Chapter III

Description of experimental methods
3.1 The slotted line

The slotted line is one of the important measuring instruments at microwave frequencies. It is designed to measure the standing wave pattern of the electric field intensity, which is a function of longitudinal position in the guiding structure. A probe is mounted on a carriage, which slides along the outside section of the coaxial section of the line or wave-guide, which has a longitudinal slot. The probe extends into the slot and is provided with an adjustment for varying the probe penetration into the slot and with a tuning adjustment used to cancel the reactive component of the probe impedance. The probe is connected to a barrater or crystal detector, which detects the r. f. voltage. This voltage is amplified and applied to the appropriate indicating meter.

A slotted section used over a frequency range from about 300 to 5000 MHz and the modified form is used over the frequency range in GHz is shown in figure (3.1.1). The standing wave ratio is measured by sliding the probe along the line for a maximum to minimum indication on the output meter.

The wavelength of the signal frequency can be measured by obtaining the distance between the minima since the distance between the successive maxima or minima is equal to half the wavelength.

*Slotted section model is shown in figure Figure 3.1.1*
Figure 3.1.1 Slotted section model
3.1.1 Errors in slotted line technique

The possible sources of errors associated with slotted line measurements must be carefully evaluated and proper operating technique must be applied in order to minimize these errors. The connectors on the coaxial slotted section limit the accuracy of standing wave ratio measurements since the slotted section probe responds to the combined reflections from the connector and the load beyond the connector. The uncertainty in the standing wave ratio measurements can be evaluated when the inherent S.W.R. of the slotted section is known.

Variation of maxima and minima at different points on the line is referred to as the slope error. It can be caused by the variation of the probe depth as the probe carriage is varied and also by energy leakage through the slot. This error can be adjusted to a minimum value in some slotted sections. [1]

3.1.2 Probe tuning error

One of the major sources of error in standing wave measurements is excessive probe penetration. The presence of probe affects the voltage standing wave ratio (V.S.W.R.) because it is essential an admittance shunting the line. Excessive coupling to the line cause a shift in maxima and minima and also cause the measured V.S. W. R. to be lower than the true (V.S.W.R.) In addition to the distortion of the field pattern, reflections from the probe vary when the probe is moved. Errors in measurement of low V.S.W.R. arise when these reflections are reflected from a mismatched source. Therefore, the probe coupling should be kept as small as possible except in case where it is only desired to examine the minimum point on the standing wave pattern. Using
can minimize excessive probe penetration high sensitive detector, assuming that there is adequate signal source power available. [1]

### 3.1.3 Harmonics and spurious signals

Harmonics are usually present in the signal sources that have coaxial outputs. Errors are possible when the probe is tuned to the harmonic of the fundamental. The frequency to which the probe is tuned can easily checked by measuring the half wavelength distance between two voltage minima on line. Harmonics are usually reduced to negligible value in coaxial system by the use of low pass filter.

Spurious signals are usually arise from improper adjustment of the modulating voltages used to square wave modulate a signal source such as a klystron which has several methods of oscillations. [1]

### 3.1.4 Frequency modulation

A variation of the instantaneous signal source frequency is referred to as frequency modulation. The minima of the standing wave pattern are obscured in the presence of frequency modulation since the minima of the standing wave pattern at different frequencies do not appear at the same position on the line. If the frequency modulation becomes excessive, it is possible that the other portion of the standing wave pattern can be distorted. Poor regulations of potentials applied to the oscillator are often responsible for frequency modulation problems. In ordered to prevent the frequency modulation of the modulated signal source, square wave modulation is used. The presence of frequency modulation on modulated source usually displayed on a scope using a frequency meter. The frequency modulation can also be detected by
investigation the minima of the standing wave pattern when the slotted section terminated with a short circuit. [1]

3.1.5 Detector characteristics

The characteristics of the barrater and crystal detector determine the power levels in the measurement system. These detector elements are usually used with equipment, which is calibrated in terms of squarer law response. In order to prevent a departure from the square law response, the barrater should be operated at power level less than 200µm and crystal should be operated at power levels less than 10 µm. In either case, the departure from the square law can be checked by noting the detector response on the standing wave indicator for different levels of input power. Measuring a fixed mismatch at different power inputs to the detecting element performs this check [1].

