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has been approved for the

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To my parents Kazi Harunur Rashid and Kazi Hasina Rashid

for their love and support when required and when not, to my

sister Kazi Safina Shahreen for her unconditional love
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# TABLE OF CONTENTS

## CHAPTERS

### I. INTRODUCTION

1.1 TERAHERTZ ................................................................. 2  
   1.1.1 Overview of Terahertz ............................................... 2  
   1.1.2. Terahertz Source ................................................ 3  
      1.1.2.1 Electronics based Terahertz Source ..................... 3  
      1.1.2.1.1 Vacuum based Terahertz Source ..................... 3  
      1.1.2.1.2 Solid-state based Terahertz Source ............. 4  
      1.1.2.2 Optics based Terahertz Source .............................. 4  
   1.1.3 Application of Terahertz Spectrum .......................... 4  

1.2 SCOPE AND MOTIVATION OF THE THESIS: ........................................ 8  

1.3 NOVELTY OF THE PROPOSED WORK ........................................... 10  

### II. DIELECTRIC PROPERTIES AND MEASUREMENT THEORY

2.1 DEFINITION OF DIELECTRIC PROPERTIES ........................................... 11  

2.2 POLARIZATION OF DIELECTRICS .................................................. 13  
   2.2.1 Electric Polarization .................................................. 14  
   2.2.2 Orientational (Dipole) Polarization: ........................................ 15  
   2.2.3 Ionic (Molecular) Polarization: ........................................... 16  
   2.2.4 Frequency Dependence of Dielectric Mechanism ..................... 16  

2.3 DIELECTRIC MEASUREMENT COMPLICATION FACTOR ............................. 18
2.3.1 Dispersion ................................................................. ................................................... 18
2.3.2 Anisotropy ................................................................. ................................................... 22

2.4 Complex Permittivity Measurement ................................................................. 23

2.4.1 Complex Permittivity measurement of solids ................................................... 23
2.4.1.1 Single Layer Medium ................................................................. ................................................... 23
2.4.2 Compound Materials ................................................................. ................................................... 26
2.4.3 Complex Permittivity Measurement of Liquids ................................................... 27

III. METHODS FOR MATERIAL CHARACTERIZATION

3.1 Measurement Techniques .................................................................................. 34

3.1.1 Coaxial Probe Method .................................................................................. 34
3.1.1.1 High-Performance Probe .......................................................................... 36
3.1.1.2 Slim probes ............................................................................................... 36
3.1.1.3 Performance Probe .................................................................................. 37
3.1.1.4 Strength and Limitation .......................................................................... 38

3.1.2 Transmission Line Method ........................................................................... 38
3.1.2.1 Condition of Transmission Line Measurement ......................................... 39

3.1.3 Free space Measurement Method .................................................................. 41
3.1.3.1 Free space Measurement procedure: ....................................................... 42
3.1.3.2 Condition for Accurate Free space Measurement ..................................... 47
3.1.3.3 Free-space Calibration ............................................................................ 49
3.1.3.3.1 TRL Calibration .................................................................................. 50
3.1.3.3.2 TRM calibration .................................................................................. 52
3.1.3.3.3 GRL calibration .................................................................................. 53

3.1.4 Resonant Cavity Method .............................................................................. 55
5.5.2 Measurement data for 170-260 GHz ................................................................. 120
5.5.3 Comparison Between Measurement and Simulation Data .................................. 128
5.6 DISCUSSIONS ........................................................................................................ 134
5.7 LIMITATION OF STUDY ...................................................................................... 135
5.8 FUTURE WORK ..................................................................................................... 136

REFERENCES .............................................................................................................. 137

APPENDICES .............................................................................................................. 142

☐ APPENDIX A ............................................................................................................ 142
  Quasi Optic Measurement System Configuration Software ........................................ 142

☐ APPENDIX B............................................................................................................ 150
  Completions of the Institutional Review Board (IRB) with Collaborative Institutional Training Institute (CITI) ...... 150
CHAPTER 1

INTRODUCTION

The modern world we live in today is built on the utilization of electromagnetic spectrum. Science and technology based on the electromagnetic spectrum is being used in every aspect of our lives; from electricity generation to daily communication to outer space exploration. Every part of the spectrum plays part in our daily life today. For instance, radio spectrum is used for television, mobile, radar and radio communication, x-ray for medical imaging and security screening, gamma ray for cosmic observation, infra ray for fiber telecom, and remote sensing and ultraviolet ray is used for dental curing. This has become possible due to the dedication and efforts of scientists and engineers for over two centuries in the research exploiting electromagnetic spectrum. However, there is a region in electromagnetic spectrum called terahertz region which is still relatively unexplored. This is mainly due to the lack of information about its characteristics in nature. Due to the potential of the utilizing this spectrum in various including medical imaging, high-speed communication and chemical characterization, scientific communities from all over the world are actively conducting research in this area. In this thesis, a unique characteristic of terahertz spectrum has been exploited to perform research in dielectric spectroscopy for cancer screening and detection.
1.1 Terahertz

1.1.1 Overview of Terahertz

Variety of definition is present regarding lower and upper edge of terahertz frequency range. It is mainly due to different terahertz sources and various terahertz application with different operation bands. Usually, the electromagnetic waves with frequencies from 100 GHz to 10 THz with the corresponding wavelength of .03 – 3 mm are referred to as terahertz (THz) region. In some cases, frequencies between .3 to 3 THz is considered as terahertz band as well. In the electromagnetic spectrum, the terahertz region lies between the microwave and infrared bands as illustrated in Figure 1. Although, terahertz radiation was observed almost 100 years ago, it was relatively unexplored due to the lack of the development of terahertz sources and detectors. Hence, a gap, mostly referred as ‘terahertz gap’ had been created. However, a significant amount of research have been conducted for the last decade in developing effective powerful terahertz sources and detectors.

Figure 1.1: Terahertz portion in the Electromagnetic spectrum [2]
1.1.2. Terahertz Source

Depending on the application, terahertz sources can be different. For some applications, the source is desired to have spectral purity, large bandwidth, and tunability. On the other hand, some application may require the source to have the ability to generate enough power so that it can overcome atmospheric attenuation. In technical aspects, the terahertz source can be classified into two broad categories: electronics and optics [1].

1.1.2.1 Electronics based Terahertz Source

Electronics based terahertz sources are compact in sizes. Electronics sources are divided into two categories: solid state and vacuum. Vacuum devices generate higher average power than solid state source. Vacuum devices incorporate vacuum tube in which electron with certain kinetic energy travels through and is controlled in the tube.

1.1.2.1.1 Vacuum based Terahertz Source

Vacuum devices are grouped into two categories depending on how fast electron travels through the tube. One is the fast-wave and the another is the slow-wave. The fast-wave vacuum source include gyrotron and free electron laser (FEL). FEL uses two component, accelerator and undulator to generate terahertz signal. Undulator produces a magnetic field which drives an electron beam from the accelerator to achieve coherent terahertz pulsed radiation. Gyrotron generates continuous terahertz wave when it moves along the tube in gyration motion. Slow-wave sources are traveling wave tube (TWT), klystrons and backward oscillators (BWOs). In traveling wave tube (TWT) and backward oscillators (BWO), velocity modulation is applied by the RF coil surrounded by the tube while traveling through the tube. It is referred to as slow-wave since the electron velocity decreases from the speed of the light in free-space.
1.1.2.1.2 Solid-state based Terahertz Source

Solid-state devices are classified into two categories: active and passive depending on how the device is implemented. Active device source uses transistor whereas passive device uses diodes. Depending on the technology used for fabrication, the active device can be of two types: Si-technology device and III/V-technology device. Silicon-Germanium (SiGe) Heterojunction Bipolar Transistor (HBT) and Si MOSFET are Si-technology devices. High Electron Mobility Transistor (HEMT) is an example of III/V-technology device. Passive devices are Resonant Tunneling Diodes (RTD), IMPact Ionization Transit Time (IMPATT), and Gunn diodes. The concept of negative resistance across two terminal is utilized to generate terahertz signal in these devices.

1.1.2.2 Optics based Terahertz Source

Optics based terahertz sources are classified into two categories: continuous wave and pulse. Quantum Cascade Laser (QCL), gas lasers, p-type germanium laser are the continuous-wave terahertz optic sources which follow photo mixing method for wave generation. Photo mixing method is basically the combination of two different waves that results in the difference between two frequencies of the incident wave. The other type of optics based source is the pulsed one. This is the most widely used source in imaging and spectroscopy. For example, it allows femtosecond laser to incident onto photoconductive antenna to generate terahertz wave.

1.1.3 Application of Terahertz Spectrum

Vast efforts have been made in last couple of decades in research and development of terahertz technology as it has been proved that terahertz has unique characteristics that can be applied to various fields such as imaging, security, spectroscopy, medical field, and
communication. It can provide very high-resolution images. For example, researchers from Max-Planck-Institute of Biochemistry (MPIB) was able to demonstrate the mapping of free-carriers in state-of-the-art industrial transistors of 65 nm-technology by developing high-resolution atomic microscope exploiting terahertz’s sensitivity to the conductivity of semiconducting materials [2]. It is observed that certain portion of terahertz band offers 100-fold increased sensitivity to the conductivity of semiconducting material when compared to infrared light [2]. Its ability to offer high-resolution image is also being used in medical imaging. When screening cancer tissues, its high-resolution feature allows observing precise delineation of cancer tissue from the surrounding normal tissues. It is also useful in detecting early detection of tooth decay. Moreover, terahertz is non-ionizing radiation which means it doesn’t have enough energy to alter the structure.

![Figure 1.2: Terahertz fingerprints of the explosives][2]
of human tissues by moving atoms or molecules from them. One significant characteristic of terahertz is the chemical fingerprint. Most of the explosive and hazardous materials absorb photons when stimulated at terahertz frequencies. These absorption bands act as chemical fingerprints of those corresponding materials. This particular behavior of terahertz frequency is used in developing terahertz Sensor which is a technology for security and defense application. It has the ability to detect different types of explosives even if it’s hidden in the layer of different materials. It is considered a major advancement as conventional technology can only detect metal. An illustration of terahertz fingerprints for explosives materials is shown in Figure 1.2. It can be seen that different explosives have different absorbance characteristics over the range of 300 GHz to 3 THz. Another great feature of terahertz is that it offers high-speed data transfer and communication. It increases the data transfer speed by 6-7 times compared to microwave frequencies. However, it should be mentioned that terahertz signal is not suitable for long range communication. It

![Figure 1.3: Terahertz atmospheric absorption for horizontal transmission at sea level [6]](image)

is mainly due to its unique absorption characteristics to water. It has been seen that when
terahertz signal travels through air, the large portion of its signal gets attenuated due to absorption of water present in the air. The Figure 1.3 illustrates how significant the attenuation can be in the atmosphere. It shows the amount of absorption the terahertz signal faces per kilometer when transmits through the atmosphere at sea level for the frequency range of 10 GHz to 10 THz. It is seen that negligible amount of absorption around .01 dB/km occurs at 10 GHz. At frequencies around 100 GHz to 500 GHz, the absorption starts to rise but it is relatively lower than .5 THz - 10 THz. The absorption at 100 GHz to 500 GHz is between 1 dB to 10 dB per kilometer. But the signal absorption shows a step rise after 500 GHz. The absorption becomes significantly large around 100 thousand dB/km at 10 THz. As a consequence of this nature, it is theoretically impossible to transmit data at terahertz frequency through atmosphere to longer distance unless very powerful terahertz source is not used. Figure 1.4 provides a clear idea about the signal power requirement to establish effective communication over specific distances for frequency up to 5 THz. When distance is within 100 m, communication can be performed effectively even with very

![Figure 1.4: Transmitted power requirement over THz frequency ranges for specific distances [6]](image)
small amount of transmitted power usually less than 0 dB for up to 1 THz since signal doesn’t get absorbed significantly for travelling 100 m distance in the atmosphere. However, communication over 1 km and 6 km distance is quite impossible for frequency above 800 and 200 GHz respectively. As a result, currently researchers are considering terahertz spectrum only for short distance communication.

1.2 Scope and Motivation of the Thesis:
Though terahertz’s sensitivity to water absorption is a hindrance to the development of long-distance high-speed communication in this spectrum, the same feature could be very useful in medical imaging and diagnosing diseases. In this thesis, the absorption characteristic of Terahertz is exploited in detecting cancer. In last few decades, several experiment had been performed in determining the absorption coefficient of water in the terahertz region, sometimes in the name of the far-infrared region. Most notable research in this aspects was from M. N. Afsar and J. B. Hasted, Hans R. Zelsmann and, Ronne and Keiding [3-4, 5]. Afsar and Hasted measured the absorption coefficient of normal liquid

![Figure 1.5: Absorption of normal water vs. wavenumber at 19 degree Celsius [5]](image)
water $H_2O$ and heavy water $D_2O$ in the frequency range from 200 GHz to 13.5 THz corresponding to wavenumber 6 cm$^{-1}$ and 450 cm$^{-1}$ [5]. Figure 1.5 illustrates the absorption of normal liquid water across 200 GHz to 13.5 THz at 19 degree Celsius. From the Figure 1.5, it is seen that the absorption spectrum of water exhibits a peak at around 5.6 THz corresponding to the wavenumber 185 cm$^{-1}$ which is due to the resonant stretching of hydrogen bonds between water molecules. This absorption characteristics has been extended to lower region of the mm-wave spectrum. This absorption to water at lower region of THz spectrum has been exploited in this study to distinguish cancer from healthy tissue. From numerous study for past several decades, it has been found that cancer cell contains different amount of water than the normal healthy tissue. One of the first of this kind study was performed by Ross and Gordon [7]. They measured the water content of the tumor bearing tissue and healthy tissue of a rat liver with the use of nuclear magnetic resonance. It was found that tumor cell had 5 percent more water in cytoplasm then normal healthy tissue from their study. In addition to Ross and Gordon study, several other studies have also demonstrated the variation of water content in cancer tissues from healthy one [7-14]. Most of the studies have suggested cancer tissues including breasts and lungs contain more water than healthy counterpart. However, a study from Ali J.H. shows cancer tissues actually having less water content than the healthy sample. Ali J.H. performed measurement on cancer and healthy human prostate sample of same subject using near infrared spectroscopy [8].

Since the amount of water is different between neoplastic and normal tissue at terahertz spectrum, representing the water absorption by an appropriate parameter such as dielectric properties, malignant cancer tissue can be identified from the healthy tissue. In this thesis,
the theory of the dielectric properties, its determination techniques are discussed and then an effective method is chosen to conduct the measurement in terahertz spectrum.

1.3 Novelty of the Proposed Work

The study proposes a novel way to diagnose cancer from human body. The cancer cells have different hydration level than the healthy tissue. Using the transmission and reflection approach and the sensitivity of terahertz absorption to water, the degree of hydration is determined for cancer and healthy tissue in order to identify the cancer. The use of terahertz frequency to screen cancer is safe for human body since terahertz photons are not energetic enough to break chemical bonds or ionize atoms and molecules of the human body.

This study is considered a novel approach in detecting cancer and it is expected to perform screening and detection of cancer from human body without in need of biopsy or other invasive method.
CHAPTER 2

DIELECTRIC PROPERTIES AND MEASUREMENT THEORY

For past few decades, the use of RF and Microwave materials in various fields such as communication technology, electronics, aviation, military and biomedical applications has been significantly increased. Though, the study of material science is an edge old field of science, the increasing requirement of understanding the properties of materials functioning at the RF and microwave frequency for the development of high-speed circuit and system has brought researchers to lay emphasis on this field. This chapter discusses this background theory of material characterization; the fundamental physics that governs the interaction between materials and electromagnetic waves.

