Chapter I

Introduction

1.1 Introduction:

Over the last decade use of microwave energy has become the emerging field for a variety of applications. Different aspects of microwaves have been applied in various branches of science and technology like industrial measurements [1], medical and domestic applications [2], agricultural [3,4] etc. Measurement of dielectric properties of moist biogranular materials such as cereal grains and oil seed is essential for understanding their electrical behaviour and the development of indirect nondestructive methods for determining their physical characteristics including moisture content and bulk density [5].

At microwave frequencies, dielectric properties of bio vegetation are primarily a function of frequency, water saturation, porosity, texture, component geometry and electrochemical interactions. Microwave techniques and instrumentation can be utilized in agriculture to improve the efficiency of the crop production, handling and processing, and improve the quality of the products. Dielectric properties of agricultural materials and products are useful in increasing numbers of applications, as new technology is adopted for use in agriculture and related industries. Measuring and controlling moisture content is an important aspect in the harvesting, storage marketing and processing of oil seeds. The use of dielectric properties for measuring the moisture content of products such as cereal grains has produced a variety of methods using the RF range of electromagnetic radiation.

Recent airborne and deep space applications require the broadband and real time dielectric characterization of the materials at microwave frequencies apart from need of planarization. Microstrip components are very suitable to overcome the planarization aspects of various microwave devices.

Agricultural biomaterials are materials consisting of organic and inorganic inclusions along with water. Oil seeds form a major constituent of the agricultural and food sector. The agricultural application of microwaves,
particularly crop growth studies like classification, mapping of various crops, monitoring of crop growth parameters, leaf area index, moisture content, dry matter production are very important aspect.

The present work deals with use of the Ag thick film microstripline circuits in the Ku band (13-18 GHz) for dielectric characterization of moisture laden oil seeds. This frequency band has been chosen because there are no reports on use of microstrip components to study oilseeds in this frequency range. In general, this frequency range is studied very less. Microwave characterization and moisture content prediction at room temperature using Ag thick film non resonant passive component has been done of three major seeds (Soybean, Sunflower and Groundnut) having significant structural and compositional differences. The moisture levels studied are those of interest to the grain industry.

Thick film technology is a very cost effective technology [6,7] for fabrication of miniaturized microwave components. The overlay technique has been used to study the microwave characteristics of moisture laden oil seeds. The change in the response of the Ag thick film microstrip passive components has been used to characterize the moisture dependant microwave properties of biomaterials (oil seeds). The thick film circuits fabricated and investigated were Ag thick film microstripline of width 25 mil and Ag thick film equilateral triangular microstrip patch antenna. All the designing, stencil preparation, fabrication was done in the Thick and Thin Film Device Lab itself. The samples of three different seeds (Soybean, Sunflower and Groundnut) of Krishna Valley (grown at Sangli district in Maharashtra state of India in June-Sept 2006) were used as overlay.

Apart from the overlay technique, waveguide reflectometer and VSWR slotted line method was also used to characterize these moisture laden seeds at various moisture levels. The calibration curve of the actual moisture content versus measured moisture content from microwave measurement has been plotted for the various seeds using the different measurement techniques.
1.2 Microwaves: Properties and applications:

Microwaves play an ever increasing role in modern life. The electromagnetic spectrum from 300MHz to 300GHz is called the microwave spectrum. This corresponds to the range of wavelength from 30 cm to 0.3 mm in the free space. Microwaves occupy a region in the EM spectrum that is bounded by radio waves on the side of longer wavelength and infrared waves on the side of shorter wavelengths.

In this range many interesting properties of microwave radiations are offered for a variety of applications [8],

1. Microwave radiation does not alter nor contaminate the material under test, enabling fast, nondestructive and continuous monitoring.
2. Microwave can propagate through the free space allowing remote sensing to be accomplished, where lower frequency signals gets reflected from the ionosphere.
3. Relatively insensitive to environmental conditions like dust in contrast to IR radiation.
4. Most of solid dielectric materials and atmosphere is transparent to microwaves.
5. Large penetration depth compared to the IR radiation.
6. Microwaves are absorbed by the water and thus can be used to dry, heat and cook food. Their absorption can be used to determine the amount of moisture content within the materials.
7. Microwaves propagate along straight line.

Applications for which microwave provides moisture measurement are,

1. Pulp and paper manufacturing industries,
2. Packaging laminates
3. Sand and ceramics
4. Animal feed, powders and palletized product
5. Grain drying and agricultural harvesting and moisture measurement in grain, leafy vegetation like tea, alfalfa etc.
6. Frozen food, fruit, milk processing and milk water measurement
7. Moisture in soil and wood measurement

8. Oil moisture measurement to detect good and bad oil.

The dimensions of microwave components are nearly equal to wavelength and therefore new techniques and instruments are needed for dielectric characterization of the materials in the microwave region of electromagnetic spectrum.

For applications, the transmission structures for the microwave frequency range includes wave-guides, open wire TEM lines, co-axial lines and planer transmission structure which are the miniaturized version of transmission structure.

1.3 Techniques for microwave measurements:

Various techniques can be used for measurement of microwave properties of materials in the time and frequency domain. The correct choice of microwave measurement techniques have to be made and various restrictions exist in these techniques. The ideal method includes the following merits,

1. It has good measurement accuracy for the properties.
2. The measurement procedures are simple and easy.
3. The required sample dimension is small.
4. The measurement frequency range is as wide possible and it has the swept frequency capability.

At microwave frequencies, different measurement techniques can be used [9,10].

- Transmission / reflection techniques (Waveguide, coaxial)
- Impedance bridge methods
- Cavity resonator methods
- Slotted line method
- Free space methods
- Overlay technique
All these methods fall into two categories: either destructive methods in which sample preparation is needed for accurate evaluation or non destructive methods which require very little or no sample preparation. In all these methods the measured quantity enables the computation of its permittivity.

1.3.1 Transmission / reflection techniques [11-14]:

Transmission lines (waveguide) consisting of region between two concentric cylinder (co-axial line) or the interior of hollow pipe (Rectangular or circular waveguide) are used for measurements. Microwave signals are propagated through these lines as electromagnetic waves and scattered from the associated junction of these lines to travel in well-defined direction or ports. Conventional open wire lines are unsuitable for microwave transmission because of the high radiation losses that occur when the wavelength becomes smaller than the physical lengths of these lines at high frequencies. The structures [8] considered are:

1. Multi-conductor lines – coaxial lines, strip lines, microstriplines, slot lines and coplanar lines.

2. Single conductor lines – rectangular waveguides, circular waveguides and ridge waveguides.

3. Open boundary structures such as dielectric rods.

In the first category of lines, the mode of transmission is a TEM or quasi-TEM wave. In the second category of lines, the modes are either TE or TM waves or both. The third category of lines support, in general, a combination of TE and TM waves called hybrid HE modes, except possible axis-symmetric modes which are either purely TE or TM waves.

The relation for input impedance, reflection coefficient, transmission coefficient, characteristic impedance, all are applicable to microwave guides operating in a single mode.

1.3.2 Impedance bridge methods

Microwave analogs of radio-frequency impedance bridges have been
constructed. These microwave bridges have the advantages that null detection is used and uncertainties in the detector response law are unimportant. However, impedance standards at microwave frequencies are difficult to construct. Variable attenuators which serve as standards of resistance, have the undesirable property of shifting the phase along the power level. Thus reference elements of the bridge must be calibrated by another method before one can take advantage of the speed of measurement possible with the bridge methods.

Most bridges are usually of symmetrical structures, to simplify the balance equations. Much of the simplicity introduced by symmetry is lost in practice because of fringing fields and associated frequency dependent reactance between the arms of the bridge for comparing transmitted waves. At the single frequency, the symmetry may be restored by means of appropriate matching devices. Waveguide along of r-f impedance bridge was described by [10] who constructed a waveguide bridge for measurement of dielectric loss of dilute solution.

1.3.3 Cavity resonance technique [15-17]:

If a section of waveguide is closed at its two ends by metal plates; it is called as cavity resonator. A small sample in the resonant cavity will cause a shift of the resonant frequency and change in the quality factor of the cavity. The permittivity of the specimen can be calculated from these changes. The cavity can be either rectangular or cylindrical as shown in figure 1.1 and figure 1.2.

Another configuration of perturbation technique is the re-entrant cavity [18,19] as shown in figure 1.3.

1.3.4 Dielectric resonance techniques [20]:

In this technique, the dielectric sample is placed between metal shields and the transmission ($S_{21}$) and/or reflection ($S_{11}$) measured. One very popular dielectric resonance method is the Hakki-Coleman resonance method [21].
1.3.5 Slotted line and Double minima method [22]:

- **Slotted line**: The electromagnetic field at any point of a transmission line may be considered as the sum of two traveling waves. The wave from the generator incident on the load is reflected towards the generator due to mismatch, and so maximum power will not occur. The reflected wave will combine with the forward wave to give a standing wave pattern. The maximum field strength is found where the two waves add in phase and the minimum occurs where the
two add in opposite phase, the distance between two successive minima (or maxima) is half the wavelength in a transmission line. From the knowledge of the standing wave pattern, the wavelength of the exciting wave can be measured by obtaining the distance between the successive minima, preferably with a loss less load to obtain the greatest resolution.