3.2 Elementary concepts of wave-guide

It is difficult to attach the physical significance of the mathematical procedure involved in determining the possible field and current distributions along the wave-guide structure. One method commonly used to derive the rectangular wave-guide from the two-wire transmission line shown in figure (3.2.1). Two-quarter wavelength sections support the transmission line and since the input impedance of the each section is theoretically infinite, they have no effect on the transmission power.
If the number of stub increased to infinite the rectangular wave-guide is formed as shown in figure. (3.2.1). It can be seen that a dimensions of the wave-guide cannot be less than one half wavelength. In fact, it must be slightly more than one half wavelengths in order to completely accommodate the transmission line function and the same time preserves the insulating properties of the quarter wave sections. Any frequency lower than that which makes the dimension a less than one half wavelength will cause the circuit to become an inducting shunt and there is no propagation. The frequency at which the dimension ‘a’ is one half wavelength (free space wavelength) is called the cutoff frequency and is designated \( f_c \). The free space wavelength is associated with this cutoff frequency is the cutoff wavelength designated ‘\( \lambda_c \)’ (\( \lambda_c = 2a \)).
3.3 Advantage of hallow wave guide

There is no power loss by radiation from the metal pipe of any type including coaxial line, if the ends are closed. Hallow wave guides are superior to coaxial line in ability to handle the large concentration are kept further apart so that the electric field is less intense.

The construction of the wave-guide is simpler than that of the coaxial cable since the inner conductor and its support are eliminated. It is also more rugged and less susceptible to variations and shock. Elimination of the insulating supports also results in a decrease in attenuation. Wave-guides are usually air filled, and for practical purpose they are consider to have no dielectric loss.

The current carrying capacity is grater since, in practice; the wave-guide is likely to have much grater-conducting surface than a coaxial line. The over all power lost as heat in the walls of the wave guide is lower than the heat dissipating in the conductors of conventional size coaxial lines [2].
3.4 Theory of rectangular wave guide

Consider a rectangular metallic wave guide having dimensions $a$ along X-axis and $b$ along Y-axis as shown in figure (3.4.1).

Figure 3.4.1 Metallic rectangular wave-guides

![Figure 3.4.1 Metallic rectangular wave-guides](image)

The transverse magnetic field (TM) where all magnetic fields are transverse to the direction of propagation $Z$. We have $H_Z = 0$, and all field components can be derived from a single longitudinal component $E_Z$, considering the equation

$$[\nabla^2 s + \omega^2 \mu \varepsilon - K_Z^2] E_Z = 0$$

(3.4.1)
and considering the bounding conditions of vanishing tangential electric field on the metallic wall surface, we obtain

\[ E_Z = \sin Kx \sin Ky e^{ikz} \quad (3.4.2) \]

The dispersion relation is

\[ K^2x + K^2y + K^2z = \omega^2\mu\varepsilon = k^2 \quad (3.4.3) \]

The traverse components are

\[ Ex = \frac{i kx kx}{\omega^2\mu\varepsilon - k^2z} \cos Kx \sin Ky e^{ikz} \quad (3.4.4) \]

\[ Ey = \frac{i ky ky}{\omega^2\mu\varepsilon - k^2z} \sin Kx \cos Ky e^{ikz} \quad (3.4.5) \]

\[ Ez = \frac{-i \omega \varepsilon ky}{\omega^2\mu\varepsilon - k^2z} \sin Kx \cos Ky e^{ikz} \quad (3.4.6) \]

\[ Hy = \frac{i \omega \varepsilon kx}{\omega^2\mu\varepsilon - k^2z} \cos Kx \sin Ky e^{ikz} \quad (3.4.7) \]

We see that at \( X = 0 \) and \( a \), \( Ex \) and \( Ey \) vanish, and at \( Y = 0 \) and \( b \), \( Ez \) and \( Ex \) vanish, provided that

\[ Kx^a = m\pi \quad (3.4.8) \]

\[ Ky^b = n\pi \quad (3.4.9) \]

Where \( m \) and \( n \) are integer numbers, Eq. (3.4.8 and 9) are the guidance conditions. For TM waves neither \( m \) nor \( n \) can be zero because then \( Ez \) will be zero, too. Substituting the guidance conditions (3.4.8 and 9) in the field equation, we see that for large \( m \) there is a large variation for the fields as a function of \( x \), and for large \( n \) there will be more field variation along the \( Y \) direction.
The dispersion relation (3.4.3) and the guidance conditions (3.4.8 and 9) combine to give the propagation constants,
\[
K_z = \sqrt{\frac{\omega^2 \mu \varepsilon}{\varepsilon} - \left( \frac{m\pi}{a} \right)^2 - \left( \frac{n\pi}{b} \right)^2}
\]  
(3.4.10)

According to the particular values of m and n, the TM waves inside the rectangular wave-guide are classified into TMmn modes. The first index m associated with the number of variations along the X direction, and the second index with the number of variations along Y direction.