2.1 Definition of Dielectric Properties

Whenever an electric field is applied to a particular material, it induces electric polarization within the material. If the electric field is increased, the polarization in the material also increases [15]. This increment is linear and is expressed by a constant called permittivity, ε. In other words, Permittivity can be defined by the ability of a material to polarize in response to the field or the resistance that is encountered when forming an electric field in a medium. Usually in theory of electromagnetics, when permittivity of a material is discussed, it points to relative permittivity. Relative permittivity is the ratio of material’s permittivity to the permittivity of vacuum which is expresses by

\[ \varepsilon_r = \frac{\varepsilon}{\varepsilon_0} \]  

(2.1)
The permittivity of vacuum is constant in nature and it is \( \varepsilon_0 = 8.854 \times 10^{-12} \). Relative permittivity is also called as dielectric constant. The above equation is true when the source is direct current source. But for time varying electric field, an additional loss term is required and is represented by complex number. Then the permittivity representation becomes

\[
\varepsilon = \varepsilon' - j\varepsilon''
\]  

(2.2)

where, \( \varepsilon' \) and \( \varepsilon'' \) are the real and imaginary part of the permittivity respectively. The real part of permittivity expresses how much energy is stored in the material for the applied electric field and the imaginary part implies how much energy is absorbed in the material. It is often referred to as ‘loss factor’. Another term which is used to express a materials level of absorption is ‘loss tangent’ which is the ratio of imaginary permittivity to real permittivity.

\[
\tan\delta = \frac{\varepsilon''}{\varepsilon'}
\]  

(2.3)

Higher loss tangent suggests higher absorption of energy for the applied electric field. If a material possesses conduction characteristics due to the applied field, the conduction term is defined by a quantity \( \sigma \).

\[
\sigma = \omega\varepsilon_0\varepsilon''
\]  

(2.4)

where \( \omega \) is angular frequency, \( \omega = 2\pi f \)

Permeability is the response of magnetic materials under the influence of magnetic field. Due to electric and magnetic field in microwave and terahertz, magnetic materials can strongly interact with the incident radiation. Magnetic permeability is a constant that relates
the magnetic flux density to the magnetic field, which depends on intrinsic material properties such as magnetic moment and domain magnetization. Magnetic permeability, \( \mu_0 \), is a constant that has a value of \( 4 \pi \times 10^{-7} \) H/m. Under the oscillating electric field, similar to permittivity, permeability is a complex number which can be represented by

\[
\mu = \mu' - j\mu''
\]  

(2.5)

where, \( \mu' \) and \( \mu'' \) are the real and imaginary part of the permeability respectively. Usually, most of the materials have little or no magnetic response. It means their magnetic property has the same value as that of free space. This can be explained in quantum mechanic. The magnetic moment of the material is associated with the spin of the electron. It creates small magnetic dipole moment around each electron. Electron always pairs up when they share orbit shells. If two electron spins in opposite direction, their net dipole moment becomes zero because each dipole moment is in opposite direction. However, in some substrate, there could be unpaired electron in the shell. This class of material tends to show permeability [15]. These materials are well known as a ferromagnetic material.

2.2 Polarization of Dielectrics

Unlike conductor, charges of dielectric materials are bound with the molecules and atoms and therefore doesn’t move over macroscopic distance under the influence of the applied electric field. Instead these electron clouds associated with the molecules and atoms gets distorted, realigned or displaced. This results in electric dipole and materials becoming polarized. The polarizability differs and depends on the type of dielectric substrate.
2.2.1 Electric Polarization

When a dielectric material of nonpolar molecules is exposed to the external electric field, the electrons move away from the nucleus by the amount of electric field applied to it. As a result, small electric dipole forms which make themselves align according to the electric field as shown in Figure 3.1. When the external field is removed, the electron comes back to their original state.

![Figure 2.1: (a) Atoms when no external electric field applied (b) Electric dipoles induced as the externally applied electric field distorts the electron cloud [16].](image)

The phenomena of shifting electron from applied electric field can be compared to the stretched spring. A spring only stretch the amount from its ideal position, the amount of force applied to it. After removal of the force, it returns to its original state [16, 17].

The electric dipole which is created due to the external field, is a measure of the polarity of the electric charges. This polarity is directly proportional to the electric field. If ‘P’ vector is the polarization vector, ‘E’ is the applied electric field, then
\[ P = \alpha_e E_r \]  \hspace{1cm} (2.6)

where \( \alpha_e \) is called electronic polarizability of the material and it can be expressed as \( \alpha_e = 4\pi\varepsilon_0 r_e^3 \). Here \( r_e \) is the radius of the electron cloud.

### 2.2.2 Orientational (Dipole) Polarization:

This type of polarization can be seen in materials which has polar molecules in it. For instance, liquid water. The substance such as water possesses permanent dipole moment due to the polar molecules. Under ideal condition, all molecules follow random orientation due to thermal agitation. Hence, they don’t show net polarization as shown in the Figure 2.2.

![Figure 2.2](image)

**Figure 2.2:** (a) Polar molecules with no applied electric field, random orientation due to thermal agitation (b) applied field makes the polar molecules aligned [16]

When the material is exposed to an electric field, the polar molecules tend to orient themselves with the applied electric field. This results materials to achieve net polarizability in the material. This type of polarizability is termed as orientational polarizability [16, 17].
2.2.3 Ionic (Molecular) Polarization:

Ionic polarization considers polarization of the dissolved ionic substance. It’s not ideal dielectrics. For instance, NaCl dissolved in water exists as individual Na$^+$ and Cl$^-$ ions, not as neutral ions.

In the absence of electric field, the polar molecules of the water tend to keep the total polarity neutral by bonding with the positive Na$^+$ ions. When an external electric field is applied to the NaCl solution, the polar water molecules dissociate from the Na$^+$ and Cl$^-$ ions and align themselves according to the applied electric field creating a net polarizability in the solution as demonstrated in Figure 2.3 [16,17]. This type of polarizability is known as ionic polarizability and denoted by $\alpha_i$.

2.2.4 Frequency Dependence of Dielectric Mechanism

Usually, permittivity of a material decreases with the increase of frequency. The relationship between materials permittivity and frequency is demonstrated in Figure 2.4. From Figure 2.4, it can be seen that every dielectric mechanism has its own cutoff
frequency, $f_c$. At the cutoff frequency, the real part of the permittivity rolls off whereas loss factor, imaginary part of the permittivity becomes largest. The material with dipolar characteristics has lower cutoff frequency than material with atomic or electronic mechanism characteristics. Water molecule is polar molecules, hence it is dipole in nature. As a result, water like other material with dipolar mechanism characteristics has sharp roll off and cutoff frequency around 22 GHz. Materials tend to show dielectric mechanism behavior only when an AC signal is applied to those. Since, AC signal changes its direction when it propagates, when an AC signal is applied to the material, the material’s molecules realign every half cycle according to the applied electric field. For instance: polar molecule with positive charge will orient to the negative cycle of AC signal and negative charge molecule will be aligned by the positive AC cycle. Depending

![Figure 2.4: Frequency dependence dielectric behavior [17]](image)

on the dielectric mechanism, the behavior of the atoms/molecules/charge could be different as discussed in sections 2.2.1 to 2.2.3. Below the cutoff frequency, materials structure maintains realigning respective to the electric field but above the cutoff frequency it fails to do so. As a result, energy absorption decreases.
2.3 Dielectric Measurement Complication Factor

Measurement and characterization of the dielectric material is complicated. It gets even more complicated if the material is engineered or material behavior changes over frequency ranges. There are two complicated mechanism that governs dielectric materials [15].

- Dispersion
- Anisotropy

2.3.1 Dispersion

Dispersion is the property of a material that exhibits frequency dependent behavior [15]. Real part undergoes a step decrease while imaginary increases. These changes are called dispersion. There are several models to describe the dispersion. These include Drude, Debye and Lorentz model [18-20]. Drude model best fits for metal and other conductive model. Lorentz model is suitable for expressing magnetic relaxation and used for magnetic material. Debye model is used for dielectric materials. These three models are expressed by the equation in the below table.
Table 2.1: Summary of classical dispersion theories and their application [15, 18-20]

<table>
<thead>
<tr>
<th>Relaxation Model</th>
<th>Expression</th>
<th>Application</th>
</tr>
</thead>
<tbody>
<tr>
<td>Debye</td>
<td>[\epsilon = \epsilon_U + \frac{\epsilon_R - \epsilon_U}{1 + j\omega\tau}]</td>
<td>Dielectric materials</td>
</tr>
<tr>
<td>Lorentz</td>
<td>[\epsilon = \epsilon_U - \frac{\omega_0^2}{\omega_0^2 - \omega^2 + 2j\omega\delta}]</td>
<td>Magnetic materials</td>
</tr>
<tr>
<td>Drude</td>
<td>[\epsilon = \epsilon_U - \frac{\omega_p^2}{\omega^2 - j\omega\delta}]</td>
<td>Conductive materials</td>
</tr>
</tbody>
</table>

For the expression in table, \(\epsilon_R\) and \(\epsilon_U\) are the high and low frequency limits of the permittivity, \(\tau\) is the characteristics relaxation time, \(\omega_o\) is the relaxation frequency and \(\omega_p\) is the plasma frequency. Note that, though these equations are specified for the permittivity, it can be applied to magnetic permeability as well.

The dielectric constant achieved from this model often may not be accurate. It has been seen that relaxation phenomena in many materials occur over a broad range of frequency than these models assume. To accommodate relaxation phenomena over wider bandwidth,
Cole and Cole has modified the Debye model by adding one additional term, $\alpha$ [21]. The expression is

$$\varepsilon = \varepsilon_U + \frac{\varepsilon_R - \varepsilon_U}{1 + (j\omega \tau)^{1-\alpha}}$$

(2.7)

Here, the $\alpha$ term extends the relaxation to the much wider bandwidth.

Besides modifying the Debye model in order to express the relaxation characteristics in a broader frequency range of certain material, the model requires another modification to express real and imaginary permittivity for the material with unbound charges. Figure 2.5 shows the real and imaginary permittivity of the polymid/graphite. The charges of this material are bound to the molecules and dipole polarizability is the primary mechanism for dispersion. As it is less lossy, the real and imaginary permittivity is low. On the other hand, Figure 2.6 shows the real and imaginary permittivity of carbon loaded foam of which charges are not bound to the molecules. As a result, it demonstrates conductivity.

Therefore, for Debye model to show accurate real and imaginary permittivity of a material similar to this with unbound charges, a conduction term, $\sigma$, is required. With the addition of $\sigma$ term, the Debye model becomes

$$\varepsilon = \varepsilon_U + \frac{\varepsilon_R - \varepsilon_U}{1 + j\omega \tau} + j\frac{\sigma}{\omega \varepsilon_0}$$

(2.8)

where $\sigma$ is the DC conductivity term and $\varepsilon_0$ is the permittivity of the free space.
Table 2.2: Debye formula for materials with bound and unbound material

<table>
<thead>
<tr>
<th>Debye model for bound change material</th>
<th>Debye model for unbound change material</th>
</tr>
</thead>
<tbody>
<tr>
<td>[ \varepsilon = \varepsilon_U + \frac{\varepsilon_R - \varepsilon_U}{1 + j\omega\tau} ]</td>
<td>[ \varepsilon = \varepsilon_U + \frac{\varepsilon_R - \varepsilon_U}{1 + j\omega\tau} + j\frac{\sigma}{\omega\varepsilon_0} ]</td>
</tr>
</tbody>
</table>

In Figure 2.6, real and imaginary permittivity are plotted for simple Debye model, Cole-Cole model, Debye model with conductivity term and compared with the measured data.

Figure 2.5: Real and imaginary permittivity comparison of bound charge material (Graphite) for Cole-Cole and Debye models with the measured data [1]
Figure 2.6: Real and imaginary permittivity comparison of unbound charge material (Carbon foam) for Cole-Cole, Debye and Debye-conductivity models with the measured data [15]

It can be seen that Cole-Cole model fits better than simple Debye model with measured data. However, Debye model with conductivity fits best with the actual data.

2.3.2 Anisotropy

Anisotropy means the material which have multiple dielectric characteristics depending on the orientation. Examples include Honeycomb, Fiber reinforced composites and flake field materials. These kind of materials will give complete different dielectric constant depending on how it is placed on the sample holder during material characterization.
2.4 Complex Permittivity Measurement

Calculation of permittivity requires a signal to be transmitted through the material which permittivity will be measured. It considers the signal’s electromagnetic interaction with the material. When a signal passes through the material, its insertion loss, return loss and phase changes. It generates new insertion loss and return loss each time it interacts with the new medium. Therefore, calculation of permittivity differs for solid than liquid or compound material.

2.4.1 Complex Permittivity measurement of solids

2.4.1.1 Single Layer Medium

Consider an electromagnetic wave is applied to the solid dielectric medium of a material with the thickness ‘d’ as shown in Figure 2.8. Some amount of the wave goes enters the
medium 2 while some reflects back. Inside the medium 2, some portion of the wave again reflects back and some part passage reach medium 3. Thus, while changing medium, the EM wave generates multiple transmission and reflection at boundaries of each the two different mediums. All of the transmission and reflection quantity result transmission coefficient and reflection coefficient, $S_{21}$ and $S_{11}$, respectively. The propagation constant for the medium can be expressed by

$$\gamma = \mu_0 \varepsilon_0 \sqrt{\varepsilon_r \frac{\sigma}{\omega\varepsilon_0}}$$  \hspace{1cm} (2.9a)

$$\gamma = \frac{j\omega \sqrt{\varepsilon_r - j\varepsilon_r''}}{c}$$  \hspace{1cm} (2.9b)

If the wave is incident from medium 2 (dielectric) from medium 1 (air), then the reflection coefficient would be

$$\Gamma_{12} = \frac{Z_0 - Z_1}{Z_1 + Z_0}$$  \hspace{1cm} (2.10)

where $Z_1 = Z_0 / \sqrt{\varepsilon_r}$, impedance of the dielectric medium and $Z_0 = 120\pi$ is the impedance of the air.
With the use of Ray-Tracing model, the total reflection coefficient at the input would be

\[
\Gamma_{in} = \Gamma_{12} + T_{12} T_{21} \Gamma_{21} e^{-2j\beta d} + T_{12} T_{21} \Gamma_{21}^3 e^{-4j\beta d} + \cdots \quad (2.11a)
\]

\[
= \Gamma_{12} + \frac{T_{21} \Gamma_{21} e^{-2j\beta d}}{1 - \Gamma_{21}^2 e^{-2j\beta d}} \quad (2.11b)
\]

where, \( \beta \) is the propagation constant, \( d \) is the thickness of the slab, \( \Gamma_{12} \) is the return loss at the boundary of air-dielectric slab for the reflected signal of dielectric slab-air boundary from inside the slab, \( \Gamma_{21} \) is the return loss while the EM wave first interacts with the air-dielectric slab.

\[
\beta d = \frac{2\pi d}{\lambda} = \frac{2\pi d \sqrt{\mu_r \varepsilon_r}}{\lambda_0} \quad (2.12)
\]

Here, \( \lambda \) is the wavelength of the medium, \( \lambda_0 \) is the free space wavelength, \( \varepsilon_r \) and \( \mu_r \) are the relative permittivity and permeability of the dielectric medium respectively. It should also be mentioned that \( T_{12} = 1 + \Gamma_{12} \), \( T_{21} = 1 + \Gamma_{21} \) and \( \Gamma_{21} = -\Gamma_{12} \). Now, replacing \( \beta d \) in the equation 2.11b with the 2.12 and arranging, \( \Gamma_{in} \) can be found.

\[
\Gamma_{in} = S_{11} = \frac{\left(1 - e^{-2j\omega d / c} \sqrt{\varepsilon_r - j\varepsilon_r''} / \varepsilon_r''ight)}{\left(1 - \Gamma_{12}^2 e^{-2j\omega d / c} \sqrt{\varepsilon_r - j\varepsilon_r''} / \varepsilon_r''ight)} \Gamma_{12} \quad (2.13)
\]

Equation 2.13 provides the total reflection coefficient of the dielectric slab for the incident EM wave. Similarly, the total transmission coefficient, \( S_{21} \) is evaluated by

\[
S_{21} = \frac{(1 - \Gamma_{12} e^{-2j\omega d / c} \sqrt{\varepsilon_r - j\varepsilon_r''} / \varepsilon_r'')}{1 - \Gamma_{12}^2 e^{-2j\omega d / c} \sqrt{\varepsilon_r - j\varepsilon_r''} / \varepsilon_r''} \quad (2.14)
\]
In equation 2.13 and 2.14, \( \sqrt{\varepsilon' - j\varepsilon''} \) can be written as \( \sqrt{\varepsilon_r} \) where \( \varepsilon_r \) is the complex dielectric constant. \( \sqrt{\varepsilon' - j\varepsilon''} \) value can be found by solving 2.13 and 2.14 independently. Therefore, \( \varepsilon_r \) can be determined.

### 2.4.2 Compound Materials

If two or more dielectric materials are combined together, it is called compound material. The property ‘P’ of the two dielectric combined together for its non-iterative and distributive mixture is

\[
P = v_1 P_1 + v_2 P_2
\]  

(2.15)

where \( v_1 \) and \( v_2 \) are the volume fraction of those two dielectrics. As two dielectric components are combined together and each has its own dielectric constant, therefore an effective dielectric constant is evaluated for the compound materials [12].

\[
\varepsilon_{r\text{ eff}} = v_1 \varepsilon_r + \varepsilon_r V_2
\]  

(2.10a)

\[
\varepsilon_{r\text{ eff}} = \frac{v_1 \varepsilon_r V_1 + \varepsilon_r V_2}{V_1 + V_2}
\]  

(2.16b)

here, \( \varepsilon_{r1} \) and \( \varepsilon_{r2} \) are the dielectric constant and \( V_1 \) and \( V_2 \) are the volume of the two material respectively. If the two components have same area of cross-section, then the

\[\text{Figure 2.9: Configuration of compound materials with two components [22]}\]
volume fractions can be replaced with the thickness of the components. If \( t_1 \) and \( t_2 \) are the thickness of the two samples shown in Figure 2.9, then effective dielectric constant become

\[
\varepsilon_{\text{eff}} = \frac{t_1 \varepsilon_{\text{r}1} + t_2 \varepsilon_{\text{r}2}}{t_1 + t_2}
\]

(2.17)

2.4.3 Complex Permittivity Measurement of Liquids

While determining the dielectric constant of any liquid, an electromagnetic wave goes through five different medium. It is because measurement of any liquid requires container to hold the liquid. As a result, if container is made of glass then EM wave first interacts at the boundary of air-glass before reaching into liquid. The calculation of the permittivity becomes complicated when it involves five medium. Most of the liquid are considered lossy, therefore additional loss term should be considered as well.