The ratio of the amplitude of the maximum to the minimum field strength of the wave is called the standing wave ratio, or more generally, voltage standing wave ratio (VSWR), and is given by

\[
\text{VSWR} = \frac{E_{\text{max}}}{E_{\text{min}}} = \frac{(E_1 + E_2)}{(E_1 - E_2)} = \frac{1 + E_2}{1 - E_2}\]

Where \(E_1\) and \(E_2\) are respectively the amplitude of the incident and reflected electric field strength.

Further, the ratio of the reflected to the incident electric field intensities is defined as reflection coefficient, i.e., \(\Gamma_r = \frac{E_2}{E_1}\). Hence,

\[
\text{VSWR} = S = \frac{1 + \Gamma_r}{1 - \Gamma_r}
\]
When $E_{\text{min}} = E_{\text{max}}$ i.e. there is no reflection [$E_2 = 0$], the resulting VSWR = 1, this is the requirement of a matched circuit. In other words, under given ideal conditions, the VSWR of a matched load is 1.00. Under extreme mismatch conditions, $E_{\text{min}} = 0$, i.e. total reflection, [$E_2 = E_1$], the resulting VSWR = $\infty$. However, in most of the cases, VSWR > 1. It is not rare to obtain VSWR as low as 1.02 in well designed set up having matched components.

Thus the degree of mismatch between the load and the transmission line can be calculated from the measurement of VSWR. Complex load will also shift the phase and hence minimum position, so load impedance can be calculated from the VSWR and the position of the minimum point on the line with respect to the load.

The standing wave patterns are usually studied by means of the traveling detector, which probes the electric field intensity along the axis of propagation. The basic set up for coaxial line for such studies may be as shown in the block diagram of fig. 1.4.

- **Double minima method:**

To Find VSWR the double minima method is used. The direct method is not useful to measure the high VSWR because for measuring $E_{\text{min}}$, a greater probe penetration is required which produces distortion in VSWR pattern and the voltage range becomes too broad to permit operation of crystal rectifier entirely in Square law region.

Accurate measurement of VSWR greater than 1 can be made using the double minima method. The variation of square law detector output, i.e. amplitude square of the standing wave along the slotted section, is shown in fig.1.5.

The equation for $E_x^2$, the square of the amplitude of standing wave, at any point $x$ which lies between the minimum and maximum amplitude is given by

$$E_x^2 = E_{\text{max}}^2 \sin^2 \theta + E_{\text{min}}^2 \cos^2 \theta$$

Where $\theta = \pi \Delta x / \lambda$; $\Delta x$ being the distance from the minimum to some point $x$ (Fig.1.6). Thus,
Fig. 1.4 Coaxial line set up for the study of standing waves.

Figure 1.5 Typical standing wave pattern for the measurement of high VSWR

Fig. 1.6 Double minima method for high VSWR measurement (Illustrated).
\[ \text{VSWR} = \frac{E_{\text{max}}}{E_{\text{min}}} = \left\{ \frac{(E_x/E_{\text{min}})^2 - \cos^2 \theta}{\sin \theta} \right\}^{1/2} \]

Thus, by determining the value of field at any point \(x\) rather than at maximum, one can compute VSWR. However, if we consider \(\Delta x\) to correspond to twice the distance between minimum power points, calculations are simplified. Thus, \( (E_x/E_{\text{min}})^2 = 2 \)

Then

\[ \text{VSWR} = \frac{E_{\text{max}}}{E_{\text{min}}} = \sqrt{\frac{2 - \cos^2 \theta}{\sin \theta}} = \sqrt{\frac{2 + \sin^2 \theta}{\sin \theta}} \]

For high VSWR, \( \theta = \pi \Delta x / \lambda \) is very small.

So

\[ \sin \theta \rightarrow 0 \]
\[ \cos \theta \rightarrow 1 \]

\[ \text{VSWR} = \frac{E_{\text{max}}}{E_{\text{min}}} = 1 / \theta = \frac{\lambda}{\pi \Delta x} \]

### 1.3.6 Free space methods [23-25]:

The methods depending on transmission in waveguide could be readily extended to free space. The process being regarded as an extension, despite the slight simplification of the theory, because of non-standard experimental technique requirements. This extension has advantages as well as disadvantages, not offered by waveguide method.

In all the methods requiring the use of a guide, the preparation of the sample usually takes much more time than the actual process of measurement. But the free space method will eliminate the primary step of preparing the sample before measurements, as most of the industrial materials are normally supplied in sheet form. It results in saving the time for measurements. Besides this increase in speed and simplicity, the experimental procedure has the additional advantage that there are no errors from clearance in cell. Error analogous to clearance i.e. due to diffraction may be eliminated altogether if the sample is sufficiently large. Nelson et al [26] have been used free space technique for the measurement of complex permittivity of wheat over a very broad frequency range.
1.3.7 Overlay technique [27-29]:

A new technique, which employs a microwave microstripline component in the multilayer (overlay) configuration as a nondestructive and miniaturized novel sensor, has been recently developed. The loading and unloading of the samples is facilitated by its open structure.

If the microstrip component is a resonating circuit, a shift in the resonance frequency is observed which can be translated into the overlay material property. If it is a non resonating circuit, the changes in reflectance and transmittance can be converted into the material properties. The various types of microstrip components that can be used for overlay technique are simple microstripline, λ/2 rejection filter, ring resonator, straight resonator and patch antenna. All these components being in the planar form, miniaturize sensors can be fabricated.

1.4 Microwave Methods for Agricultural biogranular material:

Farming is one of the oldest and most important human activities and food production is essential to man’s survival. Intrinsic nature of the complex permittivity makes it the best entity to describe the wave–material interaction. The moist granular substances are mixtures of components exhibiting different dielectric behaviors, it is rather difficult to use the measured complex permittivity directly for moisture content. Characterization of dependence of dielectric properties of the seed to its moisture content is an important step in this endeavor. The seed show wide range of variation in shape, size, internal characteristics depending on the chemical composition of seeds that are different due to nutrition, stress, water, soil etc. Many food and agricultural products are granular. Therefore, in addition to moisture content, bulk density and other factors such as geometry and dimensions of the granular particles affect the complex permittivity.

Though a number of reports [30-33] are available for use of microwaves for biovegetation and biomaterials, only those work dealing with seeds have been referenced in this thesis. Dielectric properties of materials can be useful
for nondestructive, continuous determination of important characteristics such as the moisture content and bulk density of agricultural and food products. For food and agricultural products, the permittivity is the effective permittivity of a mixture of components with different dielectric behaviours. Microwaves can be used in the active and passive mode.

Microwave energy can be used in two different ways for moisture content determination in agricultural products. The correlation of dielectric properties with moisture content permits the sensing of moisture through electronic measurement of related quantities and other is the rapid drying of samples in microwave ovens to low levels and determining moisture content by sample weight loss and calibrating to a standard air or vacuum oven method [34]. The standard oven method requires high temperature and long time and also they are not useful for on-line process.

Of all the microwave applications to agricultural biomaterial moisture content determination has the longest and most successful history and the most promising perspectives.

1.4.1 Methods for Dielectric properties and Moisture content measurement of the seeds:

There are different methods for moisture content determination,
Water is an important component of many natural and man made material. Understanding water behaviour is fundamental in improving and preserving the quality of many of the materials.

Moisture content of cereal grains and seeds is one of the most important characteristics determining quality, proper time for harvest, the potential for safe storage and selling price because the dry matter of grain has more value than the water it contains and because costs of drying must be taken into account when grain must be dried to safe storage levels. Rapid determination of moisture content in the natural and manufactured products is important for process and quality control.

In this work microwaves have been used to characterize the seeds, the various dielectric properties and moisture content determination techniques have been divided into non radio frequency technique and radio frequency technique.

1.4.1.1 Non radio frequency techniques:

This article reviews the various non radio frequency techniques for permittivity and moisture content measurement in brief.

Air-oven method is the most common reference method for grain moisture determination. Other methods such as phosphorous pentoxide (P_2O_5) method [35] and Karl-Fischer titration method [36] are also used. These methods provide accurate results for moisture content but are very time consuming and require various chemicals. Electronic moisture meters have been used to determine moisture content in sunflower [37]. Low resolution NMR has also been used to simultaneously determine water and oil content in seeds [38]. Near IR transmittance and reflectance technology has also been used for the analysis of grains and seeds [39- 43].

1.4.1.2 RF/ Microwave techniques:

Compared to direct techniques indirect techniques are more suited for rapid determination of moisture content for automated industrial process and
the growing need for real time decision making. Radiation based methods offer on-line processing.

Nuclear- type sensors are expensive and present potential hazards. IR sensors are limited to sensing surface moisture content. The RF and microwave techniques have emerged as an alternative for rapid and non destructive sensing of moisture.