Cutoff occurs when Kz becomes imaginary such that the wave attenuates exponentially along the direction of propagation. For a TM$_{mn}$ mode, the cutoff wave number is
\[
K_{cmn} = \sqrt{\left( \frac{m\pi}{a} \right)^2 + \left( \frac{n\pi}{b} \right)^2}
\]  
(3.4.11)

The lowest order TM mode is seen to be the TM11 mode. In figure (3.4.2), we plot the propagation constant Kz as a function of K for the case a = 2b. For example, if we take a = 3cm. and b = 1.5cm., we find the cutoff wave number for TM11 mode to be K$_{c11}$ = 234 m$^{-1}$
Next we examine TE fields which are derived from a single longitudinal component $H_z$ with $E_z = 0$, from the below equation (3.4.12).

$$\left[ \nabla^2 + \omega^2 \mu \varepsilon - K^2 Z \right] H_z = 0 \quad (3.4.12)$$

and considering the boundary condition of vanishing tangential electrical field on the metallic wall surface, we obtained

$$H_z = \cos K_x \cos K_y \, e^{ik_z z} \quad (3.4.13)$$

The dispersion relation is identical to equation (3.4.3), the transverse field components are found from the following equation,
\[
\bar{E}_s = \frac{1}{\omega^2 \mu \varepsilon - k^2 z} \left[ \nabla S \frac{\delta E_z}{\delta z} + i \omega \mu \nabla S. \bar{H}_z \right] \tag{3.4.14}
\]

\[
\bar{H}_s = \frac{1}{\omega^2 \mu \varepsilon - k^2 z} \left[ \nabla S \frac{\delta H_z}{\delta z} + i \omega \mu \varepsilon S. \bar{E}_z \right] \tag{3.4.15}
\]

We obtained,

\[
H_x = -\frac{i k_x k_z}{\omega^2 \mu \varepsilon - k^2 z}. \sin K_x x. \cos K_y y e^{i k z} z \tag{3.4.16}
\]

\[
H_y = -\frac{i k_y k_z}{\omega^2 \mu \varepsilon - k^2 z}. \cos K_x x. \sin K_y y e^{i k z} z \tag{3.4.17}
\]

\[
E_x = -\frac{i \omega \mu k_y}{\omega^2 \mu \varepsilon - k^2 z}. \cos K_x x. \sin K_y y e^{i k z} z \tag{3.4.18}
\]

\[
E_y = \frac{i \omega \mu k_x}{\omega^2 \mu \varepsilon - k^2 z}. \sin K_x x. \cos K_y y e^{i k z} z \tag{3.4.19}
\]

The guidance conditions are obtained from the boundary conditions of \(E_x = 0\) at \(Y = 0\) and \(E_y = 0\) at \(x = 0\) and \(a\).

The result is identical to the equation (3.4. 8and 9) the propagation constant again given by equation (3.4.10) and cutoff wave number by the equation (3.4.11).

There is, however, a very significant difference between TE and TM modes. We noted that for TM mn modes neither \(m\) nor \(n\) could be zero. For the TEmn modes it is possible to have \(m\) or \(n\) or both \(m\) and \(n\) are equal to zero. For \(m = n = 0\),

We find that \(H_z = e^{i k z}\). The equation \(\nabla \times \bar{H} = 0\), implies \(K = \omega(\mu \varepsilon)^{1/2} = 0\) and consequently the TM\(_{00}\) mode is static field solution in the wave-guide.
We now argue that the TEM mode for which $E_z = H_z = 0$ cannot exist inside the hollow wave-guide. From $\nabla \cdot H = \nabla \cdot H_S = 0$

And the boundary conditions of vanishing normal H field on the wall, we find that H field must form a closed loop and $\nabla \times H_S = \hat{z} j z - \hat{z} \omega \varepsilon \varepsilon_z \neq 0$. Thus either $J_z \neq 0$, which implies that the wave-guide is not hollow or $E_z \neq 0$, which implies that the mode can be TEM. As a corollary, we have in fact shown that TEM mode do exist when there is another conductor to support the conduction current $J_z$, as in the case of a coaxial line.

Assuming $a > b$ we see from equation (3.4.11) that the lowest order TE mode will be $TE_{10}$. With cutoff wave number

$$Kc_{10} = \frac{\pi}{a} \quad (3.4.20)$$

The field component of the $TE_{10}$ mode is

$$H_z = \cos \frac{\pi x}{a} e^{i k z} \quad (3.4.21)$$

$$H_x = -i k z a / \pi \sin \frac{\pi x}{a} e^{i k z} \quad (3.4.22)$$

$$E_y = i \omega \mu a / \pi \sin \frac{\pi x}{a} e^{i k z} \quad (3.4.23)$$

The electric field has only a y component. A field plot of $TE_{10}$ mode is shown in figure (3.4.2). Equation (3.4.23) Becomes

$$E_y = \frac{\omega \mu a}{2\pi} \{ e^{i \pi z/a + i k z z} - e^{-i \pi z/a + i k z z} \} \quad (3.4.24)$$