From Figure 2.10, consider the thickness of the glass and the liquid are \( d_1 \) and \( d_2 \) respectively. As the medium 1 is simply air, as a result

Impedance of the medium 1 (air), \( \eta_1 = 120\pi = 377 \) \( \Omega \)

Impedance of the medium 2 (glass), \( \eta_2 = \eta_1 / \sqrt{\varepsilon_{\text{glass}}} \) \( \Omega \)

Impedance of the medium 2 (liquid), \( \eta_3 = \eta_3 / \sqrt{\varepsilon_{\text{liquid}}} \) \( \Omega \)
Figure 2.10: EM wave propagation through five medium to measure liquid’s dielectric constant [22]

Reflection and transmission coefficient of the different boundaries of the medium are calculated using the impedance of air, glass and liquid [23]. Transmission coefficient for air-glass, glass-liquid, liquid-glass and glass-air are $T_{21}, T_{32}, T_{43}$ and $T_{54}$ respectively.

\[ T_{21} = \frac{2\eta_2}{\eta_1 + \eta_2} \quad (2.11) \]
\[ T_{32} = \frac{2\eta_3}{\eta_2 + \eta_3} \quad (2.18b) \]
\[ T_{43} = \frac{2\eta_4}{\eta_3 + \eta_4} \quad (2.18c) \]
\[ T_{54} = \frac{2\eta_5}{\eta_4 + \eta_5} \quad (2.18d) \]

Likewise, the reflection coefficient for the mentioned interfaces would be –
\[ \Gamma_{21} = \frac{\eta_1 - \eta_2}{\eta_1 + \eta_2} \quad (2.19a) \]

\[ \Gamma_{32} = \frac{\eta_2 - \eta_3}{\eta_2 + \eta_3} \quad (2.12) \]

\[ \Gamma_{43} = \frac{\eta_3 - \eta_4}{\eta_3 + \eta_4} \quad (2.13c) \]

\[ \Gamma_{54} = \frac{\eta_4 - \eta_5}{\eta_4 + \eta_5} \quad (2.14) \]

When an EM wave passes through the medium, attenuation and phase change occurs depending on the medium. For the measurement of liquid, phase change happens at medium 2, medium 3 and medium 4 and change of phase are denoted by \( \theta_1 \), \( \theta_2 \) and \( \theta_3 \). In this complex permittivity calculation of liquid, liquid is considered as lossy and attenuation factor is accounted for this lossy behavior.

\[ \theta_1 = \gamma_1 d1 = (\alpha_1 + j\beta_1) d1 \quad (2.20) \]

\[ \theta_2 = \gamma_2 d2 = (\alpha_2 + j\beta_2) d2 \quad (2.21) \]

\[ \theta_3 = \gamma_3 d3 = (\alpha_3 + j\beta_3) d1 \quad (2.22) \]

The above equation has a specific term called \( \gamma \) which can be defined as \( \gamma = \alpha + j\beta \) where \( \alpha \) and \( \beta \) represents the attenuation constant and \( \beta \) represents the phase constant, respectively [23]. These two constants can be defined by the following equation below –

\[ \alpha = \frac{\omega \sqrt{\mu \varepsilon}}{\sqrt{2}} \sqrt{\left(1 + \left(\frac{\sigma}{\omega \varepsilon}\right)^2\right)} - 1 \quad (2.23) \]

\[ \beta = \frac{\omega \sqrt{\mu \varepsilon}}{\sqrt{2}} \sqrt{\left(1 + \left(\frac{\sigma}{\omega \varepsilon}\right)^2\right)} + 1 \quad (2.24) \]
In order to calculate the total transmission coefficient $S_{\text{total}}$ of the system, multiple reflections at each medium and their effect at each boundary need to be evaluated. Therefore, a new operator ‘O’ is defined. This operator takes multiple transmissions between dielectric interfaces into account and express the remaining output signal at any interface [22].

The EM wave coming out from the DI$_4$ due to the signal entering the interface DI$_1$ is denoted by $O_1$ and given by

$$O_1 = T_{54} T_{43} T_{32} e^{-j(\theta_1 + \theta_2 + \theta_3)} \quad (2.25)$$

Similarly for the signal emerging from DI$_2$ and DI$_3$ can be expressed as

$$O_2 = T_{54} T_{43} e^{-j(\theta_2 + \theta_3)} \quad (2.26)$$

$$O_3 = T_{54} e^{-j(\theta_3)} \quad (2.27)$$

Now, consider EM wave just enters the medium 2. The amplitude of the wave would be $T_{21}$ at the interface of DI$_1$. The effect of this wave at the output is $O_1 T_{21}$. A part of the $T_{21}$ will be reflected back from DI$_2$ to DI$_1$. The reflected signal again will be reflecting and the term is defined by $\Gamma_{23} \Gamma_{21} T_{21} e^{-j(2\theta_1)}$. Defining all the reflections inside the medium 2, a series term is obtained

$$\text{Set 1} = O_1 T_{21} \left[ 1 + \Gamma_{23} \Gamma_{21} e^{-j(2\theta_1)} + \Gamma_{23}^2 \Gamma_{21}^2 e^{-j(4\theta_1)} + \cdots \right] \quad (2.28a)$$

$$= O_1 T_{21} \frac{1}{1 - \Gamma_{21} \Gamma_{23} e^{-j(2\theta_1)}} \quad (2.28b)$$
Multiple reflections at the medium third and their effect at the output is given by

\[
\text{Set 2} = T_{54} T_{43} e^{j(\theta_2 + \theta_3)} \times T_{21} T_{32} \Gamma_{34} \Gamma_{32} e^{-j(\theta_3 + \theta_2)} \left[ 1 + \Gamma_{34} \Gamma_{32} e^{-j(2\theta_2)} + \Gamma_{34}^2 \Gamma_{32}^2 e^{-j(4\theta_2)} + \ldots \right]
\]

\[= O_2 T_{21} T_{32} \Gamma_{34} \Gamma_{32} e^{-j(\theta_1 + \theta_2)} \frac{1}{1 - \Gamma_{34} \Gamma_{32} e^{-j(2\theta_2)}} \quad (2.29a)\]

Multiple reflections and their effect at fourth medium can be expressed

\[
\text{Set 3} = O_3 T_{43} \Gamma_{45} \Gamma_{43} T_{32} T_{21} e^{-j(\theta_1 + \theta_2 + \theta_3)} \frac{1}{1 - \Gamma_{45} \Gamma_{43} e^{-j(2\theta_3)}} \quad (2.30)
\]

Similar procedure can be used at the following terms to obtain the output power

\[
\text{Set 4} = O_2 T_{32} \Gamma_{21} \Gamma_{23} T_{21} e^{-j(3\theta_1)} \frac{1}{1 - \Gamma_{34} \Gamma_{32} e^{-j(2\theta_2)}} \quad (2.31)
\]

\[
\text{Set 5} = O_2 T_{32} \Gamma_{21}^2 \Gamma_{23} T_{21} e^{-j(5\theta_1)} \frac{1}{1 - \Gamma_{34} \Gamma_{32} e^{-j(2\theta_2)}} \quad (2.32)
\]

\[
\text{Set 6} = O_2 T_{32} \Gamma_{21}^2 \Gamma_{23}^2 T_{21} e^{-j(7\theta_1)} \frac{1}{1 - \Gamma_{34} \Gamma_{32} e^{-j(2\theta_2)}} \quad (2.34)
\]

\[
\text{Set 7} = O_2 T_{32} T_{23} \Gamma_{34} \Gamma_{21} T_{21} e^{-j(2\theta_1 + 2\theta_3)} \frac{1}{1 - \Gamma_{21} \Gamma_{23} e^{-j(2\theta_1)}} \quad (2.35)
\]

There will be infinite number of reflected signal inside the dielectric medium. The terms which contributes less than one percent is neglected from the computation. Therefore, the total transmission coefficient becomes

\[S_{\text{total}} = \text{Set 1} + \text{Set 2} + \text{Set 3} + \text{Set 4} + \text{Set 5} + \text{Set 6} + \text{Set 7} + \text{Set 8} + \text{Set 9} + \text{Set 10} \quad (2.36)\]

The \(S_{\text{total}}\) is the transmission coefficient of the liquid when Electromagnetic waves pass through it. And the measurement of this coefficient is obtained by network analyzer.
In the discussion above, a theoretical approach is taken to determine the transmission and reflection coefficient of the liquid and solid materials. In practice, there are measurement methods exist which is used to measure the coefficients and subsequently the coefficients are converted into permittivity and permeability by algorithms. Following chapter will discuss about the measurement methods associated with determining these dielectric properties.
CHAPTER 3

METHODS FOR MATERIAL CHARACTERIZATION

For over a century, there has been a long history of research in material characterization for over a century. Some of the earliest known research was conducted by J.C Bose at 1890 [15]. He performed research on polarization of the electromagnetic waves with the use of technique similar to free space measurement system. His research work was groundbreaking. This field drew attention of the scientific community during World War II when understanding of the material properties of RF and microwave frequency was crucial for the development of radar, aircraft and electronic communication devices [15].

During in the last few decades many new measurements methods have been invented and applied. This chapter discusses with the measurement methods which can be applied to characterize the material.

Several measurement methods exist to characterize the material but selecting the proper measurement method can be often difficult. Four methods are discussed in this chapter. These are coaxial probe, transmission line, free space, and resonant cavity techniques. The coaxial probe technique is usually suitable for measuring liquids, transmission line technique is best fit for solids. Free space technique is useful for measuring sample with high temperature and in high frequency regime. The resonant cavity should be used to measure low loss sample and the measurement where highest accuracy is desirable. In the following few sections, these four techniques have been explained. Measurement steps, procedures, their application, and limitations are discussed.
3.1 Measurement Techniques

3.1.1 Coaxial Probe Method

The coaxial probe method is one of the basic and widely used methods of all the measurement techniques. The measurement setup of this method consists of a vector network analyzer and an open-ended coaxial probe. To perform the characterization of a material with this technique for liquid or semi-liquid material, the tip of the probe is immersed into the material as demonstrated in Figure 3.2. The probe is pressed against the surface of solid if the material under test (SUT) is solid. Since the impedance of the sample under test (SUT) is different from the impedance of the transmission line, when electromagnetic waves from VNA goes through the transmission line and incident onto the material at the interface, the portion of the wave reflects back from sample in Figure 3.1. The amount of reflection of the incident signal is expressed using scattering parameter $S_{11}$.

![Figure 3.1: Open circuit reflection of Open-ended coaxial probe technique [24]](image)

Scattering parameter explains the relationship between ports in an electrical system [25]. In a two port system with port 1 and port 2 being transmitting antenna and receiving antenna, $S_{21}$ represents power transferred from port 1 to port 2 and $S_{11}$ represents how much power reflects back to port 1. This scattering parameters are used to determine permittivity and permeability. As the impedance of the material varies one from the other, the reflection
Coefficient also becomes different for each material. Thus, when permittivity is obtained from the reflection coefficient, it shows different permittivity value for each material. Agilent Technologies provide a special material measurement software namely 85071E which calculates the permittivity and permeability values automatically from the VNA measured transmission and reflection coefficient data. For the coaxial probe method, reflection coefficient is only measured [17, 24]. Hence, the software only considers reflection coefficient to provide permittivity.

The coaxial probe method works for the frequency range from .2 GHz to 50 GHz depending on the probe type. The different types of probes are shown in Figure 3.3.
Figure 3.3: Three different coaxial probe (a) high-temperature probe (b) slim probe (c) performance probe [26]

3.1.1.1 High-Performance Probe

The tip of the high-temperature probe is sealed by hermetic glass to metal and it can survive in abrasive and corrosive substances. This probe can withstand between -40 to 200 degree Celsius. The large flange of this probe allows it to measure the flat surfaced solid material in addition to liquid and semi-liquid as shown in Figure 3.4. With this probe, the measurement can be performed from 200 MHz to 20 GHz.

Figure 3.4: Dimension of high-performance probe [26]

3.1.1.2 Slim probes

Due to its smaller diameter, the slim probe is suitable for measurement of small sample. A slim probe with its dimension shown in Figure 3.5. It is also used in the tight space.
However, it is not recommended to use this probes for measuring solids because of the small tip size. The slim probes can operate from 500 MHz to 50 GHz. The slim probes easy to build and cheap, which makes it favorable for measuring the materials that would destroy the probe such as epoxy or cement.

3.1.1.3 Performance Probe
The performance probe combines the features of the high temperature probe and slim probe which means it withstand extreme temperature typically from -40 to 200 degree Celsius and can operate from 500 MHz to 50 GHz such as slim probes [17]. This probe is sealed on both probe tip and connector end allowing it to be autoclaved, so it is suitable for food and medical industry where sterilization is important. The dimension of the probe is shown in the Figure 3.6.
3.1.1.4 Strength and Limitation

Coaxial probe technique is easy to use and does not require a delicate setup for the measurement. While performing the measurement, it doesn’t damage the sample which makes it non-destructive. There are several factors which should be maintained in order to have an error free measurement from this method. During liquid material characterization, the probe to liquid contact should be such that no air bubble form around the tip. Same condition applies for solid and semi-solid measurement as well so that no air gap exists at the contact. Sample should be homogeneous and isotropic. Sample should also be thick enough usually 1 cm so that electromagnetic waves from VNA sees the medium as infinite [17, 26].

3.1.2 Transmission Line Method

In the transmission line method, the sample under test in inserted into a small portion of the transmission line. Usually coaxial cable or waveguide is used as transmission line in this technique. To establish the setup, the two port of a calibrated VNA is connected to

![Figure 3.7: A simple transmission line setup [17]](image-url)
the section of the coaxial cable or waveguide as demonstrated in Figure 3.7. Sample under test (SUT) is placed into the cable or waveguide and electromagnetic wave is applied to the SUT through the sample. The reflection and transmission amount of the incident wave is measured through the VNA. These transmission and reflection coefficient is then converted into dielectric parameters by Agilent 85071E material measurement software. The converted dielectric parameter can be observed from the software directly installed into the VNA or by installing the software in an external computer and then communicating it to the VNA by LAN or GPIB interface. In the coaxial cable method, only reflection coefficient is used to extract the dielectric properties. In the transmission line technique, all four scattering parameters $S_{11}$, $S_{22}$, $S_{21}$ and $S_{12}$ are used [17, 24, 27]. Hence, it is a full two port measurement.

3.1.2.1 Condition of Transmission Line Measurement

Since the transmission line measurement requires sample to put into the transmission line, the sample under test needs to be shaped into certain configuration before placing it into the sample holder. For coaxial cable method, the sample may need to be shaped into toroidal form. For the waveguide, sample length and width should be similar to the inner

![Figure 3. 8: Coaxial 7mm airline and X-band waveguide with samples [17]](image)

Figure 3. 8: Coaxial 7mm airline and X-band waveguide with samples [17]
dimension of the waveguide as shown in Figure 3.8. As a result, this technique is best suited when the material under test is solid. Liquid and powder samples can also be measured but requires dielectric dam to be placed into the sample holder. The reason why samples need to be completely filled in shape in the inner cross section of the cable and waveguide is to avoid air gaps around the sample. If sample doesn’t fit properly and leaves air gaps, then acquired data will have inaccuracy. The sample also needs to be flat faced and uniform throughout. The sample should be thick enough to contain at least portion of one wavelength. Transmission line technique works from .1 GHz to 110 GHz. Use of coaxial cable as transmission line allows measurement at broad range of frequency whereas waveguide offers banded frequency coverage. As waveguide dimension gets smaller with the higher frequency, the sample requires to be smaller in order to fit inside the sample holder. This method provides more accurate data than probe method but comes with few limitation as well. The strength and limitation are shown in the Table 3.1.
Table 3.1: Advantages and limitations of the transmission line method [17, 28]

<table>
<thead>
<tr>
<th>Aspects</th>
<th>Advantage</th>
<th>Limitation</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sample configuration</td>
<td>Can measure anisotropic and magnetic material</td>
<td>Liquid, powder or gasses are hard to measure</td>
</tr>
<tr>
<td>Frequency coverage</td>
<td>Broad frequency of operation range</td>
<td>Sample shaping required and certain thickness should be maintained</td>
</tr>
<tr>
<td>Measurement accuracy</td>
<td>More accurate measurement with minimum error</td>
<td>Sample become large at low frequency</td>
</tr>
<tr>
<td></td>
<td>Ideal for solid material</td>
<td>Not recommended for extreme low loss material</td>
</tr>
</tbody>
</table>

3.1.3 Free space Measurement Method

Free-space measurement can be considered as the other form of transmission line method. In transmission line technique, cable and waveguide are used as the transmission medium of the electromagnetic waves whereas air acts as transmission medium of energy in free space technique. Aside from that, similar to transmission line method, free space system also measures both transmission and reflection coefficient and use the same algorithm to extract the dielectric properties of the material.
3.1.3.1 Free space Measurement procedure:

The free space measurement procedure is relatively complicated than other measurement techniques. It requires precise alignment of the fixtures, otherwise measurement uncertainty arises. A typical free space measurement requires following equipment to establish the setup.