Various attempts have been made to correlate the electrical properties of agricultural products with moisture content. At microwave frequencies, the electromagnetic energy is mainly absorbed by water. This selective absorption is mainly due to the polar character of water molecules and their high permittivity compared with that of dry matter.

The moisture content of material ‘M’ is expressed as [44]

\[ M (\%) = \frac{M_w}{M_w + M_d} \times 100 \]

Where

- \( M_w \) = mass of water
- \( M_d \) = mass of dry material

Interaction of EM waves with wet material takes place at sensor. Many radiating elements in waveguide, co-axial line or strip line configurations can be used as sensors. Some examples of sensors are shown in fig. 1.7.

Dielectric based sensors are divided according to their operating frequency at two classes: MHz frequency and GHz frequency. The sensors are divided into resonant and periodic groups, into open and close structures and into reflection and transmission types. Output of sensor is function of material characteristics. The parameters like absorption in material, attenuation, phase shift, reflection coefficient, and resonant frequency can be used to measure the moisture content. All these parameters can be measured on the basis of deflection, null method including substitution, comparison and compensation methods.

The group of S. O. Nelson, S. Trabelsi, and A. W. Kraszewski have used waveguide technique, free space method and resonator method for studying a
variety of biomaterials. A large number of publications exists of this group. Only those published in the last 15 years have been included in this thesis.

Determining moisture content in granular and particulate materials from measurement of microwave parameters at a single frequency or several frequencies requires compensation for, or elimination of temperature and bulk density effects [45]. Jacobsen et al [46] have proposed the first density-independent calibration function for moisture content determination from permittivity data. It was later tested by others [47-51] for a large variety of granular and particulate materials.

Success of microwave sensing is related to the polar nature of the water molecule and its high dielectric constant (~80). Both $\varepsilon'$ and $\varepsilon''$ are intrinsic properties of the material and both depend on frequency, temperature, moisture content and for granular material, they also depend on bulk density [45,52]. At higher frequencies, bulk density effect eliminates and compensates for temperature effects [53]. Bulk density measurement is more troublesome and several density independent algorithms [54-65] for moisture sensing in granular and particulate materials have been proposed. For prediction of moisture content density independent calibration functions were used in shelled peanuts from free space measurements [66]. A unified method for moisture sensing in cereal grain and oil seed has been reported by S. Trabelsi et al [67].

Funk et al [68] has reported the RF impedance material analyser and various grains models were developed to predict unifying calibration parameters and temperature correction coefficient from the chemical and physical properties of grains.

The various microwave techniques in agriculture has been explained by Kraszewski et al [8]. RF impedance method have proved to be nondestructive method for moisture content determination for in-shell peanuts [69] by Kandala et al. Nonequilibrated water has been used for sensing moisture content on microwave dielectric properties of wheat and related errors[70]. Microwave permittivity measurements on unshelled pod useful in non-destructive and simultaneous determination of moisture have been reported by Trabelsi et al.
A non destructive multiparameters microwave sensor for simultaneous determination of bulk density and moisture content in most granular materials has been reported by Trabelsi et al [74,75].

On microwave aquametry, a review articles has been published by Kraszewski [76]. Various calibration techniques for microwave moisture sensors have been proposed [52,77]. Dielectric property measurement of most granular biomaterials have been studied by Nelson et al [5,78]. Frequency and temperature dependent permittivity have also been investigated [79-81]. The permittivity of grains have been used to sense moisture content by the same group [82-84]. Apart from this group, other workers have also studied the moisture dependent microwave characteristics of granular agricultural materials [85-88]. The interaction between electromagnetic field and the grain, air and moisture mixture is of complex nature. Trab elsi et al have reported both density and material independent permittivity based calibration function was used for prediction of moisture content in cereal grain and oil seed [89,90].

The artificial neural network (ANN) and advanced statistical techniques has been studied for determining moisture content [91-93]. Using radio frequency (RF) dielectric properties and sensing moisture content have been done by Nelson et al [94]. Data on dielectric properties of grains and sensing moisture content as a function of frequency, bulk density and temperature have become available [95-99]. Sachs et al [100] has proposed integrated network analyser module for sensing moisture content.

Broad frequency range RF and microwave dielectric property measurements are of great interest for many dielectric heating and nondestructive property sensing applications. The open-ended co-axial line sensor has been considered for measurements and developed commercially for use with network analyzers on a wide range of materials [101,102]. A novel technique for the moisture determination in single peanut pods by complex RF impedance measurement have been reported [103].

The frequency, moisture and temperature dependent dielectric properties using an open ended co-axial probe with an impedance analyser in the 10-
1800MHz range has been reported [104]. Dielectric properties of seeds using parallel plate capacitor in the 50 KHz-100MHz have been investigated by Sacilik et al [105,106].

Fig. 1.7 Classification of Microwave sensors being used for moisture content meter

Gillay et al has reviewed prediction equation relative to dielectric constant to moisture content developed for a range of RF higher measurement frequency yielded smaller predicted moisture difference than lower frequency studies on corn [107]. The use of overlay technique to study biomaterials has been done by group at Pune University and in our lab. The Pune University group mainly used the thin film microstrip ring resonator in the X-band, whereas in this work the thick film resonant and non resonant microstrip components have been used. There are very few works available for the seeds using microstripline components. The permittivity aspects of these biomaterials are also not explained in detail using this technique.

The use of thin film microstrip ring resonator as a moisture sensor in wheat grains [108], chickpea [109], mug grains [110], feasibility study of a novel moisture sensor microstripline component also [111,112] have been reported. All this work is in the X band (8-12GHz). The size dependent leaf overlay effects on thin film microstrip patch antenna has also been studied [113]. A comparative study of thin film and thick film component response to
leaf overlay has been reported by Ghorpade et al [114-116]. The moisture dependent leaf overlay effects on thick film microstrip components has also been studied by Kamble et al [117-118]. The limited studies on moisture laden seeds using microstrip components indicates the scope for further studies of these components to be used for biogranular material characterization.

1.5 Microwave Transmitting Structures: A Brief Introduction

The waveguide components, a simple Ag thick film microstripline and thick film equilateral triangular microstrip patch antenna were used, therefore only those structures are elaborated in the proceeding articles.

1.5.1 Transmission and reflection parameters:

At microwave frequency, the operational wavelength becomes comparable to the dimensions of the device geometries. It is more appropriate to consider the response of a component in the path of microwaves as it propagates through a transmission medium. The components at microwave frequency can be characterized by various transmission and reflection parameters.

The important transmission/reflection parameters include-

1. Transmission coefficient
2. Reflection coefficients
3. Insertion loss / Gain
4. Standing wave ratio

In this work rectangular waveguide with sample kept transverse to the direction of propagation has been used. The theory of propagation of microwave through rectangular waveguide in given in brief in the next article.

1.5.1.1 Rectangular waveguide:

Microwave signals were traditionally transmitted through hollow metal pipes called waveguides. Wave bounce back and forth from the conductor walls, producing signal propagation when frequencies exceed a lower bound
called cut off frequency. Metal waveguides are inherently rigid that is the realization of waveguide circuits leads to some interesting topological problems. The circuit themselves tend to be rather heavy and bulky. Waveguides are therefore used mainly where high power operation is required and at the upper limit of microwave range (millimeter range).

Rectangular waveguides were one of the earliest types of transmission lines used to transport microwave signals and are still used today for many applications. A large variety of components such as couplers, detectors, isolators, attenuators and slotted lines are commonly available for various standard waveguide bands from 1 GHz to over 220 GHz.

(The hollow rectangular waveguide can propagate TM and TE modes, but not TEM waves, since only one conductor is present.)

A rectangular (cross section) waveguide is most commonly used [119] and is shown in fig 1.8

![Fig. 1.8 Rectangular waveguide Geometry](image)

The fields inside the guide can be obtained from the solution to the wave equations which satisfy the boundary condition for a rectangular waveguide of infinite length. Two basic sets of solution exist, each of defining a set of modes. Thus there are two modes of propagation possible inside a hollow metal waveguide.

TE waves in which the E-field is wholly transverse and TM waves in which H-field is wholly transverse. No TEM waves are possible in this case.
since it require an axial conductor for axial current flow or an axial displacement current (axial E field) to support a transverse magnetic field.

For an infinite guide and propagation along the Z-direction ($e^{-jK_z z}$) the wave equations are:

\[
\nabla^2 E + K^2 E = 0 \quad 1.2 \\
\nabla^2 H + K^2 H = 0 \quad 1.3 
\]

This can be written as,

**TM:** \[
\nabla^2 E_z + K_c^2 E_z = 0 \quad 1.4 
\]

**TE:** \[
\nabla^2 H_z + K_c^2 H_z = 0 \quad 1.5 
\]

Where

\[
\nabla_i = \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} \quad \text{and} \\
K_c = \sqrt{K^2 - K_z^2} = \sqrt{K_x^2 - K_y^2} = \frac{2\pi}{\lambda c} \quad 1.6 
\]

and is called the cut off wave number. The propagation constant in the guide is, therefore,

\[
K_z = \pm j\sqrt{w^2 \mu \epsilon - K_c^2} \quad 1.7 
\]

There are three special cases apparent

1. For no propagation (evanescent mode) to take place $K_z = 0$ or $K_c = w_c \sqrt{\mu \epsilon}$ Here $w_c$ is the cut off frequency in radian/sec.