Higher order TE and TM modes can also be interpreted as plane wave bouncing around the four walls and propagating along Z with propagation constant $K_z$. The propagation constant ($K_z$) for the various modes plotted in figure 5.4.2, for the case of $a = 2b$. Since the $TE_{10}$ mode has the lowest cut off...
frequency, it is the fundamental mode or the dominant mode of the rectangular wave-guide [3]

3.5 Review of experimental methods

The method is divided into two groups
1) Resonance method and
2) Non-Resonance method

3.5.1 Resonance method

Let us consider a cylindrical resonator of volume V, natural frequency $f_0$ and quality factor $Q_0$. Upon introducing the dielectric into the cavity, the resonance frequency will change to $f_1$. Similarly, the quality factor will also change to $Q_1$, depending on the shape, size and dielectric constant the sample introduce into the cavity. Under idealized conditions of very small coupling, cavity shape of high symmetry, high conductivity and low dielectric loss, the power $P$ at the detector as the input frequency $f$, is given by,

$$P_{\text{max}} = \frac{1}{\left[ 1 + 4Q^2 (f-f_{\text{max}})^2 / f_{\text{max}}^2 \right]^{-1}}$$  (3.5.1)

Where $P_{\text{max}}$ is the maximum power at which maximum power reaches the detector. The expression for dielectric permittivity can be obtained as follows,

$$\varepsilon' = (f_0 / f_1)^2$$  (3.5.2)

$$\varepsilon'' = (1/Q_1 - 1/Q_0)$$
Different groups have been used different types of cavities for this experimental technique, depending upon the frequency range and the type of material used. Works, Dakin and Boggs have used a double re-entrant cavity of high Q\(>\)(2000) made from a length of coaxial line short circulated at both ends, with a small cylindrical section removed from central conductor near one end of cavity. The sample, the cylinder of the same dimension as the removed section is placed near the cut. The accuracy of this found to be ±1 percent in \(\varepsilon\) and ± 5 \(\times \)10\(^{-5}\) percent in \(\tan\varepsilon\). A microwave resonance technique studied for high loss material has been developed by Birnbaum, Kryder and Lyons, with the accuracy of ±4 percent in \(\varepsilon\) and ± 2 percent \(\varepsilon''\). The method can be used from 3GHz to 24 GHz. Pitt and Smyth has developed a short circular coaxial line of adjustable length to which the generator and detector are connected loosely. The cavity can be used over the range of 0.5 to 3 GHz with an accuracy of 0.5 ± percent in \(\varepsilon'\) and 3 percent in \(\varepsilon''\).

**3.5.2 Non-resonance method**

The classical methods of Durbe for the centimeter wave have been employed for this technique. The method can be described as follows.

Consider a wave-guide shorted at one end shown in figure (3.5.1) with wave propagating along it. As a result of the interference between the incident and reflected waves a standing wave will from in the wave-guide; its distribution is shown in Figure (3.5.1).
Figure 3.5.1 Standing waves in a wave guide

(a) Without dielectric

(b) Loaded with dielectric of any length

(c) Loaded with dielectric of length

Description of experimental methods
If the ends of the wave-guide are filled with dielectric layer of thickness d then as the result absorption, the amplitude of the reflected wave will differ from the amplitude of incoming wave, end the distribution of amplitude in standing waves changes as shown in figure (3.5.1). The amplitude of the electric field at nodes is no longer zero, and the maxima minima of wave shift. First of all, the shift is related to the changes of the wavelength in the dielectric as compared with its length in air Roberts and Von Hipple worked out a method determining the complex permittivity form measurements of the parameters of the standing wave, which arises in the air filled part of the wave guide.

The principle equation relating the parameter, which described the standing in the wave-guide

\[
\frac{\tan h\gamma_2d}{\gamma_2d} = -j \left( \frac{\lambda}{2\pi d} \right) \frac{ks - j\tan\theta}{1 - j ks \tan\theta} = c e^{j\xi} \quad (3.5.1)
\]

where,

\[
\gamma_2 = j\omega\sqrt{\varepsilon_0 \mu_0} = \alpha_2 + j\beta_2 \quad (3.5.2)
\]