- Two antennas (horn)
- Two standoffs antenna mounting system
- Sample holder
- Metal reflector plate
- Sample
- Broadband Absorber (TRM calibration)
- Rail fixture/ Precise positioning fixture (TRL calibration)

Free-space technique usually requires only the aforementioned equipment. However, depending on what type of calibration is performed before measurement, additional equipment requirement may arise. This method follows three calibration techniques which are TRL, TRM and GRL. The main difference between TRL and TRM is that they both employ Thru and Reflect standard but TRL considers Line standard whereas TRM uses Match standard as the third standard. If either of these two calibration is performed before measurement, then a broadband absorber and precise positioning fixture is required for TRM and TRL respectively. In a typical free space setup, the port one antenna (left) acts as a transmit antenna and port two antenna (right) acts as a receiving antenna as shown in Figure 3.9. Ports 1 and port 2 antennas are connected to ports 1 and 2 of the VNA, respectively. The material under test (SUT) is placed at the middle and equal distance from
two antennas. During the measurement, electromagnetic energy is allowed to go through the sample to measure the reflected \([S_{11}, S_{22}]\) and transmitted \([S_{21}, S_{12}]\) energy of the sample via VNA.

![Diagram of a typical free space measurement setup](image)

**Figure 3.9: A typical free space measurement setup [29]**

There are several conditions that need to be met in order to establish an effective free space setup and perform measurement accurately without errors. The conditions are discussed below

- Both antennas of the setup should be of the same band. Half power beam width (HPBW) should be such that material under test (SUT) can contain 3 dB beam spot within its body [17, 28, 30]. In other words, SUT should be large enough to contain the 3 dB beam spot. If the beam spot becomes equal or larger than the sample size, then the incident waves will scatter and provide inaccurate scattering parameter data. The 3 dB beam spot is much larger for antennas operating at lower frequency band. For instance; while performing X-band free space measurement, sample dimension requires being several square feet to contain the 3 dB beam spot within the area of the sample. However, sample dimension decreases at the higher
frequency. For example, the sample area needs to be only several square inches (6 in x 6 in) for WR-10 waveguide horn antenna for the operating frequency of 75-110 GHz.

- Distance from the SUT to the antenna should be greater than $2D^2/\lambda_g$ in order to have sample far field region where $D$ is the diameter of the antenna and $\lambda_g$ is the guided wavelength of the air [30]. If the sample is located at the far field region, then waves with planar wave front can incident upon the sample under test and phase will be identical. Below is the far field distance calculation for the X-band horn antenna. The larger dimension of X-band antenna $D$, is 5.5 inches or .1397m.

Therefore, the required sample to antenna distance = $2D^2/\lambda_g = \frac{2 \times (0.1397)^2}{0.03} = 1.3$ m.

Whereas, the guided wavelength at free space is, $\lambda_g = \frac{C}{f} = \frac{3 \times 10^8}{10 \times 10^9}$. Agilent uses quasi-optical system to achieve far field at the sample. Figure 3.5 shows the free space quasi-optical system setup for W-band frequency. The difference between conventional free space setup and quasi-optical system shown

![Image](https://via.placeholder.com/150)

Figure 3. 10: Quasi-optics free space measurement system [28]
in figure 3.10 is that it has two additional mirrors. These mirrors are concave mirrors. The concave shape can be seen by looking closely at upper reflector of Figure 3.10. This pair of concave reflector turns focused beam into collimated beam. Collimated beam provides planar wave. As a result, sample is considered to be in far field when collimated beam incidents on it.

- The algorithm which is used to convert the S-parameter responses to permittivity and permeability (Nicholson Ross, NIST) requires the thickness of the sample smaller than the length of ‘one guided wavelength’ [31]. Otherwise, the algorithm generates multiple roots if the thickness of the material exceeds the length of the guided full wavelength at the operating frequency [31]. Within one full guided wavelength, the thicker the sample, better the response be in terms of dielectric properties. The sample thickness should be between $\frac{20}{360}\lambda_g$ to $\lambda_g/2$. Table 3.2 shows the calculated minimum and ideal sample thickness for the frequency from 10 GHz to 300 GHz. It can be noted that, sample thickness gradually decreases with the increase of operating frequency.

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Wavelength (mm)</th>
<th>Minimum Sample Thickness (mm)</th>
<th>Ideal Sample Thickness (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>30</td>
<td>1.67</td>
<td>15</td>
</tr>
<tr>
<td>20</td>
<td>15</td>
<td>.83</td>
<td>7.5</td>
</tr>
</tbody>
</table>
When free space calibration is performed, a metal plate is used to create ‘Reflect’ standard. The metal plate is placed at the exact location where SUT is placed. It is recommended for metal plate to have same thickness as the sample. Otherwise, adjustment should be made at the material measurement software in order to achieve accurate material properties. An ideal reflect standard reflects all of the energy which incidents up on it, nothing passes through the reflector and reach on the other side. In practice, ideal reflector can’t be found. Instead, copper or aluminum are used. To ensure energy does not pass through the reflector, skin depth of copper or aluminum depending on which one is used as reflector need to be calculated for the frequency range of operation. Below, skin depth has been calculated for Aluminum and Copper for the frequency of operation.

\[
\delta_s = \sqrt{\frac{2}{\omega \mu \sigma}} = \sqrt{\frac{\rho}{\pi f \mu}} \tag{3.1}
\]

where \(\omega\) is the radian frequency, \(\rho\) is the bulk resistivity, \(\sigma\) is the conductivity, \(\mu\) is the permeability of the material. At room temperature, the resistivities of Aluminum and

<p>| | | | |</p>
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<thead>
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<tbody>
<tr>
<td>40</td>
<td>7.5</td>
<td>.41</td>
<td>3.75</td>
</tr>
<tr>
<td>50</td>
<td>6</td>
<td>.33</td>
<td>3</td>
</tr>
<tr>
<td>100</td>
<td>3</td>
<td>.16</td>
<td>1.5</td>
</tr>
<tr>
<td>200</td>
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<td>.75</td>
</tr>
<tr>
<td>300</td>
<td>1</td>
<td>.056</td>
<td>.5</td>
</tr>
</tbody>
</table>
Copper are 2.65 µΩ.cm and 1.69 µΩ.cm respectively. For 10 GHz, 140 GHz and 215 GHz, the calculation has been performed for both Aluminum and Copper.

\[
\delta_s = \sqrt{\frac{\rho}{\pi f \mu}} = \sqrt{\frac{1.69 \times 10^{-8}}{\pi \times 10 \times 10^9 + 4 \pi \times 10^{-7}}} = 0.654 \text{ um}
\]

Table 3.3: Skin depth calculation for Aluminum

<table>
<thead>
<tr>
<th>Material</th>
<th>Frequency, f</th>
<th>Skin depth, (\delta_s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Aluminum</td>
<td>10 GHz</td>
<td>0.819 um</td>
</tr>
<tr>
<td></td>
<td>140 GHz</td>
<td>0.218 um</td>
</tr>
<tr>
<td></td>
<td>215 GHz</td>
<td>0.176 um</td>
</tr>
</tbody>
</table>

Table 3.4: Skin depth calculation for Copper

<table>
<thead>
<tr>
<th>Material</th>
<th>Frequency, f</th>
<th>Skin depth, (\delta_s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Copper</td>
<td>10 GHz</td>
<td>0.654 um</td>
</tr>
<tr>
<td></td>
<td>140 GHz</td>
<td>0.174 um</td>
</tr>
<tr>
<td></td>
<td>210 GHz</td>
<td>0.141 um</td>
</tr>
</tbody>
</table>

3.1.3.2 Condition for Accurate Free space Measurement

Free space method is a sophisticated measurement system. Measurement at low frequency involves less complexity and accuracy than higher frequency. The degree of accuracy
decreases when the measurement is performed at higher frequencies. Usually focused beam method is used at high-frequency measurement to increase the accuracy. There are several precautions which should be kept in mind while conducting free space measurement [22].

- The sample holder position and antenna stand should be fixed. Even the slight change of position can add up error, especially the phase which is very critical for accurate permittivity measurement.
- The fixtures shouldn’t be moved during calibration and measurement. If fixtures are moved after calibration, the calibration reference plane changes and the data obtained will be inaccurate.
- The metal plate which is used as reflector standard during calibration and the sample should be exactly in the same location and normal to the incident wave. Reflector/metal plate shouldn’t be bent.
- Sample size should be larger than $5\lambda$. It is the rule of thumb to ensure beam size to be within the area of the sample. It minimizes the edge diffraction.
- When TRL (Thru, reflect, line) calibration is performed, port 2 should be moved by quarter wave distance from its previous reflect standard position for line standard calibration [30].
- The use of thick sample in comparison to the wavelength of operation should be avoided. It creates ambiguity in phase values.

Even though free space system is quite delicate than another system, it has some significant advantages over other techniques. Strength and limitation of this technique are given below in the Table 3.5.
Table 3.5: Strength and limitation of the Free space technique [17, 28, 30]

<table>
<thead>
<tr>
<th>Aspects</th>
<th>Strength</th>
<th>Limitations</th>
</tr>
</thead>
<tbody>
<tr>
<td>Sample shape consideration</td>
<td>Non-destructive and non-contacting measurement of the sample</td>
<td>Samples need to be flat and parallel faced</td>
</tr>
<tr>
<td>Frequency coverage</td>
<td>Covers broad range of frequency, up to 500 GHz</td>
<td>Large sample requires for low frequencies</td>
</tr>
<tr>
<td>High temperature measurement</td>
<td>Sample can be measured at high temperature</td>
<td>Liquid, powder or gasses are hard to measure</td>
</tr>
<tr>
<td>Calibration</td>
<td>GRL calibration allows avoiding expensive fixtures</td>
<td>Small change of alignment gives significant error</td>
</tr>
</tbody>
</table>

3.1.3.3 Free-space Calibration

There are two calibration types at VNA in terms of standard. One is- SOLT and other one is TRL. VNA usually does the SOLT calibration which uses Short, Open, Load, Thru standards. Most of the calkits file inside VNA is of SOLT calibration. There are only couple of calkits which supports TRL calibration. Ex: 85052C. This calkit uses Agilent 85052C mechanical calkit to perform the TRL calibration. The free space calibration follows TRL calibration family because SOLT calibration standards are hard to achieve for free space measurement.
Calibrating the network analyzer for a free space measurement is challenging. Free space calibration standards present special problems since they are “connector-less”. There are three calibrations for free space techniques. Those three are –

1. Gated Reflect Line (GRL)
2. Thru Reflect Line (TRL)
3. Thru Reflect Match (TRM)

The free space calibration is achieved by defining the reference plane. Reference plane is the plane up to which calibration is achieved. It is usually the surface of both side of the metal plate used for reflect standard and is located at the middle having equal distance from the two antenna.

3.1.3.3.1 TRL Calibration

TRL calibration follows three standards which are Thru, Reflect and Line. When calibration is performed Thru comes first, then Reflect and at last Line. Thru is performed by putting two horn antenna facing straight to each other and placing nothing between them. For Reflect standard, a metal plate is placed at the middle between two antennas. The port 2 antenna should be moved away by the thickness of the metal plate as shown in Figure 3.11 during measuring reflect standard. And for Line standard, the port 2 antenna
should be moved away from its initial position by quarter wavelength of measurement frequency.

It should be noted that the quarter wavelength’s frequency should be the center frequency of the measurement band. For example: for X-Band free space measurement, quarter wavelength will be calculated for 10.3 GHz. It’s the center frequency of X-Band. After measuring the Line standard, the port 2 antenna should be brought to its initial Thru standard position.

A new calibration kit need to be created by defining offset delay and offset loss at VNA calkits if TRL calibration is required to perform. The below equations value should be used to calculate offset delay and loss while defining ‘Line standard’ at VNA calkit. Offset delay is the delay which is used in compensating the difference between the physical reference plane and electrical plane [21]. It is expressed as

\[
\text{Offset delay} = \frac{t \sqrt{\varepsilon}}{c}
\]  

(3.2)
where ‘l’ is the physical offset length from reference plane. $\varepsilon$ is the signal transmitting medium. For free space, its air.

Offset loss is the propagation loss of the transmission line per unit length at a normalization frequency [32]. It shifts the electrical plane to physical plane for adjustment of the loss [33]. The calculation of the offset loss is performed by using the following equations [32]

$S_{11}$ data:

Linear mag: (Offset loss) $= -2 \ln( |S_{21}|_{1GHz} ) \left( \frac{Offset\ Z_0}{Offset\ delay} \right)$ (3.3)

Log mag (dB): (Offset loss) $= \left( -\frac{\ln(10)\ dB_{1GHz}}{10} \right) \left( \frac{Offset\ Z_0}{Offset\ delay} \right)$ (3.4)

$S_{21}$ data:

Linear mag: (Offset loss) $= -\ln( |S_{11}|_{1GHz} ) \left( \frac{Offset\ Z_0}{Offset\ delay} \right)$ (3.5)

Log mag (dB): (Offset loss) $= \left( -\frac{\ln(10)\ dB_{1GHz}}{20} \right) \left( \frac{Offset\ Z_0}{Offset\ delay} \right)$ (3.6)

where $Offset\ Z_0 = Lossless\ characteristic\ impedance\ of\ transmission\ line$

3.1.3.3.2 TRM calibration

TRM calibration is same as TRL calibration except instead of using line standard, match standard is used as shown in Figure 3.12.
Like TRL, here T stands for thru, R stands for Reflect. Reflect actually is a short corresponding to transmission line theory and Match standard acts as a matched load corresponding to transmission line theory.

Figure 3.12: TRM calibration steps [30]

3.1.3.3.3 GRL calibration

This calibration is from TRL class family. GRL calibration is achieved by doing 2 step calibration.

1st step: 1st step calibration means calibrating the VNA itself. It can be done simply by doing Ecal or by using mechanical calkits available to the VNA. This calibration removes error up to the connector (Antenna input) of VNA cable as shown in Figure 3.13. Hence, after 1st step of calibration, those connectors become the reference point.
54

2nd step: The second step of GRL calibration involves with the free space measurement setup. In order to do 2nd step calibration, first ‘Thru standard’ and ‘Reflect standard’ are measured. From the Figure 3.14, it is seen that ‘Thru’ and ‘Reflect’ standards are the same as ‘TRL’ calibration’s thru and reflect. After that time gating is applied. After successful completion of GRL calibration, the error free calibration reference plane shifts from VNA connector/antenna input side to the surface of the metal plate.

It should be noted, usually sample thickness should be same as metal plate thickness. Hence, if sample thickness changes, the calibration reference plane changes. When doing
the calculation for dielectric characteristics, this should be kept in mind. For example: metal plate thickness is 15 mm but sample has thickness of 10 mm. In this case, calibration will still assume the sample thickness as 15 mm without knowing that the original sample is of 10 mm sample and rest 5 mm is air. Because, when free space calibration is done, calibration is achieved up to the surface of the metal plate. Therefore, any changes in the thickness and position of the metal plate create shifting in the reference plane.

### 3.1.4 Resonant Cavity Method

The resonant cavity technique is the most accurate and easiest technique among all the methods. The measurement setup requires a vector network analyzer, resonant cavity, and the sample. The resonant cavity is connected between two ports of the analyzer as shown in Figure 3.15.

![Resonant cavity measurement system](image)

Figure 3.15: Resonant cavity measurement system [34]

Permittivity is calculated from the measurement of the quality factor, $Q$ and resonant frequency, $f_c$. These two parameters are measured in two steps. The first measurement is performed for empty cavity without the sample. The sample is inserted in the next measurement. The resonant cavity and quality factor at the empty cavity are mostly
evaluated by the dimension of the cavity. The quality factor and the bandwidth widens when the sample is inserted into the cavity. The resonant frequency also shifts downwards. The relative real and imaginary permittivity are obtained from the formula associated with the quality factor and resonant frequency. The formula is mentioned below \[2,7\]

\[
\varepsilon'_r = 1 + \frac{V_c (f_c-f_s)}{2V_s f_s} \quad \text{(3. 7)}
\]

\[
\varepsilon''_r = \frac{V_c}{4V_s} \left( \frac{1}{Q_s} - \frac{1}{Q_c} \right) \quad \text{(3. 8)}
\]

where \( f_c \) = Resonant frequency of the empty cavity

\( f_s \) = Resonant frequency of the filled cavity

\( Q_c \) = Quality factor of the empty cavity

\( Q_s \) = Quality factor of the filled cavity

\( V_s \) = Volume of the empty cavity

\( V_c \) = Volume of the sample

The resonant cavity measurement system can only show dielectric properties for a single frequency at a time. It doesn’t show the data for broad frequency range. There are few differences between previously discussed three broadband transmission techniques to the resonant technique. The Table 3.6 compares the Resonant technique with the broadband techniques.
Table 3.6: Resonant vs. broadband transmission techniques [17, 28]

<table>
<thead>
<tr>
<th>Aspects</th>
<th>Resonant technique</th>
<th>Broadband technique</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency coverage</td>
<td>Single frequency measurement coverage</td>
<td>Broadband or banded measurement coverage</td>
</tr>
<tr>
<td>Calibration</td>
<td>Calibration not required</td>
<td>Calibration required</td>
</tr>
<tr>
<td>Measurement of lossy material</td>
<td>Extremely low loss material can be measured such as ceramic</td>
<td>Material with high loss is measured</td>
</tr>
</tbody>
</table>

3.2 Measurement Uncertainty

Measurement uncertainty is the deviation of the measured value from the actual value. Usually systematic error is responsible for measurement uncertainty in a particular system. Measurement uncertainty is present in the above-discussed techniques as well for characterization of the material. The determination of the measurement uncertainty can often become complicated and takes more effort than the actual measurement.