2. For propagation of real power, $K_z > 0$ or $w \sqrt{\mu \epsilon} > K_c$

3. If $w_c \sqrt{\mu \epsilon} < K_c$, $K_z$ is imaginary ($j\infty$), so that the wave will be attenuated in accordance with $e^{-\infty z}$.

### 1.5.1.2 TE waves solution:

For TE waves propagating in $+Z$ direction $E_z = 0$, and solution may be derived from the $H_z$ component of equation (1.5)

\[
H_z = \cos \left( \frac{m\pi x}{a} \right) \cos \left( \frac{n\pi y}{b} \right) \quad 1.8 
\]
This satisfies the boundary condition of the walls. Harmonic solution in the x and y directions show that the waves produce a standing wave between the longitudinal walls, giving the values \( K_x = \frac{m\pi}{a} \) and \( K_y = \frac{n\pi}{b} e^{-\mu z} \) factor is omitted but this is present in all field components. Here \( m = 0, 1, 2 \ldots \) denotes the number of half waves in the x direction and \( n = 0, 1, 2 \ldots \) those in the y direction, when \( m \neq n = 0 \). To each set of integer’s \( m \) and \( n \) a solution exists, and these modes are designed as the TE\(_{mn}\) modes. The other field components are obtained form the Maxwell’s equation as follows.

\[
H_y = \frac{jK_z}{K_c} \frac{n\pi}{b} \cos \left( \frac{m\pi x}{a} \right) \sin \left( \frac{n\pi y}{b} \right) \tag{1.9}
\]

\[
H_x = \frac{jK_z}{K_c} \frac{m\pi}{a} \sin \left( \frac{m\pi x}{a} \right) \cos \left( \frac{n\pi y}{b} \right) \tag{1.10}
\]

\[
E_z = 0 \tag{1.11}
\]

\[
E_y = \frac{\omega \mu}{K_z} H_x \tag{1.12}
\]

\[
E_x = \frac{\omega \mu}{K_z} H_y \tag{1.13}
\]

The cut-off frequency and cut-off wavelength as defined in equation for the TE\(_{mn}\) modes are,

\[
F_c = \frac{1}{2\sqrt{\mu \varepsilon}} \sqrt{\left( \frac{m}{a} \right)^2 + \left( \frac{n}{b} \right)^2} \tag{1.14}
\]

\[
\lambda_c = \frac{2}{\sqrt{\left( \frac{m}{a} \right)^2 - \left( \frac{n}{b} \right)^2}} \tag{1.15}
\]

The propagation constant is expressed by,

\[
K_z = \beta = \omega \sqrt{\mu \varepsilon} \sqrt{1 - \left( \frac{\omega_c}{\omega} \right)^2} \tag{1.16}
\]

The phase velocity of the waves

\[
0 < \beta < \omega \sqrt{\mu \varepsilon} \sqrt{1 - \left( \frac{\omega_c}{\omega} \right)^2}
\]
\[ u_p = \frac{\omega}{\beta} = \frac{u}{\sqrt{1 - \left(\frac{f_c}{f}\right)^2}} \]  

1.17

Where \( u = \frac{1}{\sqrt{\mu \varepsilon}} \), the phase velocity in an unbounded dielectric (\( \mu, \varepsilon \)).

The velocity of energy propagation called the group velocity is given by,

\[ V_g = \frac{u^2}{u_p} \]  

1.18

The propagation wavelength inside the guide is called the guide wavelength and is given by,

\[ \lambda_g = \frac{1}{\sqrt{1 - \left(\frac{\lambda}{\lambda_c}\right)^2}} \]  

1.19

Where, \( \lambda = \frac{\mu}{f} \), the wavelength in an unbounded dielectric.

The characteristic wave impedance in the guide can be derived as, for TE_{mn} mode.

\[ Z_w = \frac{E_x}{H_y} = -\frac{E_y}{H_x} = \frac{a \mu}{\beta} = \frac{n k}{\beta} \left( \frac{\eta \lambda_g}{\lambda_0} \right) = \frac{\lambda}{\lambda_0} \]  

1.20

where \( \eta = \sqrt{\frac{\mu}{\varepsilon}} \) is the intrinsic impedance in an unbounded dielectric.

### 1.5.1.3 TM Waves Solution

The TM waves propagating in +z direction are characterized by \( H_z = 0 \).

A Field solution to equation (1.4) gives

\[ E_z = \sin \left( \frac{m \pi x}{a} \right) \sin \left( \frac{n \pi y}{b} \right) \]  

1.21

\[ E_y = \frac{j n k}{bk \varepsilon} \sin \left( \frac{m \pi x}{a} \right) \cos \left( \frac{n \pi y}{b} \right) \]  

1.22

\[ E_x = \frac{j m k}{ak \varepsilon} \cos \left( \frac{m \pi x}{a} \right) \sin \left( \frac{n \pi y}{b} \right) \]  

1.23
\[ H_z = 0 \]  
\[ H_y = \frac{\omega \varepsilon}{K_z} E_x \]  
\[ H_x = -\frac{\omega \varepsilon}{K_z} E_y \]  

Where \( m = 1, 2, 3, \ldots \) and \( n = 1, 2, 3, \ldots \). Here \( m \) or \( n \neq 0 \), since either \( m=0 \) or \( n=0 \) leads to zero field intensities. For \( \text{TM}_{mn} \) modes, the equations for the cut off frequency \( f_c \), propagation constant \( \beta \), the phase and group velocities \( u_p \) and \( u_g \), and the guide wavelength \( \lambda_g \) are the same as those for \( \text{TE}_{mn} \) modes given in equation (1.14) to (1.19).

The characteristics wave impedance is given by,

\[ Z_w = \frac{E_x}{H_y} = -\frac{E_y}{H_x} = \frac{\beta}{\omega \varepsilon} = \frac{n \beta}{k} = \eta \sqrt{1 - \left(\frac{f_c}{f}\right)^2} = \frac{377 \, \lambda_g}{\lambda_x} \text{ohm} \]  

\( \text{TE}_{10} \) mode is a simple and useful field distribution, has minimum attenuation arising from various factors. The ribs of the waveguide operating on \( \text{TE}_{10} \) mode carry the maximum transverse current, consequently, a very good electric contact between the waveguide walls at the ribs has to be insured, if hole is fabricated from the separate strip of metal, size the \( \text{TE}_{10} \) mode is dominant in rectangular waveguide.

### 1.5.2 Transmission lines for Planar MIC:

Planar transmission lines are those in which the entire transmission line components can be fabricated in one step by thin film or photolithographic techniques. A transmission line is made up of a symmetrical pair of strip conductor or a single strip and ground plane, at opposite face to of sheet of dielectric material. These techniques are simple methods for broadband characterization. Transmission lines \([120-121]\) can be in the form of microstripline, slot line, coplanar waveguide and many other types of related geometries as shown in figure 1.9. Mostly they are made of conducting strips deposited on dielectric substrate along with other active and passive components they are called microwave integrated circuits [MIC]. Strip line and
Microstrip line were first introduced before 1950’s. Since then the use of MIC’s for non-military and military applications have increased tremendously.

MIC is miniaturized form of microwave circuit. The concept exploits small size, low weight, improved reliability and reproducibility with potential low cost with the bandwidth performance with low parasitic impedance. The main application of MIC’s is in low power system. Production of MIC’s involves number of intermediate specialized processing, design and assembly stages.

1.5.2.1 Microstrip Line:

The microstrip line is an inhomogeneous transmission line structure. It consists of strip conductor on a flat dielectric substrate, the reverse side of which is metallized to provide a ground plane. The cross sectional view of microstrip line along with electric and magnetic field lines is shown in figure 1.10. In this structure, pure TEM mode cannot exist. Due to absence of a ground plate and the dielectric substrate above the strip, the electric field lines remain partially in the air and partially in the lower dielectric substrate. This makes the mode of propagation not pure TEM but what is termed as the quasi TEM mode [114]. Due to the open structure and presence of discontinuity, the microstrip line radiates electromagnetic energy. The radiation loss is proportional to the square of frequency. The use of thin and high dielectric materials reduces the radiation loss of the open structure where the fields are mostly confined inside the dielectric. Under the quasi-static approximation, the propagation parameters of such lines can be defined by replacing the mixed dielectric by equivalent uniform dielectric medium.