\(\gamma_2\) is characteristics propagation coefficient for the dielectric layer, \(\alpha_2\) is the attenuation coefficient and \(\beta_2 = 2\pi/\lambda_2\) is the phase shift for the dielectric layer. \(\lambda\) is the wavelength in the air filled part of the wave guide,

\[Q = 2\pi \varepsilon_0 / \lambda_0\]

where \(\varepsilon_0 = (\lambda / 2) - d - 1\) is the distance between the first minima of the wave from the surface of the dielectric and 1 is the shift of minima brought about by the introduction of the dielectric into the wave guide figure (3.5.2) on the other hand is the reciprocal of the voltage coefficient of the standing wave.
Figure 3.5.2. Determination of standing wave coefficient

\[ K_s = \frac{E_{\text{min}}}{E_{\text{max}}}, \]  \hspace{1cm} (3.5.3)

\[ E_{\text{min}} \text{ and } E_{\text{max}} \] respectively, are the minimum and maximum amplitudes of the standing wave electric field in the wave-guide figure (3.5.2). For low loss material (\( k_s < 0.1 \))

\[ K_s = \frac{\pi \delta x}{\lambda}, \] \hspace{1cm} (3.5.4)

Where \( \delta x \) is the distance between the points of the standing wave where the intensity of the current flowing into the detector is twice as high as the minimum value. Figure (3.5.1) and \( d \) is the thickness of the dielectric.
layer. From measurements parameters $\lambda$, $d$, $k$, and $X_0$ we can find the functions $C \exp^{j\xi}$ and the function

$$\gamma_2 \ d = \tau \exp^{j\tau} \quad (3.5.5)$$

can be found from the plot of equation

$$\frac{\tan hT \ e^{j\tau}}{T \ e^{j\tau}} = c \ e^{j\xi} \quad (3.5.6)$$

In the plot $\tau$ is on Y-axis, and the absolute magnitude of $T$ is on X-axis. $C$ and $\xi$ are the parameters of the intersection curves. From equation (3.5.3) we get

$$C = \frac{\lambda}{2\pi d} \ \sqrt{(ks + \tan^2\theta) / (1 + k^2 s \tan^2\theta)}$$

and

$$\tan \xi = \frac{k (1 + \tan^2\theta)}{(1 - k^2 s) \tan \theta} \quad (3.5.7)$$

The complex electric permitivity of the dielectric is given by the equation.

$$\varepsilon^* = \frac{(\gamma_2 / 2\pi)^2 + (1 / \lambda)^2}{(1 + \lambda c)^2 + (1 + \lambda)^2} \quad (3.5.8)$$

where $\lambda c$ is the limiting length of the wave-guide for rectangular wave-guide $\lambda c = 2 \times$ the width, i.e. $\lambda c = 2 \times a.$
3.6 Basic equation for low loss dielectric materials

Consider an Em wave at the surface separating air and dielectric specimen by a thin mica foil at \( Z = 0 \). It is represented by \( E_0 \exp \{ j \omega t \} \) where \( E_0 \) is the maximum amplitude and \( \omega \) is the angular frequency of the microwave electric field. The wave is transmitted through the sample of length \( d \) and reflected at the shorting plunger. The reflected wave at \( Z = 0 \) is given by

\[
E_0 \exp \{ j \omega t - 2 \gamma_1 d + j \delta \}. \tag{3.6.1}
\]

Where \( \gamma_1 \) is the propagation constant in the medium is given by

\[
\gamma_1 = \alpha_1 + j \beta_1
\]

here \( \alpha_1 \) is propagation constant and \( \beta_1 \) is phase constant, and \( \epsilon \) is the phase change by reflection due to plunger. The total amplitude of the resulting wave in the air medium is given by

\[
E = E_0 \exp \{ j \omega t \} + E_0 \exp \{ j \omega t - 2 \gamma_1 d + j \delta \}. \tag{3.6.2}
\]

The power \( P \) at \( Z = 0 \) is given by

\[
P = EE^* = |P|^2
\]

\[
P = P_0 [1 + e^{-4 \alpha_1 d} + 2 e^{-2 \alpha_1 d} \cos (2 \beta_1 - \delta)] \tag{3.6.3}
\]

Keeping the plunger at \( Z = 0 \) (mica sheet) the slotted line probe is moved, till the digital power meter (DPM) connected to the crystal detector resisters zero current. Now keeping the slotted line probe at this position, the shorting plunger is moved back and the standing wave power i.e. current at that location is recorded for different sample thickness.
Equation 2.6.3 is used to determine the different parameters via; attenuation constant ($\alpha_1$), phase constant ($\beta_1$), power at infinite thickness $P_0 = \{E_0\}^2$ and the phase change $\delta$.