3.2.1 Coaxial Probe System Uncertainty

In the coaxial probe method, the source of error comes from the air gaps between cable and the sample [24, 27]. Air gap may exist either between the sample and inner conductor or between the sample and the outer conductor. The effect of the air gap presence between sample and the inner conductor is more significant than the later one.
Layer capacitor model (Baker-Jervis) is used to calculate the effect of air gaps [6]. Air gaps are assumed to be capacitor in this model and coaxial airline segment filled with sample can be taken as capacitor in series. Then the equivalent capacitor becomes

\[
\frac{1}{c_m} = \frac{1}{c_1} + \frac{1}{c_2} + \frac{1}{c_3}
\]  

(3.9)

For length L the coaxial line capacitance become

\[
C = \frac{2\pi \varepsilon_r L}{\ln\left(\frac{D_2}{D_1}\right)}
\]  

(3.10)

The equivalent capacitor for the coaxial airline can be found by equation 3.6

\[
\frac{\ln\left(\frac{D_4}{D_1}\right)}{\varepsilon_m} = \frac{\ln\left(\frac{D_2}{D_1}\right)}{\varepsilon'_1} + \frac{\ln\left(\frac{D_4}{D_2}\right)}{\varepsilon'_c} + \frac{\ln\left(\frac{D_4}{D_3}\right)}{\varepsilon'_1}
\]  

(3.11)

where D_1, D_2, D_3 and D_4 are the diameter of inner conductor, its boundary, outer conductor and its shield respectively as shown in Figure 3.16. \(\varepsilon'_c\) is the corrected relative real permittivity, \(\varepsilon'_m\) is the measured the real permittivity, and the \(\varepsilon'_1\) is the real permittivity of
the air gap. From the above the equation 3.11, the corrected real relative permittivity can be found \([27, 31]\) as

\[
\epsilon'_c = \frac{\epsilon'_m \ln \left( \frac{D_3}{D_1} \right)}{\epsilon'_m \ln \left( \frac{D_2}{D_1} \right) - \epsilon''_m \ln \left( \frac{D_2}{D_1} \right) + \ln \left( \frac{D_4}{D_3} \right)} \tag{3.12}
\]

To simplify parameters, \(L_1, L_2\) and \(L_3\) are incorporated. The expression of those three parameters are

\[
L_1 = \ln \left( \frac{D_2}{D_1} \right) + \ln \left( \frac{D_4}{D_3} \right) \tag{3.13a}
\]

\[
L_2 = \ln \left( \frac{D_2}{D_2} \right) \tag{3.13b}
\]

\[
L_3 = \ln \left( \frac{D_4}{D_1} \right) \tag{3.13c}
\]

By replacing the value of \(L_1, L_2\) and \(L_3\) with them in the equation 3.12, the corrected real and imaginary part become

\[
\epsilon'_c = \epsilon''_m \frac{L_2}{L_3 - \epsilon''_m L_1} \tag{3.14}
\]

\[
\epsilon''_c = \left( \epsilon'_c \frac{\epsilon''_m}{\epsilon'_m} \right) \frac{L_3}{L_3 - L_1 \epsilon''_m \left( 1 + \left( \frac{\epsilon''_m}{\epsilon'_m} \right)^2 \right)} \tag{3.15}
\]

And the corrected real and imaginary permeability can be expressed as

\[
\mu'_c = \mu''_m \frac{L_3 - L_2}{L_2} \tag{3.16}
\]

\[
\mu''_c = \mu''_m \frac{L_3}{L_2} \tag{3.17}
\]

\(\mu_c\) is the corrected permeability and \(\mu_m\) is the measured permeability
3.2.2 Transmission line Uncertainty

Transmission line technique uses coaxial cable or waveguide as transmission line. The above coaxial cable uncertainty calculation can be used to evaluate the measurement error when coaxial cable is used for in transmission line technique. However, when waveguide is used as transmission line, a different approach should be used.

3.2.2.1 Uncertainty for Air gap in Waveguide

One of the conditions of transmission line with waveguide is that sample dimension should be similar to the inner dimension of the waveguide. Otherwise, the measurement data will have uncertainties.

![Figure 3.17: Air gap demonstration in waveguide](image)

(a) Air gap at height and width of the waveguide (b) Air gap at width only [27]

When sample is smaller than the inner dimension, air gaps form along the height and width of the waveguide. The error for air gap of height is negligible but error becomes significant for the width. For sample dimension ‘d’ and waveguide width ‘b’ as shown in Figure 3.17, the revised formula for real and imaginary permittivity considering the error part becomes

\[
\varepsilon' = \varepsilon_m' \frac{d}{b-(b-d)\varepsilon_m} \quad (3.18)
\]
where \( \varepsilon_c \) is the corrected permittivity and \( \varepsilon_m \) is the measured permittivity. Similarly, the relative real and imaginary permittivity can be shown as

\[
\mu'_c = \mu'_m \left( \frac{b}{d} \right) - \left( \frac{b-d}{d} \right) \\
\mu''_c = \mu''_m \left( \frac{b}{d} \right)
\]

\( \varepsilon'_c \) is the corrected real part of the permittivity, \( \varepsilon''_c \) is the corrected imaginary part of the permittivity, \( \varepsilon'_m \) is the corrected real part of the permittivity, \( \varepsilon''_m \) is the corrected imaginary part of the permittivity, \( \mu'_c \) is the corrected real part of the permeability and \( \mu''_c \) is the corrected imaginary part of the permeability. The corrected loss tangent is

\[
tag_{\varepsilon_c} = \tan \delta_m \frac{b}{b-(b-d)\varepsilon'_m}
\]

Where \( \tan \delta_c \) and \( \tan \delta_m \) is the corrected and measured loss tangent, respectively.
3.2.2.2 Effect of Sample Placement

Sample displacement from calibration plane can be a source of error in transmission line measurement techniques. Calibration reference plane is the plane up to which the calibration is performed and system loss is accounted for. If sample is placed inside the waveguide with some distance from the reference plane similar to that shown in the Figure 3.18, a phase correction is required for this additional length [24]. If the distance from port 1 reference plane to sample is ‘a’ and port 2 sample to reference plane is ‘b’ then phase correction can be performed using the following expression

$$\Delta \phi_{11} = 2a \frac{2\pi f}{c}$$  (3.23)

$$\Delta \phi_{21} = (a + b) \frac{2\pi f}{c}$$  (3.24)

here f is the measurement frequency, c is the speed of light, $\Delta \phi_{11}$ and $\Delta \phi_{21}$ are the phase correction value for $S_{11}$ and $S_{21}$ respectively.
3.2.3 Free space Measurement Uncertainty

Usually, in free space measurement, three types of calibration are performed [30]. They are TRL (Thru, Reflect, Line), TRM (Thru, Reflect, Match) and GRL (Gated, Reflect, Line).

The measurement error occurs usually after the calibration. In a regular free space system, uncertainty appears due to imperfection in the calibration standard, sample misplacement and instrumentation errors. Among these, the major source of error is sample misplacement. If the sample or sample holder containing sample translates or rotates from the calibration reference plane even by small amount similar to the Figure 3.19, it includes error in the S- parameter measurement [24]. From Figure 3.15, it can be seen that the path length of $S_{11}$ varies depending on the position angle of the sample. $S_{21}$ path length is independent of the sample position. Since, $S_{11}$ path length depends on the sample position, imperfect sample position changes the phase of the $S_{11}$. Besides, all reflected power can’t be received by the transmitting antenna when sample position changes. If signal incidents

![Figure 3.19: Free space uncertainty due to sample misalignment [24]](image-url)
on the sample at angle \( \phi \) and reflects at angle \( 2\phi \), then not all the reflected signal returns to the source. Thus, the measured \( S_{11} \) magnitude becomes incorrect. If sample misalignment can’t be avoided, then such dielectric properties conversion technique should be used which doesn’t account \( S_{11} \) term in its calculation.
CHAPTER 4

FREE SPACE QUASI-OPTIC MEASUREMENT SYSTEM

For the characterization of the biological samples, free-space quasi-optical measurement system is used. In the chapter 4, the theory and measurement procedure is discussed. In addition to that, the calculation required to conduct the measurement is shown as well.

4.1 Overview

Free-space Quasi-optic measurement system can be considered as an advancement of a free-space system which exploits beam collimation to perform the measurement. Since the sample under test (SUT) is very small for the experiment, use of conventional free-space measurement would not allow intercepting most of the beam power passage through the sample due to regular beam’s expansion characteristics over propagation distance. If beam size becomes larger than the sample, then beam power scatters from the edge of the sample. The beam growth characteristics also does not permit to achieve ideal far-field at the location of the sample. These issues can be resolved by the employment of the quasi-optical measurement system. The primary difference of the Quasi-Optic from the traditional Free-space method is that it uses Gaussian beam instead of regular beam to ensure far-field and maximum possible amount of power interception by the sample. In order to conduct Quasi-Optic measurement successfully, a detail understanding of the Gaussian beam and its parameters are necessary.
4.2 Gaussian Beam Theory

Gaussian beam is defined as the beam which has relatively small beam divergence perpendicular to the beam axis. It exhibits Gaussian intensity profile along the propagation axis which is much higher than the conventional beam. In optics, Gaussian approximation is often used to demonstrate laser beam intensity.

In the free-space quasi optical measurement system, antenna with Gaussian beam profile has been used. Gaussian beam allows the measurement system to achieve far-field at the measurement location of sample and its focused characteristics ensures entire beam to incident on the sample surface.

4.2.1 Gaussian Beam Fundamentals

Any wave propagating through the uniform medium can be expressed using the homogenous Helmholtz equation \([8, 11, 23]\) which is

\[
\Delta^2 \psi + k_0^2 \psi = 0 \tag{4.1}
\]

where \(\psi\) is a component of the electric field, \(E\) or magnetic field \(H\) and \(k_0\) is the wave number.

\[
k_0 = \frac{2\pi}{\lambda_0} \tag{4.2}
\]

where \(\lambda_0\) is the free space wavelength. If, the wave is propagating through a positive \(z\) direction, \(\psi\) can be written as

\[
\psi = u(x, y, z) \exp(-jk_0z) \tag{4.3}
\]
where \( u \) is the complex function which propagates as non-plane wave \([8, 23]\). Substituting (4.3) in the rectangular co-ordinates expression of equation (4.1) gives us the following equation

\[
\frac{\partial^2 u}{\partial x^2} + \frac{\partial^2 u}{\partial y^2} + \frac{\partial^2 u}{\partial z^2} - 2 j k_0 \frac{\partial u}{\partial z} = 0
\]  

(4.4)

If the variation of the magnitude of \( u \) is small in comparison to the size of the wavelength and the variation along the \( z \) is much smaller than \( x \) and \( y \), equation (4.4) can be simplified and expressed as

\[
\frac{\partial^2 u}{\partial x^2} + \frac{\partial^2 u}{\partial y^2} - 2 j k_0 \frac{\partial u}{\partial z} = 0
\]  

(4.5)

Equation (4.5) is paraxial equation. The solution of this equation gives the expression for Gaussian beam. The simplest solution for \( u \) from equation (4.5) is

![Figure 4.1: Representation of Gaussian beam profile [23]](image)
where, \( r^2 = x^2 + y^2 \) defines the beam circle and distance from the axis, \( w \) is the beam radius, and \( R \) is the radius of the curvature of the beam as shown in the Figure 4.1.

The above equation (4.6) describes the Gaussian beam’s amplitude taper characteristics. For better understanding, the equation 4.6 can be divided into three parts. Each part contributes to certain characteristics of the Gaussian beam. The part \( \frac{w_0}{w(z)} e^{-r^2/w^2(z)} \) is called ‘amplitude factor’ and it describes the shape of the beam and how the shape varies along the radial (x-y coordinates) and propagating direction z to that shown in Figure 4.1.

Another part of the equation is \( e^{j\tan^{-1}(z/R)} \) which is called ‘longitudinal phase factor’. It expresses the beam’s shift from plane wave to spherical wave. The remaining part of the equation is \( e^{jkr^2/(2Rz)} \). It is called ‘radial phase factor’ which shows how the phase changes in the radial direction and how curve the phase front of the beam becomes when the beam travels along z direction.

4.2.2 Parameters of Gaussian Beam

The expression of Gaussian beam function from equation (4.6) is significant as it introduces several Gaussian parameters such as beam radius and radius of the curvature which play an important role in Quasi-Optic measurement system. The beam radius and radius of the curvature of the Gaussian beam can be determined from the following expressions.
Beam radius, \( w(z) = w_0 \left[ 1 + \left( \frac{Z}{Z_R} \right)^2 \right]^5 \) \hspace{1cm} (4. 7)

Radius of the curvature, \( R(z) = z \left( 1 + \left( \frac{Z_R}{Z} \right)^2 \right) \) \hspace{1cm} (4. 8)

Using the expression from equation 4.7, radius of a Gaussian beam can be determined for any location along \( z \) axis. Here, \( w_0 \) is the beam waist which is the minimum radius of the beam located at the origin \( z=0 \) and occurs at beam focus and \( z \) is the distance between the beam waist and the location from where the beam radius is measured. Equation 4.8 is used to determine the radius of the curvature as a function of position along \( z \) axis as shown in Figure 4.1. Both equations have a common term \( Z_R \) called the Rayleigh range. Rayleigh range is often used in Gaussian optics to understand the depth of focus of the Gaussian beam. It actually is the distance from the beam waist location to where the spot size increased to \( w(z_R) = \sqrt{2}w_0 \) to that shown in Figure 4.2. In this region, phase of the wave front is considered constant.

Figure 4. 2: Rayleigh range of a Gaussian beam [15]
The Rayleigh range feature of the Gaussian beam is exploited in the Free-space Quasi-Optic system to achieve planar wave incident on the sample by placing sample within the Rayleigh region [15]. The expression of the Rayleigh region is

\[ Z_R = \frac{k}{2} w_0^2 \]  
(4.9)

In order to find the beam radius of the Gaussian beam for any location along with the z axis and radius of the curvature, Rayleigh range must be determined.

In addition to beam radius \( w(z) \) and radius of the curvature \( R(z) \), beam-waist parameter, \( w_0 \), its location inside the antenna and total power incident on the lens from antenna is required to perform Free-space Quasi-Optics successfully. Beam-waist of the antenna is the radius of the beam of a specific location along the travelling direction (z-axis) where beam power drops by \( 1/e^2 \) or -8.686 dB from its peak power [35]. The most common approach of beam-waist measurement technique is to measure the \( 1/e^2 \) angle where power reduces to -8.686 dB from the far-field radiation pattern Gain vs Angle data. The \( 1/e^2 \) angle can be expressed by the following equation [36]

\[ \tan \theta_0 = \frac{w(z)}{z} \]  
(4.10)

In far-field measurement, the distance \( z \) is assumed to be infinite. If \( \theta_0 \) is considered as asymptotic beam growth angle and \( w(z) \) from equation 4.7 is replaced at equation 4.10 then angle of \( 1/e^2 \) power can be expressed as [1, 23, 26]

\[ \tan \theta_0 = \lim_{Z \to \infty} \frac{w_0 \left[ 1 + \left( \frac{Z}{Z_R} \right)^2 \right]^5}{z} \]  
(4.11a)
If angle for $1/e^2$ is determined, beam-waist can be found from equation 4.11b. The location of the beam waist is still unknown. For feed horn antenna, it can be assumed that, the field distribution at the aperture is spherical with radius of curvature $R$ equal to the slant length $\rho$ of the horn as illustrated in Figure 4.5 [36].

![Figure 4.3: Horn antenna parameter representation [23]](image)

Inverse formula which is presented in [26] can be exploited to find the location of the beam-waist.

$$z = \frac{R}{2} \left\{ 1 \pm \left[ 1 - \left( \frac{2\pi \omega_0^2}{\lambda_0 R} \right)^2 \right]^{\frac{5}{2}} \right\}$$  \hspace{1cm} (4.12)

where, $R$ is equal to slant length $\rho$ and $z$ is the distance constant phase surface and beam-waist location as shown in Figure 4.5.