Then dielectric constant of the equivalent medium is termed as the effective dielectric constant and is defined as,

$$\varepsilon_{\text{eff.}} = \frac{C}{Ca}$$ (1.28)
Fig. 1.9 The cross sections of Planer Transmission lines used in MIC's.

a) Completely non-TEM

\[ \varepsilon_0 \]

i) Image line

\[ \varepsilon \]

b) Quasi -TEM

\[ \varepsilon_0 \]

i) Microstrip

\[ \varepsilon_0 \]

ii) Coplaner waveguide(CPW)

\[ \varepsilon_0 \]

iii) Inverted microstrip

\[ \varepsilon \]

iv) Trapped inverted microstrip(TIM)

\[ \varepsilon_0 \]

v) Suspended stripline

Fig 1.10 Schematic diagrams of microstripline

i) Schematic diagram

\[ \text{Strip conductor} \]

Ground plane

ii) Field line
Where, \( C \) is capacitance per unit length of transmission line and \( C_a \) capacitance per unit length with all dielectric replaced by air.

The phase velocity \( V \), guided wave length \( \lambda_g \) and characteristic impedance \( Z \) are given by,

\[
V = \frac{Va}{(\varepsilon_{eff})^{\frac{1}{2}}} \quad \text{(1.29)}
\]

\[
\lambda_g = \frac{\lambda_a}{(\varepsilon_{eff})^{\frac{1}{2}}} \quad \text{(1.30)}
\]

\[
Z = \frac{z_{01}}{(\varepsilon_{eff})^{\frac{1}{2}} \cdot (\varepsilon_{eff})^{\frac{1}{2}}} \quad \text{(1.31)}
\]

Here \( \varepsilon_{eff} \) is effective dielectric constant of the medium and its values lies in the range \( 1 < \varepsilon_{eff} < \varepsilon_r \), \( z_{01} \) is the characteristic impedance of microstrip line with dielectric replaced by air.

### 1.5.2.2 Synthesis formulae (\( Z_0 \) and \( \varepsilon_r \) given):

For narrow strips [i.e. when \( Z_0 > \{44 - 2\varepsilon_r\} \) ohms]:

\[
\frac{W}{h} = \left( \exp H' \right) - \left( \frac{1}{4\exp H'} \right)^{-1} \quad \text{(1.32)}
\]

Where

\[
H' = \frac{Z_0 \sqrt{2(\varepsilon_r + 1)}}{119.9} + \frac{1}{2} \left( \frac{\varepsilon_r - 1}{\varepsilon_r + 1} \right) \left( \ln \frac{\pi}{2} + \frac{1}{\varepsilon_r} \ln \frac{4}{\pi} \right) \quad \text{(1.33)}
\]

We may also use, with slight but significant shift of changeover value to \( W/h < 1.3 \) (i.e. when \( Z_0 > \{63 - 2\varepsilon_r\} \) ohms):

\[
\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} \left( 1 - \frac{1}{2H'} \left( \frac{\varepsilon_r - 1}{\varepsilon_r + 1} \right) \left( \ln \frac{\pi}{2} + \frac{1}{\varepsilon_r} \ln \frac{4}{\pi} \right) \right)^{-2} \quad \text{(1.34)}
\]

Where \( H' \) is given by equation (1.33) (as a function of \( Z_0 \)) or, alternatively, as a function of \( W/h \), from equation (1.32):
Another, somewhat simpler, expression for $\varepsilon_{\text{eff}} (Z_0)$ is given by Owens is,

$$
\varepsilon_{\text{eff}} = \varepsilon_r + 1 \left\{ 1 + \frac{29.98}{Z_0} \left[ \varepsilon_r - 1 \right] \left( \frac{\ln \frac{\pi}{2} + \frac{1}{\varepsilon_r} \ln \frac{4}{\pi} \right)^2 \right\}
$$

Now consider the ranges and formulae for “Wide” strips.

For wide strips (i.e. when $Z_0 < \{44 - 2 \varepsilon_r\}$ ohms):

$$
\frac{W}{h} = \frac{2}{\pi} \left\{ \left( \varepsilon_r - 1 \right) - \ln \left( 2 \varepsilon_r - 1 \right) \right\} + \frac{\varepsilon_r - 1}{\pi \varepsilon_r} \left\{ \ln \left( \varepsilon_r - 1 \right) + 0.293 - \frac{0.517}{\varepsilon_r} \right\}
$$

Where

$$
d = \frac{59.95 \pi^2}{Z_0 \sqrt{\varepsilon_r}}
$$

With the same slight shift of changeover value as before (i.e. where $W/h > 1.3$ and $Z_0 > \{63 - 2 \varepsilon_r \}$ ohms)) Owens found that a modified form of an earlier expression due to Schneider gave very accurate results. Owens’s [122] formula is:

$$
\varepsilon_{\text{eff}} = \varepsilon_r + 1 \left( 1 + \frac{h}{W} \right)^{-0.555}
$$

Alternatively, where $Z_0$ is known at first

$$
\varepsilon_{\text{eff}} = \frac{\varepsilon_r}{0.96 + \varepsilon_r \left( 0.109 - 0.004 \varepsilon_r \right) \log \left( 10 + Z_0 \right) - 1}
$$

For microstriplines on alumina ($\varepsilon_r = 10$) this expression appears to be accurate to ± 0.2 percent over the impedance range $8 \leq Z_0 \leq 45 \ \Omega$

1.5.2.3 Analysis Formulae ($\frac{W}{h}$ and $\varepsilon_r$ given):

For ‘narrow’ strips ($\frac{W}{h} < 3.3$):
\[ Z_0 = \frac{119.9}{\sqrt{2}(\varepsilon_r + 1)} \left[ \ln \left( \frac{4h}{W} + \sqrt{16 \left( \frac{h}{W} \right)^2 + 2} \right) - \frac{1}{2} \left( \frac{\varepsilon_r - 1}{\varepsilon_r + 1} \right) \left( \ln \frac{\pi}{2} + \frac{1}{\varepsilon_r} \ln \frac{4}{\pi} \right) \right] \] ---1.41

For ‘wide’ strips \( \left( \frac{W}{h} > 3.3 \right) \):

\[ Z_0 = \frac{119.9\pi}{2\sqrt{\varepsilon_r}} \left[ \frac{W}{2h} + \frac{\ln \left( \frac{e \pi^2}{16} \right)}{\pi} + \frac{\ln \left( \frac{e \pi^2}{16} \right)}{2\pi} \left( \frac{\varepsilon_r - 1}{\varepsilon_r^2} \right) \right]^{-1} \] \[ \frac{\ln \left( \frac{\pi}{2} \right) + \ln \left( \frac{W}{2h} + 0.94 \right)}{2\varepsilon_r} \] ---- 1.42

Where, ‘e’ is the exponential base: \( e = 2.7182818 \) ----

In all cases the shape ratio \( W/h \) will be accurate to ± 1%. For narrow lines \( (W/h < 1/3) \), \( \varepsilon_{\text{eff}} \) has the error range of + 0.5 – 0.0%. When calculated using equation (1.39), \( \varepsilon_{\text{eff}} \) is accurate to ± 0.25%. The expressions for \( Z_0 \) result in inaccuracies to ± 1%.

The characteristic parameters of microstripline depend on strip width ‘W’, thickness ‘h’ and the dielectric constant \( \varepsilon_r \) of the substrate. For a given thickness and dielectric constant of the substrate, strip width controls the characteristics of the line. This feature makes microstriplines well suited for MIC’s.

The microstrip synthesis consists of finding the values of width ‘W’ and length ‘L’ corresponding to the characteristic impedance \( Z_0 \) and electrical length ‘\( \theta \)’. Initially a substrate of thickness ‘h’ and relative permittivity \( \varepsilon_r \) has to be chosen. The choice depends upon certain frequency limitations. Initially the synthesis yield \( W/h \) ratio and effective microstrip permittivity \( \varepsilon_{\text{eff}} \) from the other design parameters can be determine [122,123]. Most of the synthesis available is for substrates with \( \varepsilon_r \) less than 15. Recently Crute [124] has investigated properties of high \( \varepsilon_r \) microstrip lines.

**1.5.2.4 Effective dielectric constant of microstripline:**

Since the propagation field lines in a microstrip lie partially in air and partially inside the homogenous dielectric substrate, the propagation delay time
for a quasi-TEM mode is related to an effective dielectric constant \( \varepsilon_{\text{eff}} \) [121] given by,

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

and \( \varepsilon_{\text{eff}} \) = \[\frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + \frac{12h}{w} \right)^{-\frac{1}{2}} \] ; \( \frac{w}{h} \gg 1 \)

\[
Z_o = \frac{\sigma_o}{\sqrt{\varepsilon_{\text{eff}}}} \ln \left[ \frac{8h}{w} + \frac{w}{4h} \right] \; \text{Ohm} \; \frac{w}{h} \leq 1
\]

Where, \( \varepsilon_r \) is the relative dielectric constant of the substrate material.