### 3.7 Basic equation for high Loss Materials

For high loss materials, one has to consider multiple reflections in the material. Using boundary conditions at each interface and solving the resulting simultaneous equations can find total reflection coefficient. The condition at the plunger yields.

$$E_{10} \exp (-\gamma_1 d) + R E_{30} \exp (\gamma_1 d) = 0 \quad (3.7.1)$$

where $E_{10}$ is the transmitted electric field from dielectric media to air, $E_{30}$ is the amplitude of the wave traveling in negative X – direction in the dielectric and $R$ is the reflection coefficient due to plunger, other boundary conditions for the transmitted ($E_{00}$) and reflected ($E_{20}$) waves are

$$E_{00} + E_{20} = E_{10} + E_{30} \quad (3.7.2)$$

$$\gamma_2 (E_{00} - E_{20}) = \gamma_1 (E_{10} - E_{20}) \quad (3.7.3)$$

The above three equation to get the value of $(E_{20}/E_{00})$ as

$$\frac{E_{20}}{E_{00}} = \frac{\left(\frac{\gamma_2}{\gamma_1}\right) \tan h \gamma_1 d + R}{\left(\frac{\gamma_2}{\gamma_1}\right) \tan h \gamma_1 d + R_0} \quad (3.7.4)$$

and

$$\frac{E_{20}}{E_{00}} = \frac{\left(\gamma_2 - \gamma_1\right) \exp (2\gamma_1 d) + (\gamma_2 + \gamma_1) R}{\left(\gamma_2 + \gamma_1\right) \exp (2\gamma_1 d) + (\gamma_2 - \gamma_1) R_0} \quad (3.7.5)$$

The standing wave in the wave guide can be describe as
The power $P$ at the position is given by:

$$P = P_0 \left| \left( \exp(-\gamma_2 x) + \frac{E_{20}}{E_{00}} \exp(+\gamma_2 x) \right) \right|^2 \quad (3.7.7)$$

In our experiment, the probe position was kept at minima ($X = 0$) when the plunger was shorted. The above equation 3.7.7 becomes $P = P_0 |1+Q|^2$, where $Q = \frac{E_{20}}{E_{00}}$ as given in equation 3.7.4.

$$P = P_0 \left| e^{-\gamma_0 x} + \frac{\left( \gamma_2 \cdot \gamma_1 \right) \exp(2\gamma_1 d) + (\gamma_2 + \gamma_1) R \exp(+\gamma_0 x)}{\left( \gamma_2 + \gamma_1 \right) \exp(2\gamma_1 d) + (\gamma_2 - \gamma_1) R_0} \right| \quad (3.7.8)$$

where $\gamma_2 = \alpha_2 + j\beta_2$ is the propagation constant in the wave guide section preceding the dielectric cell, $\gamma_1 = \alpha_1 + j\beta_1$ is the propagation constant in the dielectric cell. $R = R_0 \exp(j\phi)$ is the complex reflection coefficient of the plunger in contact with the dielectric. This equation (3.7.8) was fitted in experimental data using $\alpha_1, \beta_1, \phi$ and $P_0$ as fitting parameters. The value of $\beta_z (2\pi / \lambda_g)$ was measured experimentally and $R_0, \alpha_e$ are assumed to be one and zero for ideal plunger respectively.
3.8 Method of experimental analysis for soil

The experimental technique used to measure the dielectric permittivity and water content is that of Roberts and Von Hipple and Gopal Krishna. A least squares fit programme of Sobhanadri is used to calculate the dielectric permittivity. A X-band microwave transmission line waveguide setup is used for this purpose. The experimental setup of X-band wave guide transmission line is shown in Figure 3.6.1. The dielectric permittivity and loss of soils have been measured for two different frequencies i.e. at 9-GHz and 11-GHz and at room temperature. The soil samples were collected from different locations,
which are 10 to 30 km away from Aurangabad city. The samples were collected from these areas to determine its physical and chemical properties. These samples were collected from non-irrigated farming lands in hot summer with negligible water content, called as dry soil. These soil samples are the mixture of sand, silt and clay with very high percentage of clay.

The soil sample under measurement of known volume was placed in the empty solid dielectric cell, and well pressed by a laboratory developed mechanical system (see Figure 3.7.1) to remove the air and discontinuities in the sample. The solid cell with sample was connected to the opposite end of the source of microwave bench set up. The signal generated from the microwave source was allowed to incident on the soil sample. The soil sample reflects part of the incident signal through the soil from its front surface. The values of power at different points of standing waves have been measured as a function of probe position. About (80 – 100) points were recorded for a single standing wave pattern. The least squares fit has been used to determine the values of $\lambda_0$, $\lambda_c$, $\alpha$ and $\beta$ for the sample.