The equation 4.12 can be rewritten for calculating the beam width. The modified expression becomes
\[
\Delta = \frac{\rho}{2} \left\{ 1 \pm \left[ 1 - \left( \frac{2\pi w_0^2}{\lambda_0 p} \right)^2 \right]^{\frac{5}{2}} \right\}
\]  

(4.13)

where \( \Delta \) is the distance between peak of the constant phase surface to the beam waist. However, we need to know the beam-waist distance from the aperture plan which can be determined by the equation realized from Figure 4.5. Hence,

\[
\text{Beam-waist location from aperture plane, } d = \Delta + p - \rho
\]

(4.14)

There are few other expressions derived from Gaussian function which tells how much power a Gaussian beam can contain in a given area. The expressions are significant particularly for Free-space Quasi-Optics measurement as it provides the information of the total power amount the sample and the lens contain and adjust the sample size according to it. The edge taper and fractional power expressions are used to express how much power is falling outside of the sample/lens and how much is intercepted by the sample/lens respectively. The edge taper \( T_e \) is defined by the equation below

\[
T_e (r_e) = \exp \left[ -2 \left( \frac{r_e}{w} \right)^2 \right]
\]

(4.15)

In dB, the equation can be rewritten as

\[
T_e (dB) = 8.686 \left( \frac{r_e}{w} \right)^2
\]

(4.16)

where, \( r_e \) denotes the lens/sample radius and \( w \) is the beam waist. Equation 4.16 tells how much power resides outside of the area with radius \( r_e \).

The fractional power can be described by, \( F_e \) which describes how much power is contained by a sample with radius \( r_e \).
4.2.3 Gaussian Beam Transformation through Focus Lens

The transformation of the Gaussian beam through lens is not reciprocal. When a launched beam from a Gaussian source goes through the focusing lens, it refocuses itself. As a result, the output beam-waist also changes. The relationship of the output beam-waist $w_{out}$ to the beam-waist of the input $w_{in}$ and focal length can be derived from ray tracing.

\[ F_e = 1 - T_e (r_e) \]  \hspace{1cm} (4.17)

\[ w_{out} = \frac{\lambda_0 \times f}{\pi w_{in}} \]  \hspace{1cm} (4.18)
From equation 4.18, the output beam-radius can be obtained. However, if distance between antenna apertures to input plane of the lens can’t be maintained, the formula becomes invalid. Therefore, a different formula has taken into account to calculate the output beam-waist. In 1983, Sidney A. Self derived lens transformation expression for Gaussian beam from normal geometrical optics theory. In geometric optics, when there is no lens between the point objects/source to the image, the radius of the curvature equal the distance between them. In Figure 4.5, \( w_0 \) denotes the location of point object/source before lens and \( w'_0 \) denotes the location of the image source. When a lens is introduced between them, then the change in the radius is expressed by \( 1/f \) where \( f \) is the focal length of the lens. Then the standard lens or mirror formula becomes [40]

\[
\frac{1}{s} + \frac{1}{s'} = \frac{1}{f}
\]  

(4. 19)

where, \( S \) is the distance from source to the lens and \( S' \) is the distance between lens to image.

The equation is modified for Gaussian beam by incorporating an additional term \( \frac{z_k}{s-f} \) with the ‘\( S' \)’. The equivalent Gaussian expression becomes [40]

\[
\frac{1}{s} \left( \frac{z_k}{s-f} \right) + \frac{1}{s'} = \frac{1}{f}
\]  

(4. 20)

where \( S' \) is the location of the new beam-waist after beam transformation. In comparison to Quasi-optics, ‘\( S' \)’ is the distance between aperture plane of the transmitting antenna to lens and \( S' \)’ is the distance from lens to sample position. Equation (4.20) can be rewritten as to show \( S' \). Therefore,
Output beam-waist location, \( s' = \frac{1}{f - \frac{z_R}{s + \frac{z_R}{s}}} \)  \hspace{1cm} (4. 21)

Figure 4. 5: Geometry of imaging of Gaussian beam [40]

The output beam-waist for the location \( S' \) can be determined from the magnification amount of the beam after its transformation through lens. The magnification is the ratio of output beam-waist to the input beam-waist and it is expressed as [40]

\[
m = \frac{w'_0}{w_0} \hspace{1cm} (4. 22a)
\]

\[
m = \frac{1}{\sqrt{1 - (\frac{z_R}{f})^2 + (\frac{z_R}{f})^2}} \hspace{1cm} (4. 23)
\]

Using the equation 4.22b, the new beam-waist \( w'_0 \) can be obtained. The Rayleigh range for this new beam waist can also be calculated using the magnification formula. The Rayleigh range for the new beam waist is –

\[
Z'_R = m^2 Z_R \hspace{1cm} (4. 24)
\]

where, \( Z'_R \) and \( Z_R \) are the Rayleigh range for output beam-waist and input beam-waist, respectively.
4.3. Free-space Quasi-Optical System

Free-space Quasi-Optics technique is the modified method of Free-space measurement system. One of the major drawbacks of conventional Free-space method is the beam growth of the antenna. The gradual beam growth doesn’t allow the signal power concentrated on the sample under test (SUT) [35]. It also doesn’t allow obtaining Far-field distance within short distance which could be essential especially if the test bench is quite small. Both of the issue can be resolved employing Quasi-Optics in the Free-space system.

4.3.1 Free-space Quasi-Optical Test Bench Setup

A typical Free-space Quasi-Optic setup is illustrated in Figure 4.6. It consists of two Gaussian horn antennas, two lenses and sample holder. The horn antennas are connected to the Network Analyzer. During operation, the horn antennas illuminate the lenses by launching focused Gaussian beam on it. Then the Gaussian beam refocuses with the hyperbolic lenses. The SUT is placed in mid-way between two lenses in a way that it lies at the beam waist location of the refocused beam. In this way, SUT receives most of the

![Figure 4.6: Free-space Quasi Optic test bench setup](image-url)
power of the beam intercepted through it. Another advantage include being exposed to the
planar wave since phase of the wave-front of the Gaussian beam around beam-waist
location and within Rayleigh range are equal. As Quasi-Optics setup requires Gaussian
antennas, the choice of antenna with Gaussian beam radiating capability is vital. In the
experiment, two pair of VDI WR-6.5 and WR-6.5 Gaussian Antennas of Virginia Diodes
Incorporated manufactured by Custom Microwave are used. The physical specification of
these antennas are provided below at Table 4.1. Thorlab’s two pairs of dielectric lenses
with constant dielectric parameters over THz frequency ranges are used. The Table 4.2 has
shown the physical parameters of these lenses.

Table 4.1: Physical parameter of the Gaussian Antenna uses in the experiment

<table>
<thead>
<tr>
<th></th>
<th>VDI WR-6.5 Feed Horn</th>
<th>VDI WR-4.3 Feed Horn</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency range</td>
<td>110 GHz - 170 GHz</td>
<td>170 GHz – 260 GHz</td>
</tr>
<tr>
<td>Horn type</td>
<td>Conical</td>
<td>Conical</td>
</tr>
<tr>
<td>Horn length</td>
<td>26 mm</td>
<td>16.5 mm</td>
</tr>
<tr>
<td>Aperture diameter</td>
<td>10.8 mm</td>
<td>7.1 mm</td>
</tr>
<tr>
<td>Taper half angle</td>
<td>11.7 degree</td>
<td>12.1 degree</td>
</tr>
</tbody>
</table>
Table 4.2: Common Specification of the Thorlab’s LAT100 Dielectric Lens

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Specification</th>
</tr>
</thead>
<tbody>
<tr>
<td>Material</td>
<td>PTFE Teflon</td>
</tr>
<tr>
<td>Design frequency range</td>
<td>Up to 500 GHz</td>
</tr>
<tr>
<td>Diameter</td>
<td>50 mm</td>
</tr>
<tr>
<td>Focal length</td>
<td>100 mm</td>
</tr>
<tr>
<td>Thickness at the center</td>
<td>12.8 mm</td>
</tr>
<tr>
<td>Dielectric Constant</td>
<td>1.96</td>
</tr>
</tbody>
</table>

4.3.2 Free-space Quasi-Optic Parameter Calculation

This section presents the parameters which need to be determined in order to perform Free-space Quasi-Optic measurement. The following parameter should be known before conducting measurement at Quasi-Optic setup. They are

- Distance from antenna aperture plane to lens surface
- Distance between output lens surface to sample
- Beam-waist, $w_{in}$ of Gaussian beam of the antenna
- Beam waist location of the antenna
- Beam-waist, $w_{out}$ of the refocused Gaussian beam
- Amount of power intercepted by the lens and the sample
- Size of the sample
From the equation 4.11b, beam-waist $w_0$ can be calculated if $\theta_0$ is known. For the experiment, as mentioned in previous section, two pairs of antennas from Virginia Diodes WR 6.5 and WR 4.3 have been used. WR-6.5 and WR-4.3 operates in the bands of 110-170 GHz and 170 GHz-260 GHz, respectively. To calculate their beam-waist, wavelength of the band mid-frequency is considered. The band mid-frequency for WR-6.5 and WR-4.3 are 140 GHz and 215 GHz and their wavelength are 2.143 mm and 1.34 mm, respectively. Their $1/e^2$ angle can be determined from Figures 4.7 and 4.8. From the figure and data provided by VDI, the $1/e^2$ angle of WR-6.5 and WR-4.3 are 9.45 and 8.98 degree, respectively. Therefore, beam-waist can be calculated from equation 4.11b as

$$ Beam-waist \text{ of WR-6.5} = \frac{\lambda_0}{\pi \tan \theta_0} = \frac{2.143}{\pi \times \tan 9.45} = 4.1 \text{ mm} $$

$$ Beam-waist \text{ of WR-4.3} = \frac{\lambda_0}{\pi \tan \theta_0} = \frac{1.34}{\pi \times \tan 8.98} = 2.7 \text{ mm} $$

Figure 4.7: Gain vs. Far-field radiation pattern of the WR 6.5 Virginia Diodes antenna, plot provided by VDI partner company Custom Microwave [38]
Distance of beam waist location from the aperture plane WR-6.5 and WR-4.3 antenna has been found from equation 4.1 which are 4.4 mm and 3.6 mm, respectively. As per the discussion of the special case of the Gaussian beam transformation through lens, the distance between the beam-waist locations of the transmitting antenna to lens input plane and input plane of the lens to output beam waist location are kept equal to the focal distance of the lens at the experiment setup. According to the equation 4.18, the output beam-waist is found to be

\[
\text{Output beam-waist for WR-6.5}, \quad w_{\text{out}} = \frac{\lambda_0 \times f}{\pi w_{\text{in}}} = \frac{2.143 \times 100}{\pi 4.1} = 16.63 \text{ mm}
\]

\[
\text{Output beam-waist for WR-4.3}, \quad w_{\text{out}} = \frac{\lambda_0 \times f}{\pi w_{\text{in}}}
\]
The above calculation provides relatively large beam-waist for considering fixed distance between antenna and lens. Since, the SUT is quite small, Sidney A. Self method is used to look for smaller beam-waist by closing the distance between antenna and lens. If antenna and lens distance is considered 40 mm, then using equation 4.21 and 4.22b, the beam waist location and output beam-waist for WR-6.5 is found as

Output beam-waist location for WR-6.5, \( s' \) = \( \frac{1}{s + \frac{Z_R}{s-f}} \) = 94.32 mm

Output beam waist for WR-6.5, \( w'_0 = w_0 \times \frac{1}{\left[ 1 - \left( \frac{s}{f} \right)^2 \right]^{.5}} \) = 8.08 mm

Using the same formula, beam-waist location and output beam-waist can be found for WR-4.3 antennas.

Output beam-waist location for WR-4.3, \( s' \) = \( \frac{1}{s + \frac{Z_R}{s-f}} \) = 93.78 mm

Output beam waist for WR-4.3, \( w'_0 = w_0 \times \frac{1}{\left[ 1 - \left( \frac{s}{f} \right)^2 \right]^{.5}} \) = 5.31 mm

Here, \( f \) is the focal length of the lens with value 100 mm, \( s \) is the distance from antenna to lens which is 50 mm.
In free-space quasi-optical measurement system, larger sample size is always desired as sample can intercept maximum amount of beam power. It is found that, when sample size is twice the size of the beam, 99.97% power can be intercepted. Consider the sample under test (SUT) is 10 mm of radius size and if the beam size for WR-6.5 and WR-4.3 are 8.08 mm and 5.31 mm respectively, then

Amount of power concentrated on the sample for WR-6.5, \( F_e = 1 - \exp[-2 \left( \frac{r_e}{w} \right)^2] \)

\[ = .953 = 95.3\% \]

Amount of power concentrated on the sample for WR-4.3, \( F_e = 1 - \exp[-2 \left( \frac{r_e}{w} \right)^2] \)

\[ = .9991 = 99.91\% \]

It can be realized that for the distance of aperture plane to lens of 50 mm and beam-waist of 8.08 mm and 5.31 mm for WR-6.5 and WR-4.3, the beam power interception is 95.3 and 99.91 %, respectively.

Determination of the power interception of the lens is also necessary to ensure all the power is transmitting antenna passing through the lens and signal is not attenuated while passing the lens. A perfect lens should not absorb energy. It should just transform the phase by introducing phase shift. In order to find the power interception at the lens surface, the beam radius at the surface should be known. Since, the distance from antenna to lens surface is considered 50 mm, the beam waist is found from equation 4.7.

Beam radius for WR-6.5, \( w(z) = w_0 \left[ 1 + \left( \frac{z}{z_R} \right)^2 \right]^5 \)

\[ = 24.46 \text{ mm} \]
Beam radius for WR-4.5, \( w(z) = w_0 \left[ 1 + \left( \frac{z}{z_R} \right)^2 \right]^{5} \)

\[ = 15.32 \text{ mm} \]

Therefore, the power intercepted by the lens with diameter of 50 mm as shown in the Table 4.2 for WR-6.5 and WR-4.5 are,

Power interception by the lens for WR-6.5, \( F_e = 1 - \exp\left[-2\left(\frac{r_e}{w}\right)^2\right] \)

\[ = .9998 = 99.98 \% \]

Power interception by the lens for WR-4.3, \( F_e = 1 - \exp\left[-2\left(\frac{r_e}{w}\right)^2\right] \)

\[ = 1.00 = 100 \% \]

The above calculations demonstrates that keeping antenna to lens distance to 50 mm gives beam-waist as low as 8 mm which is adequate for sample under test (SUT) in our experiment. The distance also permits 95% to 99.91 % power interception through the sample.

**4.3.3 Measurement Correction**

In free-space quasi-optical measurement system, the measurement uncertainty often comes from sample misplacement and incorporates error in phase of scattering parameters. Since,
antenna from port 1 and port 2 are in equal distance from the sample location, the recorded reflection and transmission coefficient are equal ($S_{11} = S_{22}$ and $S_{21} = S_{12}$) [35, 41].

![Diagram of sample suffering from misplacement in propagation direction](image)

Figure 4.9: Sample suffering from misplacement in propagation direction

When misplacement takes place as illustrated in Figure 4.9 where sample deviates from original location by $\delta$ in direction of propagation, the reciprocity is not maintained anymore in terms of phases of $S_{11}$ and $S_{22}$. Assume in ideal case, the S-matrix is

$$[S] = \begin{bmatrix} \Gamma_s & T_s \\ T_s & \Gamma_s \end{bmatrix} = \begin{bmatrix} |\Gamma_s| e^{j\phi_{rs}} & |T_s| e^{j\phi_{ts}} \\ |T_s| e^{j\phi_{ts}} & |\Gamma_s| e^{j\phi_{rs}} \end{bmatrix}$$ (4. 25)

where $S_{11} = S_{22} = \Gamma_s$ and $S_{12} = S_{21} = T_s$. When sample is moved from desired location by $\delta$ in the z- direction, then the resultant matrix becomes [35]

$$[S'] = \begin{bmatrix} e^{+j\beta_0 \delta} & 0 \\ 0 & e^{-j\beta_0 \delta} \end{bmatrix} [S] \begin{bmatrix} e^{+j\beta_0 \delta} & 0 \\ 0 & e^{-j\beta_0 \delta} \end{bmatrix}$$ (4. 26)
Inserting equation 4.25 in 4.26, the following expression can be found [23].

\[
[S'] = \begin{bmatrix} S'_{11} & S'_{12} \\ S'_{21} & S'_{22} \end{bmatrix} = \begin{bmatrix} |T_s| e^{j(\phi_{\Gamma S} + 2\beta_0 \delta)} & |T_s| e^{j\phi_{\Gamma S}} \\ |T_s| e^{j\phi_{\Gamma S}} & |T_s| e^{j(\phi_{\Gamma S} - 2\beta_0 \delta)} \end{bmatrix} \tag{4.27}
\]

If equation 4.27 is compared with the equation 4.23, it can be realized that, \(S_{21}\) and \(S_{12}\) are not changed, the magnitude of \(S_{11}\) and \(S_{22}\) is unaffected as well. However, the phase of \(S_{11}\) and \(S_{22}\) are changed by \(\pm 2\beta_0 \delta\). The phase can be corrected by averaging the changed phases of \(S_{11}\) and \(S_{22}\). It then cancels out the \(2\beta_0 \delta\).

\[
\phi_{\Gamma S} = \frac{\angle S'_{11} + \angle S'_{22}}{2} \tag{4.28}
\]
In this chapter, simulated and measured results of the healthy human tissue sample and cancer sample has been presented for the study of the dielectric properties of the material in two frequency bands of 110-170 GHz and 170-260 GHz. For the characterization of the biological tissues, the free-space quasi optical measurement system has been used.