**Characteristic impedance and guide wavelength:**

Characteristics impedance of microstrip lines [121] can be expressed by,

\[
Z_o = \frac{376.7}{\sqrt{\varepsilon_{\text{eff}}}} \left[ \frac{w}{h} + 1.4 + 0.667 \ln \left( \frac{w}{h} + 1.444 \right) \right] \; \text{Ohm} \; \frac{w}{h} > 1
\]

\[
Z_o = \frac{376.7}{\sqrt{\varepsilon_{\text{eff}}}} \frac{h}{w} \; \text{Ohm} \; \frac{w}{h} \gg 1
\]

The guide wavelength for the propagation of quasi-TEM mode is given by,

\[
\lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_{\text{eff}}}}
\]

### 1.6 Antenna:

All antennas operated in the microwave frequencies are called microwave antennas. Antennas are very useful components in the free transmission of microwaves. An antenna is a structure that transforms guided EM wave into free space EM waves and vice versa. An antenna is a reciprocal device that is its directional pattern as receiving antenna is identical to its directional pattern when the same is used as transmitting antenna.

**Types of Antennas:**

a) Yagi – Uda antenna
b) Reflector antenna  
c) Horn antenna  
d) Helical antenna  
e) Log periodic antenna  
f) Phased array antenna  
g) Microstrip patch antenna.

In this work the microstrip patch antenna and horn antenna has been used.

Therefore, only these two antennas are elaborated in the proceeding articles.

Antennas are characterized by following parameters:

1.6.1 Microstrip antenna:

Patch antennas first suggested by Munson [125], are essentially radiating elements in a microstrip environment and have proved to be popular in both single and arrayed configurations. Microstrip antennas have been one of the most innovative topics in antenna theory and design in recent years, and are increasingly finding application in a wide range of modern microwave system.
The simple configuration of microstrip antenna is a metallic radiator on a dielectric substrate as a support, with ground plane on the other side. The metal used is normally gold, copper or silver. Ideally the dielectric constant of the substrate should be low ($\varepsilon_r = 2.5$), so as to enhance the fringing fields, which account for radiation. The higher limit for value of $\varepsilon_r$ is 13. Various types of substrates used are quartz, glass, sapphires, alumina etc. Alumina ($\text{Al}_2\text{O}_3$) is most commonly used because of higher thermal conductivity, strength and stability and good electrical properties.

They are used in various systems such as radar, satellite, communication, missiles, space vehicles and various defense equipments because of their thin profile, low weight, planar configuration and easy integrability with other circuits.

Microstrip antennas may be of any geometrical shape and any dimensions. However all the microstrip antennas may be divided into three basic categories are:

A) Microstrip patch antennas
B) Microstrip traveling wave antennas
C) Microstrip slot antennas

1.6.2 Equilateral Triangular Microstrip Patch Antenna:

Among the various shapes of the microstrip patch antennas the rectangular, square, circular and annular ring have been studied extensively. In contrast, the triangular patch has been the subject of only a handful of investigations [126-129].

The basic triangular element can be thought of a half wavelength, open circuited patch fed by either a microstripline in the same plane, or by a coaxial probe entering under the patch in a perpendicular plane. The equilateral triangular patch has also received attention due to its interesting features [130,131]. An experimental study of this antenna was reported by Gang et al [132]. Equilateral triangular patch antennas have a smaller size compared to square or circular patch antennas for a given frequency [133-139].
The effect of inset position on the properties of EMC equilateral triangular microstrip patch antenna has been reported by Nuangnum et al [140]. Geometry of triangular patch and feedline shown in fig. 1.11 form on equilateral triangular microstrip patch antenna. This conductor patch is an one side of dielectric substrate and can be fed by an open ended microstrip line.

**Fig 1.11 Geometry of Equilateral Triangular Microstrip patch Antenna**

1.6.2.1 Analysis of an Equilateral Triangular Microstrip Patch Antenna:

A triangular patch antenna may be completely characterized in terms of its radiated field pattern, input impedance, gain, bandwidth, beam-width, efficiency, losses and Q-factor [131].

By duality principal, the TM field patterns with magnetic boundary conditions are the same as those for TE modes with electric boundary conditions. For the TM \(_{mn}\) modes, the electric and magnetic field distributions of the equilateral triangular resonator [138] are,

\[
E_z = A_{m,n,l} T(x,y)_{m,n,l} 
\]

\[
H_x = \frac{j}{\omega \mu} \frac{\partial E_z}{\partial y} \]

1.49

1.50
\[ H_y = -j \frac{\partial E}{\partial x} \quad \text{(1.51)} \]
\[ H_z = E_x = E_y = 0 \quad \text{(1.52)} \]

where,
\[
T(x, y)_{m,n,l} = \cos \left[ \left( \frac{2\Pi x}{\sqrt{3a}} + \frac{2\Pi}{3} \right) y \right] \cos \left[ \frac{2\Pi (m-n)y}{3a} \right] + \cos \left[ \left( \frac{2\Pi x}{\sqrt{3a}} + \frac{2\Pi}{3} \right) m \right] \cos \left[ \frac{2\Pi (n-l)y}{3a} \right]
+ \cos \left[ \left( \frac{2\Pi x}{\sqrt{3a}} + \frac{2\Pi}{3} \right) n \right] \cos \left[ \frac{2\Pi (l-m)y}{3a} \right] \quad \text{(1.53)}
\]

And \( A_{m,n,l} \) is an arbitrarily amplitude constant, \( a \) is the length of a side of the triangle, and \( m, n, l \) are integers which are never zero simultaneously and which satisfy the condition,
\[ m + n + l = 0 \quad \text{(1.54)} \]

The fields satisfy the wave equation
\[
\left( \frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + K_{mn}^2 \right) E_z = 0 \quad \text{(1.55)}
\]

where,
\[ K_{mn} = \frac{4\Pi}{3a} \sqrt{m^2 + mn + n^2} \quad \text{(1.56)} \]

It is observed that interchanging the three digits \( m, n, l \) leaves the wave number \( K_{mn} \) and the field patterns unchanged. The field patterns for the first two modes.

The above expressions for the fields are general. The particular case, \( m = 1 \), \( n=0 \) and \( l = -1 \) corresponds to the dominant mode, for which the field expressions are,
\[
E_z = A_{1,0,-1} \left[ 2 \cos \left( \frac{2\Pi x}{\sqrt{3a}} + \frac{2\Pi}{3} \right) \cos \frac{2\Pi y}{3a} + \cos \frac{4\Pi y}{3a} \right] \quad \text{(1.57)}
\]
\[
H_x = -j A_{1,0,-1} \xi_0 \left[ \cos \left( \frac{2\Pi x}{\sqrt{3a}} + \frac{2\Pi}{3} \right) \sin \frac{2\Pi y}{3a} + \sin \frac{4\Pi y}{3a} \right] \quad \text{(1.58)}
\]
\[
H_y = j \sqrt{3} A_{1,0,-1} \xi_0 \left[ \sin \left( \frac{2\Pi x}{\sqrt{3a}} + \frac{2\Pi}{3} \right) \cos \frac{2\Pi y}{3a} \right] \quad \text{(1.59)}
\]
\[
\xi_0 = \frac{1}{120 \Pi (mhos)}
\]
1.6.2.2 Resonant frequency:

The resonant frequency of the equilateral triangular patch [136] can be calculated by using,

\[ f_r = \frac{ck_{mn}}{2\pi \sqrt{\varepsilon_r}} = \frac{2c}{3a\sqrt{\varepsilon_r}} \left( m^2 + mn + n^2 \right)^{1/2} \]  \hspace{1cm} 1.60

where,

- \( c \) is the velocity of the light \((3 \times 10^{10} \text{ cm/s})\) in free space.
- \( a \) = length of equilateral patch
- \( \varepsilon_r \) = relative permittivity of dielectric substrate

A plot of the product \((fr \times a)\) as a function of \(\varepsilon_r\) for various values of \(m\) and \(n\). The lowest order resonance frequency is,

\[ f_r = \frac{2c}{3a\sqrt{\varepsilon_r}} \]  \hspace{1cm} 1.61

This relationship does not take into consideration end effects or effects of fringing fields. The resonant frequency may be determined with better accuracy if \(\varepsilon_r\) is replaced by effective dielectric constant which may be calculated from the average values of the effective dielectric constants for strip width’s \(W=0\) and \(w=A\).

\[ \varepsilon_e = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{4} \left( 1 + 12 \frac{h}{a} \right)^{-1/2} \]  \hspace{1cm} 1.62

Similarly the end effects are taken into account by replacing \(a\) by \(a_{eff}\) an approximate value of \(a_{eff}\) is given by,

\[ a_{eff} = a + \frac{h}{\sqrt{\varepsilon_e}} \]  \hspace{1cm} 1.63

Thus, the final expression for the lowest order resonant frequency becomes,

\[ f_r = \frac{2c}{3a_{eff} \sqrt{\varepsilon_e}} \]  \hspace{1cm} 1.64

The frequency difference between theoretical values and experimental results is approximately 3%. For high dielectric constant substrates such as alumina, \(\varepsilon_e\) should be used in place of \(\varepsilon_r\).
1.6.2.3 Radiation fields:

The triangular patch was first introduced by Helszajn and James [139] as a traveling wave antenna. Its resonant behavior was studied by Mink [141]. The equilateral triangular patch is used as resonator.