Firstly, emerging the microwave transmission line the standing wave pattern was recorded for empty cell. The soil under measurement was placed in the cell pressed it, and connecting it to the other end of the microwave bench setup. The measurements were conducted for three different lengths of the collected soils. The same procedure is applied for other soil samples. Fitting these standing wave patterns of dry soil samples along with empty standing wave pattern into least squares fit programme, the dielectric permittivity and loss were determined.
The free space wavelength was determined from the following equation

\[
\frac{1}{\lambda^2_0} = \frac{1}{\lambda^2_g} + \frac{1}{\lambda^2_c} \quad (3.8.1)
\]

Where \( \lambda_c \) is the cutoff wavelength determined by

\[
\lambda_c = 2 \times a = 2 \times a
\]

where ‘a’ is the border side of the rectangular waveguide. The real and imaginary parts of the permittivity are

\[
\varepsilon^* = (\varepsilon' - j\varepsilon'') \quad (3.8.2)
\]

have been determined from the following equations

\[
\varepsilon' = \lambda^2_0 \left[ \frac{1}{\lambda^2_c} + \frac{(\beta^2_1 - \alpha^2_1)}{4\pi^2} \right] \quad (3.8.3)
\]

\[
\varepsilon'' = \frac{\lambda^2_0 \beta_1 \alpha_1}{2\pi^2} \quad (3.8.4)
\]
3.9 X-Band microwave bench set-up

The necessary components are as follows for X-band Set-up.

1) Sweep Oscillator     2) Coaxial connector
3) Isolator            4) Attenuator
5) Slot line           6) Solid dielectric cell
7) Power Meter

Experimental set-up of Microwave X-Band is given in figure 3.9.1.
Fig. 3.9.1 Block diagram of experimental set-up of microwave X-band
Figure 3.9.2. Photograph of microwave X-band setup
3.10 Special component used in the experimentation

3.10.1 Agilent 53147A/148A/149A Microwave frequency counter/power meter/ DVM

a) MEASURING FREQUENCY

1. Connect the instrument to a power source when the instrument is connected to an AC power source, the Standby indicator on the front panel lights. The standby indicator also lights if the instrument is connected to an external DC power source or is operated from internal batteries and the battery power switch is on (with the Battery option only).

2. Press the POWER button on the front panel. The standby indicator goes off, and all segments of the front-panel display are temporarily activated. TESTING is displayed while the instrument performs its power-on self-test. If the instrument passes all of the tests, SELF TEST OK is displayed, and the instrument displays its model number, firmware version number, GPIB address, and CH2 NO SIGNAL. The Counter is now ready to measure the frequency of a signal applied to the Channel 2 input. Note that the Ch 2 Freq annunciators are activated.

3. Connect an input signal to Channel 2. The Channel 2 input path circuits contain sensitive GaAs semiconductors. To prevent damage to these components, always adhere to standard ESD (Electro Static Discharge) prevention procedures, and ensure that the maximum power specification for this channel (+27 dBm) is not exceeded.
4. To measure the frequency of a signal applied to the Channel 1 input, press the Chan Select key. CHANNEL 1 is displayed momentarily, and the Ch 1 and Freq annunciators are activated. If a signal is presently applied to the Channel 1 input, the measured frequency is then displayed. If no signal is applied, CH1 NO SIGNAL is displayed until an input signal is connected to the Channel 1 input connector.

b) MEASURING POWER

The Agilent 53147A/53148A/53149A can measure signal power in the power and frequency ranges. The power measurement, which is shown in a dedicated area of the display, includes a digital readout and an analog representation. The display, which can be configured to shows power in units of dBm or Watts, is auto-ranging when set to measure in watts.

Selecting a Power Head (Sensor)

There are a number of Agilent power heads that can be used with the Power Meter in this instrument. Choosing the appropriate power head is a matter of matching the head’s characteristics to the signal to be measured.

Before one can make any power measurement, one must determine which power head (sensor) to use for the measurement, select the power head in the instrument’s menu and configure the power meter to use the appropriate calibration factor for the frequency of the signal. The five power head models that have recorded calibration-factor tables in the instrument’s non-volatile memory are listed. One can also modify the data points (frequency/calibration-factor data pairs) in the preconfigured calibration-factor tables, add data points to these tables, and add up to three custom tables for power heads that are not
included in the instrument’s menu. Instructions for modifying and adding data points in calibration-factor tables are in “Modifying and Adding Calibration Factor Tables”.

**MAKING A POWER MEASUREMENT**

When you turn the power Meter on, you must always zero and calibrate it with the power head connected before making any measurements. If you are using a different model power head than the one used the last time the Power Meter was used, you must also set the power head model in the instrument’s menu. As part of the measurement sequence, you must input either the frequency of the signal you intend to measure or the power factor for that frequency.