5.1 Terahertz Measurement Configuration
The measurement of the samples in this thesis are performed using the Quasi Optic Free-space technique. As discussed in the previous chapter, in order to establish a proper Quasi Optic Free-space system according to the requirement of the sample and the frequency of operation, several parameters need to be determined. They are beam waist, location of the beam waist for the Gaussian beam output of the lens and its Rayleigh range, minimum thickness of the sample, and incident power amount from the source on the sample. To calculate these parameters, operating frequency, the lens focal length, antenna beam waist, antenna to lens distance of the setup and sample size must be known. The specification of the antenna and lens which has been used to conduct measurement at two bands 110-170 GHz and 170 - 260 GHz can be found in Tables 4.1 and 4.2. For the free-space quasi optical setup in the laboratory, antenna from VDI and dielectric lens from Thorlabs have been used. The specification of these equipment used for creating the laboratory measurement setup are shown at Table 5.1 whereas Table 5.2 shows the parameter calculated for the setup.
Table 5.1: Configuration for Quasi Optic setup at 110-170 GHz and 170-260 GHz

<table>
<thead>
<tr>
<th></th>
<th>Quasi Optic Specification for (110-170 GHz)</th>
<th>Quasi Optic Specification for (170-260 GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Antenna</td>
<td>WR 6.5 VDI Antenna</td>
<td>WR 4.3 VDI Antenna</td>
</tr>
<tr>
<td>Antenna Beam Waist</td>
<td>4.1 mm</td>
<td>2.7 mm</td>
</tr>
<tr>
<td>Lens Focal Length</td>
<td>100 mm</td>
<td>100 mm</td>
</tr>
<tr>
<td>Antenna to Lens distance</td>
<td>65 mm</td>
<td>65 mm</td>
</tr>
<tr>
<td>Frequency at Mid band</td>
<td>140 GHz</td>
<td>215 GHz</td>
</tr>
</tbody>
</table>

Table 5.2: Configuration for Quasi Optic setup at 110-170 GHz and 170-260 GHz

<table>
<thead>
<tr>
<th></th>
<th>Quasi Optic Specification for (110-170 GHz)</th>
<th>Quasi Optic Specification for (170-260 GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Beam waist from lens output</td>
<td>3.853 mm</td>
<td>2.503 mm</td>
</tr>
<tr>
<td>Location of output beam waist</td>
<td>169.799 mm</td>
<td>169.457 mm</td>
</tr>
<tr>
<td>Rayleigh range for Gaussian beam output of lens</td>
<td>7.98 mm</td>
<td>7.96 mm</td>
</tr>
<tr>
<td>Total power incident on sample</td>
<td>99.98 %</td>
<td>100 %</td>
</tr>
</tbody>
</table>
Using the specification from Table 5.1 and manipulating the equations 4.7, 4.17, 4.21, 4.23, and 4.24, the configuration is calculated as shown in Table 5.2. In order to create quick and repeatable free-space quasi optical measurement setup, software has been created on Matlab platform which codes can be found in the appendix A. It calculates the information required to create the Quasi Optic setup when specifications are provided to it. Figure 5.1 and 5.2 show the measurement configuration calculated by software for 110-170 GHz and 170-260 GHz band, respectively.

Figure 5.1: Quasi Optic Free-space measurement configuration calculation for 110-170 GHz
5.2 Terahertz Measurement Procedure

Measurement is performed after establishing the configuration of the Quasi Optic setup. Conducting measurement at 110-170 GHz and 170 – 260 GHz bands require THz source. The THz source has been set up by connecting THz extenders with the Vector Network Analyzer (VNA). The VNA used is PNA N5224A from Keysight Technologies that operates from 10 MHz to 43.5 GHz. Pair of VDI WR-6.5 and VDI WR-4.3 THz extenders are used for 110-170 GHz and 170 - 260 GHz frequency bands respectively.

The measurement setup involves several steps. It starts with two step calibration. The step by step procedure is mentioned below.

- First step is to calibrate up to the output port of VDI THz extender’s waveguide, usually the location at which the THz antenna connects.
• Then pair of THz antenna is connected at the port of two extenders. Then Free-space GRL calibration is performed to define the location of the sample measurement.
• Later the sample under test is placed at the defined location with the use of sample holder.
• Data is acquired by the Keysight 85071E material measurement software which is embedded in the computer of VNA.

5.2.1 Waveguide Calibration
The waveguide calibration requires full 2-port calibration with the use of calibration kit from VDI THz extenders. Two separate calibration is dedicated for two different frequency bands since extenders are different for each band. WR 6.5 and WR 4.3 calibration kit is used for WR-6.5 and WR-4.3 extenders, respectively. The calibration kit come as Short, Load (perfect match), and Shim (quarter wave, one-eighth wave) calibration standards. The WR 6.5 calibration kit consists of one waveguide, two pieces of WR-6.5 precision load, two pieces of short circuit, one piece of quarter wave delay, and two pieces of WR-6.5 eighth wave delay as shown in Figure 5.3.
Figure 5.3: WR-6.5 calibration kits

Figure 5.4: WR-4.3 calibration kits
WR-4.3 has two precision load, one waveguide for WR-4.3 band, two pieces of short circuit, two pieces of quarter wave delay and three pieces of eight-wave delay as seen from Figure 5.4.

The waveguide calibration steps are similar for both WR-6.5 and WR-4.3 bands. Before proceeding for calibration, power needs to be turned on through VNA menu for frequency of interest and the channel at which the calibration is performed. It turns on the power of the extenders. How much the signal will be multiplied from the VNA is set by ‘mm-Wave setup’ from ‘Macro’ option of VNA utility menu. RF IN frequency is multiplied by a factor of 4 and 6 for WR-6.5 and WR-4.3, respectively. The local oscillator frequency is multiplied by a factor of 6 for both bands as can be seen in Figures 5.5 and 5.6. The intermediate frequency (IF) is usually kept as 100 MHz. However, if calibrated data contains noise, then recalibration is performed with the decreased value of IF frequency. The assumption is that the more lower the IF frequency set, the less noisy the system becomes.

Figure 5.5: Configuration for “mm-Wave Setup” for frequency of 110-170 GHz
After setting the mm-wave setup, measurement of calibration standards are taken using the calibration kits. To start the calibration standard measurement, ‘Calibration Wizard is selected from response menu of VNA toolbar as shown in Figure 5.7.
This will bring Calibration wizard dialogue box which asks for the calibration types. Smart (Guided Calibration) is chosen and then next button is clicked as shown in Figure 5.8.

![Figure 5.8: Calibration types options from ‘Calibration Wizard’](image)

Then port numbers and the ports which will need to be calibrated is selected from the select port window shown in Figure 5.9.

![Figure 5.9: Calibration ports selection from ‘Select Ports’](image)

After this, the appropriate DUT connectors and specific kit needs to be chosen. For the 110-170 GHz, ‘WR-6.5 Waveguide’ and WR6p5_VDI are chosen as DUT connectors and
cal kits whereas ‘WR-4.3 Waveguide’ and WR4p3_VDI are chosen for 170-260 GHz frequency band. The DUT connectors and Cal Kits selection panel is shown in Figure 5.10.

![Guided Calibration: Select DUT Connectors and Cal Kits](image1)

**Figure 5.10: DUT connectors and Cal-kits selection panel**

Next the measurement of Short, Delay and Thru calibration standards take place. At first short standard is measured for port 1 by connecting the ‘Short’ standard that comes with the calckits. The Figure 5.11 shows port 1 short measurement of WR-4.3.

![Guided Calibration Step 1 of 4](image2)

**Figure 5.11: Port 1 short calibration standard measurement**
Similarly, port 2 short is measured by connecting the short standard at port 2 as shown in Figure 5.12.

![Figure 5.12: Port 2 short standard measurement](image)

After measuring the short standard of both ports, they are connected together to measure the through standard as shown in Figure 5.13.

![Figure 5.13: Calibration standard ‘Thru’ measurement between port 1 and port 2](image)

The last steps before saving the calibration is to perform quarter wave delay measurement. To perform quarter wave delay measurement, one standard called ‘Shim’ is placed between port 1 and port 2 as illustrated in Figure 5.14. The waveguide calibration ends once all the standards are measured and saved.
5.2.2 Free-space Calibration

Since waveguide calibration calibrates up to the waveguide port, free space calibration is required in order to calibrate out the antennas and the free-space path loss up to the location of the sample. There are three types of free-space calibration as discussed in chapter 3. In our Quasi-optic free space measurement, GRL calibration technique has been used. GRL stands for ‘Gated reflect line’ standards. GRL calibration is performed by 85071E material measurement software. The steps of GRL calibration are as follows -

- Set the calibration frequency range
- Define the reflector thickness used as ‘Reflect’ standard
- Select the calibrated file of the waveguide
- Transform the frequency domain to time domain and apply gating
- Collect the time domain parameter, enter at the software
- Measure the ‘Reflect’ standard using the reflector and ‘Thru’ standard only facing antenna to each other through the lens
- Save the free-space calibration

Figure 5.14 : Quarter-wave delay measurement between port 1 and port 2
At first, the frequency range is entered in the ‘Set Frequency’ tab from the ‘Define measurement’ menu of 85071 E material measurement software as shown in Figure 5.15.

In addition to frequency range, signal power, intermediate frequency band width and number of points are also defined. Number of points should be higher enough so that it prevents aliasing. Minimum number of points required to prevent aliasing = 1+ Time domain span × (Highest frequency – lowest frequency). Power and IF bandwidth should be identical to that of the power and IF bandwidth of the waveguide calibration. In the measurement, IF bandwidth has been set as low as 10 MHz to reduce noise in the measurement. Power has been set to 0 dBm and number of points used were 3201 points.

Next, reflector standard’s thickness is defined at ‘Metal plate thickness’ box from the sample holder tab as illustrated in Figure 5.16. Sample thickness, distance to sample and
the measurement technique is assigned as well. Then ‘Ok’ is pressed. Since the reflector used for the experiment is 5 mm thick aluminum flat plate, 5 mm has been put in the ‘Metal plate thickness’ box.

![Sample holder definition at 'Sample holder' tab of 85071E software](image)

**Figure 5.16: Sample holder definition at ‘Sample holder’ tab of 85071E software**

After this, calibration file of previously calibrated waveguide corresponding to the frequency range of operation is called. It brings the ‘Set time domain parameter’ window as shown in Figure 5.17. This window asks for the time domain parameters of reflector location in terms of ‘Search Start Time’, ‘Search Stop Time’, and ‘Gate Span’.
Figure 5.17: “Set Time Domain Parameters” are set to perform the free-space calibration

In order to collect the data of time domain parameters, frequency domain of the VNA is converted to time-domain for $S_{11}$ response. Then a metal plate is placed precisely in the middle of the two lenses of the free space quasi optical setup. Figure 5.18 shows a typical time domain response when the metal plate is present between lenses. The ‘Start time and Stop time’ of 110-170 GHz was .5260 ns and .7932 ns whereas it was 1.302 ns and 1.548 ns for 170-260 GHz frequency bands. After setting the time domain parameters for respective frequency bands, measurement of ‘Reflect’ standard is performed with the metal plate as shown in Figure 19.
Following the measurement of ‘Reflect’ standard, ‘Thru’ standard is measured by removing the metal plate fixture from the system as shown in Figure 5.20.
Figure 5.19: Reflect standard measurement with the metal plate

The time domain response without metal plate can be seen from Figure 5.21. It shows the response of antennas and lenses only. Since, there is no metal plate, hence

Figure 5.20: Thru standard measurement for free-space calibration
the peak is absent. After these steps, gated response isolation is applied at the calibration to reduce noise further from the calibration before saving it.

Figure 5.21: Time domain response without the metal plate ‘Thru’ standard between antennas

5.3 Simulation Procedure

Due to the unavailability of the malignant breast skin tissues, simulation has been performed to distinguish the difference between healthy measured breast skin sample to the simulated one. In order to replicate the measurement setup to simulation environment, free-space measurement setup has been created by designing mm-wave feed horn antennas. The specification and physical dimension which are required to design the antennas are gathered from antenna manufacturer. Then similar to free-space measurement system, the antennas are positioned facing each other and the material is placed between two antennas as shown in Figure 5.22.
In order to simulate the S-parameter response of material, the material needs to be selected from the material library or needs to be created. Material of breast tissue can be found from HFSS human body model [42]. HFSS provides female body model with the purchase of its full human body model as shown in Figure 5.23. While simulating a specific part of the body, the model needs to be imported to HFSS. When loading of full body model is complete, a specific part of the body can be selected from material library and can be modified to use as material in the simulation as illustrated in Figure 5.24.
In order to create the malignant breast tissue over simple breast skin, electrical properties of malignant breast tissue need to be entered in the frequency dependent model. In this

Figure 5.24: Defining malignant breast skin (a) Breast skin material selected from material library (b) Frequency dependent model selection for material under test (c) Electrical properties of malignant tissue found from published data requires in this window
simulation, published data of breast cancer has been used [39]. From the published data, the average dielectric constant over all malignant breast sample has been found 58.6, conductivity as 1.03 S/m and tangent loss .3457 at .915 GHz. Simulation is performed in two frequency bands for normal healthy tissue and malignant breast tissue. Once the simulation is finished, simulation result provides the scattering parameter response of the material. The S-parameters data are then exported as touch stone file and then called in 85071 E material measurement software to provide the permittivity of the desired simulation material.

5.4 Sample Description

In the experiment, five healthy skin samples of human body have been measured at 110-170 GHz and 170-260 GHz frequency ranges. These samples are extracted from skin of the breast and abdomen from donor during breast reduction surgery or abdominoplasty. The samples have been obtained from the melanoma tissue bank of Division of Medical Oncology at the University of Colorado Anschutz Medical Campus. Table 5.3 provides information about these samples
Table 5.3: Description of five measured healthy human skin samples

<table>
<thead>
<tr>
<th>Sample ID</th>
<th>Date Collected</th>
<th>Age at the time of collection</th>
<th>Ethnic Origin</th>
<th>Notes/Other information</th>
<th>Thickness (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>HD-28</td>
<td>8/27/2014</td>
<td>32</td>
<td>Hispanic</td>
<td>Tissue from abdominoplasty</td>
<td>23</td>
</tr>
<tr>
<td>HD-29</td>
<td>7/1/2015</td>
<td>35</td>
<td>Information not available</td>
<td>Panniculectomy large normal tissue collection</td>
<td>4</td>
</tr>
<tr>
<td>HD-31</td>
<td>9/30/2015</td>
<td>33</td>
<td>Caucasian</td>
<td>Normal skin collection from breast reduction surgery</td>
<td>4</td>
</tr>
<tr>
<td>HD-32</td>
<td>10/2/2015</td>
<td>31</td>
<td>Caucasian</td>
<td>Normal skin collection from breast reduction surgery</td>
<td>7</td>
</tr>
<tr>
<td>HD-33</td>
<td>10/8/2015</td>
<td>35</td>
<td>Caucasian</td>
<td>Normal skin collection from breast reduction surgery</td>
<td>5</td>
</tr>
</tbody>
</table>
Figure 5.25: Sample preservation (a) Sample preserved in the cryo-fridge of Dept. of Biology, UCCS (b) Sample box inside the fridge (c) Sample preserved at -78 degree Celsius
All skin samples obtained were frozen and had been brought to the facility in a temperature isolated box filled with solid dry ice. Then these samples had been preserved in the Cryo-Fridge at -78 degree Celsius before the measurement as shown in Figure 5.25.

During measurement, sample has been taken out from the specimen container with the use of forceps wearing gloves and masks. Each sample has been measured at the room temperature.

![Image of measurement specimen](image1.png)

(a) sample inside the temperature isolated box with the dry ice in plastic bag (b) five healthy skin samples from human body in the specimen container with the label at the cap

5.5 Measurements and Simulation Results

Measurements have been performed by placing the samples at the defined location between two calibrated reference planes. Measurements have been performed on 7 samples for 110-270 GHz. As VDI THz extenders come with 110-170 GHz and 170-260 GHz bands, measurements of the samples have been performed separately for the two frequency bands.
Figure 5.27: Quasi Optic free-space measurement setup that involves VNA with pair of THz extenders, pair of small mm-wave antennas and pair of lenses (a) Perspective view of the measurement setup at the laboratory (b) Human skin tissue as material under test is held by sample holder for measurement (c) Gloves and forceps used to handle sample
Before characterizing the human tissue samples, few materials with known dielectric characteristics are tested to verify the system accuracy. Here, the dielectric characteristics of air and wood have been measured before measurement of sample at each bands. Once the measurement is completed, the touchstone file (s2p) and ‘csv’ file has been saved to plot S-parameters and permittivity value of the sample under test. VNA provides the S-parameter files whereas 85071E material measurement software provides relative permittivity of the material.