The radiation characteristics of two-element array of equilateral triangular patch antenna are theoretically studied in presence of EM waves. The radiation fields of an equilateral triangular patch antenna may be obtained by calculating the vector electric potential [129].

The far field components may be calculated from

\[ E_r = 0 \]

\[ E_\theta = -j \eta_0 \omega \left[ -F_x \sin \phi + F_y \cos \phi \right] \quad \text{--- 1.65} \]

\[ E_\phi = j \eta_0 \omega \left[ F_x \cos \theta \cos \phi + F_y \cos \theta \sin \phi \right] \quad \text{--- 1.66} \]

The expression for the total radiation field pattern \( R(\theta, \phi) \) is obtained as,

\[ R(\theta, \phi) = |E_\theta|^2 + |E_\phi|^2 \quad \text{--- 1.67} \]

Then, the radiation field patterns in the E plane \((\phi = 0)\) and the H plane \((\phi = \pi/2)\) are given as,

\[ R_E = \eta_0^2 \omega_0^2 \left| F_x \right|^2 + |F_y|^2 \cos^2 \theta \] (E-Plane) \quad \text{--- 1.68} \]

\[ R_H(\theta, \phi) = \eta_0^2 \omega_0^2 \left| F_x \right|^2 + |F_y|^2 \cos^2 \theta \] (H-Plane) \quad \text{--- 1.69} \]

The triangular patch antenna is characterized by a single parameter namely the length of patch.

1.6.3 Horn antenna:

In this work horn antenna is used as receiving antenna. A horn antenna may be regarded as a flared out (or opened-out) waveguide. The function of the horn is to produce a uniform phase front with a larger aperture than that of the waveguide, and hence greater directivity.

Several types of Horn antennas [142, 143] are illustrated in figure 1.12
Those in the left column are rectangular horns. All are energized from rectangular waveguides. Those in the right column are circular types. To minimize reflections of the guided wave, the transition region or horn between the waveguide at the throat and free space at the aperture could be given a gradual exponential taper as in figure 1.12 (a). The types in figure 1.12 (b) and (c) are sectoral horns. They are rectangular types with a flare in only one dimension. Assuming that the rectangular waveguide is energized with a TE\textsubscript{10} mode wave electric field (E in the Y-direction), the horn in figure 1.12 (b) is flared out in a plane perpendicular to E. This is the plane of the magnetic field H. Hence, this type of horn is called a sectoral horn flared in the H-plane or simply an H-plane sectoral horn. The horn in figure 1.12 (c) is flared out in the plane of electric field E, and hence, is called an E-plane sectoral horn. A rectangular horn with flare in both planes, as in figure 1.12 (d) is called a pyramidal horn. The arrows indicate the direction of the electric field E, and their length gives an approximate indication of the magnitude of the field intensity. For small flare angles, the field variation across the aperture of the rectangular horns is similar to the sinusoidal distribution of the TE\textsubscript{10} mode across the waveguide.

Neglecting edge effects the radiation pattern of a horn antenna can be determined if the aperture dimensions of aperture field distribution are known for a given aperture. The directivity is maximum for a uniform distribution variation in magnitude or phase of the field across the aperture decrease to the direction. Since the H-plane, sectoral horn has a field distribution over X-direction which types to zero at the edge of aperture.

The principle of equality of path of length (Fermat’s principle) is applicable to the horn design but with different emphasis. Instead of requiring a constant phase across the horn mouth, the requirement is relaxed to one where the phase may deviate, but by less than a specified amount, equal to the path length difference between a ray traveling along the side and along the axis of the horn.
Fig 1.12 Types of rectangular and circular horn antenna
(Arrows indicate E field direction)

From fig (1.12)

\[
\cos\left(\frac{\theta}{2}\right) = \frac{L}{L + \delta} \quad 1.70
\]

\[
\sin\left(\frac{\theta}{2}\right) = \frac{a}{2(L + \delta)} \quad 1.71
\]

\[
\tan\left(\frac{\theta}{2}\right) = \frac{a}{2L} \quad 1.72
\]

Where,
\[ \theta = \text{Flare angle (}\theta_E \text{ for E-plane,} \ \theta_H \text{ for H-plane)} \]
\[ a = \text{Aperture (}a_E \text{ for E-plane,} \ a_H \text{ for H-plane)} \]
\[ L = \text{Horn length.} \]

From the geometry, we have also that,
\[ L = \frac{a^2}{8\delta}, \ \delta \ll L \]

And
\[ \theta = 2\tan^{-1} \left( \frac{a}{2L} \right) = 2\cos^{-1} \left( \frac{L}{L + \delta} \right) \]

In the E plane of the horn, \( \delta \) is usually held to 0.25 \( \lambda \) or less. However, in the H plane, \( \delta \) can be larger, or about 0.4 \( \lambda \), since E plane goes to zero at the horn edges.

The optimum horn dimension can be related by,
\[ \delta_0 = \frac{L}{\cos(\theta/2)} - L \]

Where,
\[ L = \frac{\delta_0 \cos(\theta/2)}{1 - \cos(\theta/2)} \]

It follows that the width of the waveguide at the throat of the horn must be between \( \lambda/2 \) and \( \lambda \), or if the excitation system is symmetrical, so that even modes are not energized, the width must be between \( \lambda/2 \) and \( 3\lambda/2 \).

1.6.3.1 **Radiation pattern:**

An antenna radiation pattern is defined as a graphical representation of the radiation properties of the antenna as a function of space coordinates. The radiation pattern can be of various types:

i) Isotropic

ii) Omni directional

iii) Pencil beam

iv) Fan beam

v) Shaped beam
Fig. 1.13 (a) Pyramidal horn antenna.
Fig. 1.13 (b) Cross section with dimensions in analysis.

The diagram can be for either E plane or H plane cross section. For the E plane, the flare angle is $\theta_E$ and aperture $a_E$ (For the H plane the flare angle is $\theta_H$ and the aperture $a_H$)

Various parts of the radiation pattern are called lobes –

a) Major lobe
b) Side lobe
c) Back lobe
d) Minor lobe

A major lobe (main lobe) is defined as the radiation lobe containing the direction of maximum radiation.

A side lobe is a radiation lobe in any direction other than the main lobe.

A back lobe usually refers to a minor lobe that occupied the hemisphere in a direction opposite to that of major lobe.
Minor lobe usually represents radiation in undesired direction and they should be minimized.

1.6.3.2 Directivity and gain:

Apart from radiation pattern, directivity and gain are very important. The directivity $D$ of an antenna is given by the ratio of the maximum radiation intensity (power per unit solid angle) $U(\theta, \phi)_{\text{max}}$ to the average radiation intensity $U_{\text{av}}$ (averaged over a sphere). Or, at a certain distance from the antenna the directivity may be expressed as the ratio of the maximum to the average Poynting vector.

Thus,

$$D = \frac{U(\theta, \phi)_{\text{max}}}{U_{\text{av}}} = \frac{S(\theta, \phi)_{\text{max}}}{S_{\text{av}}} \text{ (dimensionless)}$$

Both radiation intensity and Poynting vector values should be measured in the far field of antenna.

Now the average Poynting vector over given by

$$S(\theta, \phi)_{\text{av}} = \frac{1}{4\pi} \int_{0}^{2\pi} \int_{0}^{\pi} s(\theta, \phi) d\Omega \quad \text{(Wm}^{-2}\text{)}$$

Thus the directivity,

$$D = \frac{1 \int_{0}^{2\pi} \int_{0}^{\pi} s(\theta, \phi) d\Omega}{4\pi \int_{0}^{2\pi} \int_{0}^{\pi} S(\theta, \phi)_{\text{max}} d\Omega}$$

$$D = \frac{1}{4\pi} \int_{0}^{2\pi} \int_{0}^{\pi} P_n(\theta, \phi) d\Omega$$

or

$$D = \frac{4\pi}{\Omega_n}$$

The smaller the beam solid angle, the greater the directivity.

The gain of an antenna depends on both its directivity and its efficiency. If the efficiency is not 100%, the gain is less than the directivity.
Fig. 1.14 (a) Antenna field pattern with coordinate system.

(b) Antenna power pattern in polar coordinate (linear scale)

(c) Antenna pattern in rectangular coordinate and decibel logarithmic) scale,

Patterns (b) and (c) are the same.
Thus the gain,

\[ G = kD \quad \text{(dimensionless)} \]

Where, \( k \) = efficiency factor of antenna \((0 \leq k \leq 1)\), dimensionless.

1.7 Fabrication of the Microstrip components:

There are two types of hybrid technologies involved for the fabrication of hybrid microwave circuits.

1) Thick film technology
2) Thin film technology.

H. Sobol [144] has reported that the term thin film or thick film refers to the process used and not to the thickness of the film.

Thick films are realized by the process of screen printing of an appropriate type of paste or ink on the substrate and are fired at the specified temperature. Depending upon the composition of the inks, after firing they give rise to conductors, resistors or dielectrics. Basically, there are six techniques for fabrication of microstripline circuits [145].