**POWER MEASUREMENT EXAMPLE**

The instrument must be powered on and must remain at the same ambient temperature for 15 minutes before beginning this procedure. If the temperature changes by 5°C or more, wait another 15 minutes.

1. Connect the output cable from the power-meter head to the Power Meter connector. This example assumes that you have a power head available that is appropriate for the measurement to be taken.

2. Press the Display Power key to enable power measurement. The Power annunciator at the left side of the display is activated, and the Power Meter’s digital and analog power displays show the power measurement in Db or dBm (the default units of power measurement).

3. Press the shift key, and then press the menu (Reset / Local) key. One of the items in the instrument’s menu is displayed (if the menu has not been used since the instrument was turned on, the initial menu display is “REF OSC>INT”).
4. Use the up and/or down arrow keys to cycle through the menu until “HEAD > OFF” is shown.

5. Press the right arrow key. The flashing indicator after HEAD changes from to, and “OFF” (or the currently selected power head model number) begins to flash.

6. Select the model number of the power head you intend to use by pressing

   The up- and/or down-arrow key repeatedly until the correct model number is displayed, and then press Enter.

7. Press the Zero key. The Power Meter displays ZEROING and then returns to the display shown in step 2.

8. Connect the power-head input connector to the Power Meter Output Connector.

9. Press the Call key. During calibration, the Power Meter displays CALIBRATING. It then returns to the display shown in step 2.

10. If you know the frequency of the signal you intend to measure, press the Freq. key, enter the frequency value, and press Enter. The Power Meter uses the frequency to set the power-factor per the values in the stored calibration tables. If you prefer, you can use the Cal Factor key to enter the calibration factor value directly.

11. Disconnect the power-head input connector from the Power Meter output Connector and connect it to the signal to be measured.

12. To measure the signal power in Watts, press the Shift key, and then press the dBm/W (Display power) key. When you press the Shift key, the Shift annunciator is activated. When you press the dBm/W
(Display power) key, the Shift annunciator goes off, and the units of measurement annunciator group to the right of the digital power measurement changes from dB or dBm to Watts, mW, microwatt or nanowatt.

Photograph of the Agilent microwave power/frequency meter is shown in Figure.
Fig. 3.10.1 Photograph of the Agilent microwave

Power/Frequency meter
Modifying and adding calibration factor tables

The Head menu option provides access to reconfigure, editable calibration-factor tables for vive models of Agilent power sensor heads (models 8481A, 8481D, 8482A, 8485A and 8487A) and three custom tables. You can modify the frequency/calibration-factor values in any of the data points for any power head, and you can input data to build new calibration Tables.

<table>
<thead>
<tr>
<th>Calibration table data points</th>
<th>modify</th>
<th>delete</th>
<th>reset</th>
<th>add</th>
</tr>
</thead>
<tbody>
<tr>
<td>Factory defined data points (for 8481A, 8481D,8482A,8485A and 8487A)</td>
<td>all</td>
<td>none</td>
<td>All</td>
<td>yes</td>
</tr>
<tr>
<td>Data points added to factory-defined tables</td>
<td>all</td>
<td>all</td>
<td>N/A</td>
<td>N/A</td>
</tr>
<tr>
<td>By user</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Custom calibration-factor tables</td>
<td>all</td>
<td>all</td>
<td>N/A</td>
<td>Yes</td>
</tr>
</tbody>
</table>

Calibration table modification example

This example demonstrates how to view and modify the values in any of the
Preconfigured calibration tables or in one of the three custom calibration

tables

1. Press the shift key and then press the menu (reset/local) key.
2. Use the up and/or down arrow keys to cycle through the menu until “HEAD>OFF” is shown
3. Press the right arrow key
4. Select the model no. of the power head you intend jto modify the
data points for by pressing the up-and/or down-arrow key
repeatedly until the correct model no. is displayed, and then press Enter.

5 Press the right arrow key.

6 Press the right arrow key again

7 When you move the focus to the last digit of the frequency value, an additional flashing indicator appears at the right end of the display and indicator to the left of the frequency value stops flashing.

8 Press the right arrow key until focus moves from the last digit of the frequency value to the first digit of the calibration factor value. The flashing indicator at the right of the frequency value changes direction, which indicates that you can use the left arrow to return to frequency value, if you need to.

9. If you want to change values in additional data points, res. the right arrow key to save your changes and move to the next data point. If you are done changing data point values, press the Enter key to save your changes exit the menu.

3.10.2 Instrument used for pressing the soil

In order to remove the discontinuities in the soil sample the laboratory developed mechanical system was used in the present research work. This system is as shown in Figure 3.10.2.
Figure 3.10.2 Mechanical system for removing discontinuities in the soil sample dielectric cell
References


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