These are

1. Vacuum deposition.
   a. Evaporation through mask (additive process).
   b. Evaporate allover and photo etch (subtractive process).
2. Dc sputter allover and photo etch.
3. Dc sputtering followed by electroplating and photo etch.
4. Vacuum deposition followed by electroplating and photo etch.
5. Electro-less deposition followed by electroplating and photo etch.
6. Screen print and fire i.e. thick film technology.
   a. Pattern screened directly.
   b. Pattern photo etched from screened and fired inks.

In the present work only screen printing (thick film technology) was used for microstrip circuit fabrication, so only this technique is discussed in brief in the following article.
1.7.1 Thick film Deposition Technique:

The features and potential of thick film technology has made it one of the leading solid-state technologies. This section reviews the thick film process used for fabricating our thick films. The history of thick film technology dates back to 1950’s. It soon became clear that the result of fabricating components with different technologies could open up a whole new field in electronics. Now a days thick film technology is widely, used technology for gas sensors fabrication.

A thick film circuit is basically considered to consist of a layer of special inks or pastes deposited on to an insulating substrate. By adding integrated circuits, and sometimes film made by other techniques, a hybrid circuit is produced. One of the key factor that distinguishes a thick film, is the method of film deposition screen-printing, which is possibly one of the oldest form of graphic art reproduction.

The term thick film has gained acceptance as the preferred generic description for that field of microelectronics in which specially formulated pastes are applied and fired onto a ceramic substrate in a definite pattern and sequence to produce a set of individual components, such as resistors and capacitors, or a completely functional circuit [146-148].

The deposition process adopted for thick film production is similar to that used in traditional silkscreen printing of pencils, mugs, visiting cards etc. The degree of sophistication of the printing, and the nature of the screen materials, are somewhat different for the microelectronics circuit production requirements. A typical thick film screen is made from a finely woven mesh of stainless steel, polyester or nylon. This is mounted under tension on a metal or wooden frame and coated with an ultraviolet sensitive emulsion upon which the circuit pattern could be formed photographically. The finished screen has open-mesh areas through which the desired pattern can be printed. The screen is held in a position within screen-printing machine at a distance around 0.5mm from the surface of the substrate. The thick film paste, which is resistive, conductive or dielectric in nature, is placed on to the top surface of the stencil and a squeegee
traverses the screen under pressure. This action brings the screen into contact with the substrate and forces the paste through the open areas. The required pattern is thus deposited onto the substrate. The screen printing process is shown in figure 1.15. The next stage of the process is to dry the printed film. All the pastes contain various organic solvents, which are needed in order to produce the correct viscosity for screen-printing. These can be removed by drying the film in an infrared belt drier or a conventional box furnace at a temperature 150°C.

**Fig.1.15 Screen printing process**
After drying, the films retain a rigid pattern on the substrate and are relatively immune to smudging. In certain circumstances, another layer can be printed directly onto a dry film, but more often, the next step in the process is to fire the film at high temperatures. In addition to the organic vehicle, thick film pastes also contain finely divided particles of active material having an average diameter around five microns, and a frit (powdered glass) or some metal oxide having low melting point which act as binder. During the firing (annealing) process the glass melts, the fine powder sinters and the overall film becomes a solid composite material.

The function of the glass (metal oxide) is to bond the film to the substrate and also to bind the active particles together. Firing is performed in a belt furnace at peak temperature around 850°C (or in a Muffle furnace with controlled heating rate). After firing the film is firmly attached to the substrate. The thick film pastes used in this work are fritless (does not containing glass).

1.8 Microstripline components with overlay: A brief survey

In this work moisture dependent different types of oil seeds are used as overlay over simple microstripline and microstrip patch antenna. When any material is kept in contact with microstrip components over it is termed as overlay or superstrate. When the overlay is used over a open microstripline, it is called Multilayer Microstrip [149].

The overlay effects due to fringing field of microstrips are mainly governed by the dielectric constant ($\varepsilon_r$) [150, 151] of the overlay e.g. (a) with increase in the ($\varepsilon_r$) of the overlay, the fringing field lines get concentrated, thus increasing the fringing field capacitance, which finally results decrease in resonance frequency $f_r$ of the resonator [152]. (b) When metallic overlay is used with dielectric spacing (d), as a result of induced charges in the metal, the fringing fields, capacitance of the microstrip decreases which results in increase in $f_r$. (c) At the same time it is observed that very large and/or very small ($\varepsilon_r$) of the overlay are not much effective in changing the $f_r$ [153]. Mesa [154] has studied the leakage from microstripline due to overlay. A leaky
dominant mode can exist on microstripline without top cover, but only for relatively high frequency or thick substrates. However when top cover is introduced over the substrate, the frequency at which leakage begins is dramatically lowered.

The conductive patches of microstrip antennas are often covered by a dielectric layer for protective purpose or for other mechanical / Thermal requirements. In many of the practical design, this cover layer is electrically much thicker than the substrate thickness and can adversely affect the antenna basic radiation characteristics.

It has been found that [155], the bandwidth of the radiation pattern in the E plane will initially decrease and then increase upon increasing normalized thickness of the cover. It was found [156] that the antenna directivity can be changed by adjusting the loading of dielectric resonator core or by varying the permittivity of dielectric resonator material. The effect of coupling on various parameters of the microstrip antenna, and consequently effect of inter elements spacing with different dielectric materials placed in between the patches have been investigated [157, 158].

The effect of angle dependant perturbation of circular patch antenna due to alumina strips as overlay has been reported by Nuangnum et al [140]. Dual band microstrip antennas have been achieved by means of loading by shorting pins [159,160], crossed slot [161]. The influence of complex material covering on the quality factor of the antennas has been investigated by Tretyakov et al [162]. Multilayer microstrip antennas have been analyzed by workers [163, 164].

Most of the experimental data on overlay on thick film microstrip component in existence is from our lab [113-118,165-174] where dielectric overlays in bulk, thin film and thick film form has been studied on both thin film and thick film microstrip components. The work on thin film ring resonator as moisture sensor in X band has also been reported [108-110]. To the authors knowledge, the effect of the seeds on non resonant miniaturized Ag
thick film microstripline and resonant equilateral triangular patch antenna in the Ku band has not been studied by any worker.

1.9 **Aim of the work:**

The microstrip component being in planar form can offer an alternative miniaturized device for granular materials of uneven shape and size like oil seeds and cereal grains. The work in our lab has shown that thick film technology is cost effective and compatible for fabrication of microstrip components in the frequency range 8 - 18 GHz [175-178]. Resonant and non-resonant circuits are very useful in numerous applications in the microwave field. There are very few studies on the Ag thick film microstripline and equilateral triangular microstrip patch antenna upto 18 GHz.

To the authors knowledge using Ag thick film simple microstripline the dielectric characterization and prediction of moisture content of seeds has been reported for the first time in this work. The waveguide reflectometer and VSWR slotted section method has also been studies for the seeds for the first time. From the response of the various microstrip components, the microwave permittivity, conductivity, prediction of moisture content of seeds has been done here.

The thesis deals broadly with four aspects.

1. Design, fabrication and characterization of Ag thick film microstripline and equilateral triangular microstrip patch antenna in the Ku band.
2. Use of non-resonant simple microstripline to characterize the three different types of oil seeds by overlay technique and predict the moisture content percentage.
3. Use of resonant circuit equilateral triangular microstrip patch antenna to characterize the three different types of oil seeds by overlay technique and predict the moisture content (%).
4. The waveguide reflectometer and VSWR slotted line technique has also been used for the study of microwave dielectric properties and moisture content prediction.
In this work, an attempt has been made to study the dielectric properties of the moisture laden seeds and electrical parameters such as transmission, reflection, resonance frequency, bandwidth, and Q-factor through the Ag thick film microstripline circuits.

Three types of seeds were used for study in this work. All these seeds grown in Krishna Valley at Sangli district in Maharashtra state of India (June-Sept 2006) were selected. It has been used as overlay over two types of microstripline components viz. Ag thick film microstripline and Ag thick film equilateral triangular patch antenna. The seed was used as in touch overlay. Due to small seed size only partial overlay has been done.

The microwave dielectric properties and prediction of moisture content from overlay method have been compared with those obtained by waveguide reflectometer method and VSWR slotted line method.

The thesis is divided into six chapters. Chapter-I includes the relevant literature survey and theoretical considerations of microstrip components. The range of microwave frequency is also useful for agricultural purposes given. Various techniques has been explained. The experimental method used has been elaborated in Chapter-II. The data obtained from the experiments carried out by overlay technique using Ag thick film microstripline has been explained in the form of graphs and tables in Chapter-III. The dielectric characterization and prediction of moisture content are also elaborated here.

The data obtained from overlay studies using Ag thick film equilateral triangular patch antenna for the dielectric characterization and moisture prediction has been explained in chapter-IV. Chapter-V gives the results obtained by waveguide reflectometer and VSWR slotted section method for seeds. The discussion of the results of the studies of soybean, sunflower and groundnuts using microstrip components and waveguide techniques elaborated in chapter-VI. The summary and conclusions are also given in this Chapter.
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