5.5.1 Measurement data for 110-170 GHz

The permittivity of the air is approximately ‘1’ at 110-170 GHz. The measurement of air at 110-170 GHz provides the permittivity which tells the measurement system reads the materials accurately. The measured real and imaginary permittivity of the air are shown in the Figure 5.28.
After measurement of air, dry wood has been characterized which value is known before the measurement as 1.7 around 110-170 GHz. The real and imaginary permittivity of the dry wood are shown in Figure 5.29.
After verifying the system through the air and dry-wood measurement, the five healthy skin samples from breast and abdomen of the human body are measured. The breast skin samples are named as HD-31, HD-32, and HD-33 whereas sample from abdomen are denoted by HD-28 and HD-29. The real and imaginary permittivity of these samples are as follow.
Figure 5.30: Permittivity of breast skin tissue sample of Caucasian female across 110-170 GHz

Figure 5.31: Permittivity of breast skin tissue sample of Caucasian female across 110-170 GHz
Figures 5.30, 5.31 and 5.32 show the real and imaginary permittivity for skin of breast tissue extracted during breast reduction surgery from three different Caucasian women whose age ranges from 31 to 35. It has been found that the real permittivity of the breast skin is between 2.17 to 2.19 at the lower edge of the frequency band and it is 1.84 to 1.86 at higher edge of the band.
Figure 5.33: Permittivity of abdominal skin tissue sample of Hispanic female across 110-170 GHz

Figure 5.34: Permittivity of abdominal skin tissue sample of female (race unknown) across 110-170 GHz
Figures 5.33 and 5.34 show the dielectric properties of abdominal skin collected from two female with age between 32 to 35 years. This sample has been collected during Abdominoplasty and Panniculectomy. Donor of the sample HD-28 is Hispanic whereas race of the donor of the sample HD-29 is unknown. It is seen from those two figures that the real permittivity of the abdominal skin is between 1.90 to 1.97 at lower edge of the frequency band and 1.69 to 1.73 at higher edge of the band.

The transmission and reflection coefficients, $S_{21}$ and $S_{11}$, obtained from VNA during measurement of healthy sample are provided in Figures 5.32 – 5.37.
Figure 5.36: Measured S-parameters of healthy breast skin (HD-32) in the frequency range of 110-170 GHz

Figure 5.37: Measured S-parameters of healthy breast skin (HD-33) in the frequency range of 110-170 GHz
Figure 5.38: Measured S-parameters of healthy abdominal skin (HD-28) in the frequency range of 110-170 GHz.

Figure 5.39: Measured S-parameters of healthy abdominal skin (HD-29) in the frequency range of 110-170 GHz.
5.5.2 Measurement data for 170-260 GHz

After conducting measurement at lower frequency band of 110-170 GHz, the measurement
has been performed at 170 to 260 GHz band for all five human samples. Measurement
system is verified for this band as well similar to 110-170 GHz by measuring the
permittivity of the two known materials, air and balsa wood.

![Figure 5.40: Measured permittivity of air across 170-260 GHz frequency ranges](image_url)

Figure 5.40: Measured permittivity of air across 170-260 GHz frequency ranges
Figure 5.41: Measured permittivity of balsa wood across 170-260 GHz frequency ranges

Figure 5.42: Permittivity of breast skin tissue sample of Caucasian female across 170-260 GHz
Figure 5.43: Permittivity of breast skin tissue sample of Caucasian female across 170-260 GHz

Figure 5.44: Permittivity of breast skin tissue sample of Caucasian female across 170-260 GHz
Figures 5.42 -5.44 provide the results for breast skin sample for 170 GHz to 260 GHz. It is noticeable that the data is consistent with the previous band. The permittivity achieved at the starting frequency of this band is approximately similar to the end frequencies of the previous band which is around 1.85.

Figure 5.45: Permittivity of abdominal skin tissue sample of Hispanic female across 170-260 GHz
Figures 5.45 and 5.46 show the permittivity data of abdominal skin for 170 – 260 GHz frequency range. Similar to the data of breast skin, the measurement data of abdominal skin also conforms to the data of previous band. The permittivity of the abdominal skin measured at this band is between 1.70-1.72 for 170 GHz which is similar to the data found at higher edge of the previous band.

The following five plots from Figures 5.47 to 5.51 provide the s-parameter response of the measured human skin for 170-260 GHz ranges.
Figure 5.47: Measured S-parameters of healthy breast skin (HD-31) in the frequency range of 170-260 GHz

Figure 5.48: Measured S-parameters of healthy breast skin (HD-32) in the frequency range of 170-260 GHz
Figure 5.49: Measured S-parameters of healthy breast skin (HD-33) in the frequency range of 170-260 GHz.

Figure 5.50: Measured S-parameters of healthy abdominal skin (HD-28) in the frequency range of 170-260 GHz.
Figure 5.51: Measurement S-parameters of healthy abdominal skin (HD-29) in the frequency range of 170-260 GHz
5.5.3 Comparison Between Measurement and Simulation Data

The dielectric properties and scattering parameters acquired for healthy breast skin from the measurement has been compared with the simulated results of the healthy breast skin.

![Comparison of permittivity between measured and simulated healthy breast skin at 110 - 170 GHz](image)

Figure 5.52: Real permittivity of measured and simulated healthy breast skin at 110 - 170 GHz

Figures 5.52 and 5.53 show the comparison of the permittivity between the measured breast skins with the simulated one in two measured frequency ranges. It shows that the permittivity of simulated breast skin is similar to that of the measured skin within 2 at the lower edge of the band.
Figures 5.54 and 5.56 compare transmission and reflection coefficient between measured and simulated healthy breast skin across two measured bands. It can be seen that simulated $S_{11}$ and $S_{21}$ are in good agreement with the measured scattering parameters for both frequency bands of 110-170 GHz and 170-260 GHz.
Figure 5.54: S-parameter comparison of measured and simulated healthy breast skin at 110 - 170 GHz

Figure 5.55: S-parameter comparison of measured and simulated healthy breast skin at 170 - 260 GHz
Figures 5.56 and 5.57 show the comparison of simulated scattering parameters of healthy skin and malignant skin of breast tissues. From these two figures, it can be seen that scattering parameters $S_{11}$ and $S_{21}$ of the malignant skin tends to get separated from the healthy tissue. It can also been seen that because of high reflection, reflection coefficient increases whereas the transmission coefficient decreases.

Figure 5.56: S-parameter comparison between simulated healthy breast skin and malignant breast skin at 110 - 170 GHz
Figure 5.57: S-parameter comparison of simulated healthy breast skin and malignant breast skin at 170 - 260 GHz

Figures 5.58 and 5.59 compare the permittivity of healthy and malignant breast skin. The permittivity of the malignant breast skin is quite higher than the permittivity of the healthy breast skin and it tends to decrease at the higher edge of the band 170 to 260 GHz.
Figure 5.58: Healthy vs. malignant breast skin permittivity at 110 - 170 GHz

Figure 5.59: Healthy vs. malignant breast skin permittivity at 170 - 260 GHz
5.6 Discussions

The primary goal of this research was to diagnose cancer from human body following terahertz material characterization approach. Cancer tissue has higher water concentration than corresponding normal healthy tissue and because of having unique absorption characteristics, terahertz tends to get absorbed more at cancer tissue than healthy tissue. In order to investigate the electromagnetic wave absorption to water, dielectric properties are considered in terahertz frequency range. Material characterization method in terahertz is still being developed, therefore in order to establish effective material measurement system for human sample measurement, research began with characterizing known materials using waveguide transmission line and free-space techniques [1]. It had later been realized that measurement using waveguide-transmission line technique is not practical and feasible for the human sample in terahertz since inner dimension of the waveguide becomes extremely small. As a result, the free-space measurement system has been employed as the system for sample characterization in terahertz region. In previous study shown in Sadiq Alhuwaidi’s thesis [1], distinguishing the skin cancer tissues from the healthy tissues has been successfully demonstrated with the free-space method without the focusing lenses. However, the previous free-space measurement system had major limitation in measuring biological samples. While conducting the measurement of the sample, it was found that the system picks up more noises in the far field region and reflection coefficient of the transmitting antenna becomes larger and transmitted energy reflects back instead of communicating with the receiver antenna on the opposite side. Hence, the characterization was performed at near-field [1]. In addition to this difficulty, previous free-space system without lenses requires its antenna beam to incident in full on the sample under test in order
to avoid undesired scattering of the signal from the sample edge. With the irregular size of the sample, this condition was hard to meet. Therefore, free-space quasi optical measurement method, a modified version of the previous free-space method has been employed in this thesis work. Free-space quasi optical system can change the beam size depending on the tissue sample size and can accommodate 95 to 100% of the beam inside the sample which effectively reduces noise generated from multipath and scattering of the undesired reflected signal from the sample. Apart from this, measurement of the sample can be performed easily at the far-field region since the lenses employed refocus the antenna beam at the sample measurement location. For convenience, a software has been created which provides the distance and dimension for establishing effective and accurate measurement setup for particular sample size. With the Quasi optical free-space system, measurement has been performed on healthy breast tissue and abdominal samples. Due to the unavailability of malignant sample of these healthy tissues, a simulation of breast cancer tissue measurement has been performed. To model the cancer tissues, the dielectric properties of cancer for a known frequency is required. Dielectric properties of the breast cancer have been found from the published data, however it wasn’t found for abdominal tissue. Hence, the comparison of dielectric properties has only been performed between healthy breast skin and cancer skin. From the comparison, it has been found that permittivity of breast cancer is relatively higher than the normal healthy sample.

5.7 Limitation of Study

The measurement system used in the study only performs ex-vivo screening of the tissue. The present material characterization technique which perform in-vivo measurement only
extends up to 50 GHz [43, 44]. An appropriate thin and compact dielectric probe can be
developed for terahertz frequency ranges to establish in-vivo diagnostic system.

The absorption characteristics of terahertz to water shows that it exhibits increased
absorption characteristics with the increase of frequency within its spectrum. As a result,
in order to fully utilize terahertz sensitivity to water molecule for this study, measurement
should be performed in deeper region of the spectrum.

The screening method can identify the cancer tissues. However, it can’t define the area of
the sample affected by cancer.

5.8 Future Work

If portable terahertz source and dielectric probe for corresponding frequency ranges can be
developed, by embedding display with the material measurement software with the source,
a portable cancer diagnosis system can be developed.

At present, the material characterization can be performed up to 1.1 THz [44]. Therefore,
the measurement can be performed on cancer tissues with the corresponding tissues in other
body parts and a bank of dielectric properties can be created for the terahertz frequency
ranges.
REFERENCES


[27] Baker-Jarvis, J. Transmission/Reflection and short circuit line permittivity measurement. Technical Note NIST


propagation and applications. IEEE Press.


APPENDICES

- APPENDIX A

Quasi Optic Measurement System Configuration Software

```matlab
function varargout = Quasi_2nd(varargin)

% QUASI_2ND M-file for Quasi_2nd.fig
% QUASI_2ND, by itself, creates a new QUASI_2ND or raises the existing
% singleton*.
% H = QUASI_2ND returns the handle to a new QUASI_2ND or the handle to
% the existing singleton*.
% QUASI_2ND('CALLBACK',hObject,eventData,handles,...) calls the local
% function named CALLBACK in QUASI_2ND.M with the given input arguments.
% QUASI_2ND('Property','Value',...) creates a new QUASI_2ND or raises the existing singleton*. Starting from the left, property value pairs are applied to the GUI before Quasi_2nd_OpeningFcn gets called. An unrecognized property name or invalid value makes property application stop. All inputs are passed to Quasi_2nd_OpeningFcn via varargin.
% *See GUI Options on GUIDE's Tools menu. Choose "GUI allows only one instance to run (singleton)".
% See also: GUIDE, GUIDATA, GUIHANDLES

% Edit the above text to modify the response to help Quasi_2nd

% Last Modified by GUIDE v2.5 26-Mar-2016 03:48:12

% Begin initialization code - DO NOT EDIT
 gui_Singleton = 1;
 gui_State = struct('gui_Name', mfilename, ...
 'gui_Singleton', gui_Singleton, ...
 'gui_OpeningFcn', @Quasi_2nd_OpeningFcn, ...
 'gui_OutputFcn', @Quasi_2nd_OutputFcn, ...
 'gui_LayoutFcn', [], ... ...
 'gui_Callback', []);

if nargin && ischar(varargin{1})
 gui_State.gui_Callback = str2func(varargin{1});
end
```
if nargout
    [varargout{1:nargout}] = gui_mainfcn(gui_State, varargin{:});
else
    gui_mainfcn(gui_State, varargin{:});
end
% End initialization code - DO NOT EDIT

% --- Executes just before Quasi_2nd is made visible.
function Quasi_2nd_OpeningFcn(hObject, eventdata, handles, varargin)
% This function has no output args, see OutputFcn.
% hObject    handle to figure
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)
% varargin   command line arguments to Quasi_2nd (see VARARGIN)

% Choose default command line output for Quasi_2nd
handles.output = hObject;

% Update handles structure
guidata(hObject, handles);

% UIWAIT makes Quasi_2nd wait for user response (see UIRESUME)
% uiwait(handles.figure1);

% --- Outputs from this function are returned to the command line.
function varargout = Quasi_2nd_OutputFcn(hObject, eventdata, handles)
% varargout  cell array for returning output args (see VARARGOUT);
% hObject    handle to figure
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Get default command line output from handles structure
varargout{1} = handles.output;

function lensfocallength_Callback(hObject, eventdata, handles)
% hObject    handle to lensfocallength (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of lensfocallength as text
%        str2double(get(hObject,'String')) returns contents of lensfocallength as a double

% --- Executes during object creation, after setting all properties.
function lensfocallength_CreateFcn(hObject, eventdata, handles)
% hObject    handle to lensfocallength (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

% Hint: edit controls usually have a white background on Windows.
%       See ISPC and COMPUTER.
if ispc && isequal(get(hObject, 'BackgroundColor'),
get(0, 'defaultUicontrolBackgroundColor'))
    set(hObject, 'BackgroundColor', 'white');
end

function frequency_Callback(hObject, eventdata, handles)
    % hObject    handle to frequency (see GCBO)
    % eventdata  reserved - to be defined in a future version of MATLAB
    % handles    structure with handles and user data (see GUIDATA)

    % Hints: get(hObject,'String') returns contents of frequency as text
    %        str2double(get(hObject,'String')) returns contents of frequency as a double

    % --- Executes during object creation, after setting all properties.
    function frequency_CreateFcn(hObject, eventdata, handles)
        % hObject    handle to frequency (see GCBO)
        % eventdata  reserved - to be defined in a future version of MATLAB
        % handles    empty - handles not created until after all CreateFcns called

        % Hint: edit controls usually have a white background on Windows.
        %       See ISPC and COMPUTER.
        if ispc && isequal(get(hObject, 'BackgroundColor'),
            get(0, 'defaultUicontrolBackgroundColor'))
            set(hObject, 'BackgroundColor', 'white');
        end

    function beamwaist_Callback(hObject, eventdata, handles)
        % hObject    handle to beamwaist (see GCBO)
        % eventdata  reserved - to be defined in a future version of MATLAB
        % handles    structure with handles and user data (see GUIDATA)

        % Hints: get(hObject,'String') returns contents of beamwaist as text
        %        str2double(get(hObject,'String')) returns contents of beamwaist as a double

        % --- Executes during object creation, after setting all properties.
        function beamwaist_CreateFcn(hObject, eventdata, handles)
            % hObject    handle to beamwaist (see GCBO)
            % eventdata  reserved - to be defined in a future version of MATLAB
            % handles    empty - handles not created until after all CreateFcns called
if ispc && isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
    set(hObject,'BackgroundColor','white');
end

function antennatolens_Callback(hObject, eventdata, handles)
% hObject    handle to antennatolens (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of antennatolens as text
%        str2double(get(hObject,'String')) returns contents of antennatolens as a double

% --- Executes during object creation, after setting all properties.
function antennatolens_CreateFcn(hObject, eventdata, handles)
% hObject    handle to antennatolens (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

% Hint: edit controls usually have a white background on Windows.
%       See ISPC and COMPUTER.
if ispc && isequal(get(hObject,'BackgroundColor'),
    get(0,'defaultUicontrolBackgroundColor'))
    set(hObject,'BackgroundColor','white');
end

function samplesize_Callback(hObject, eventdata, handles)
% hObject    handle to samplesize (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of samplesize as text
%        str2double(get(hObject,'String')) returns contents of samplesize as a double

% --- Executes during object creation, after setting all properties.
function samplesize_CreateFcn(hObject, eventdata, handles)
% hObject    handle to samplesize (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

% Hint: edit controls usually have a white background on Windows.
%       See ISPC and COMPUTER.
if ispc && isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
    set(hObject,'BackgroundColor','white');
end

function rayleighrange_Callback(hObject, eventdata, handles)
    % hObject    handle to rayleighrange (see GCBO)
    % eventdata  reserved - to be defined in a future version of MATLAB
    % handles    structure with handles and user data (see GUIDATA)

    % Hints: get(hObject,'String') returns contents of rayleighrange as text
    %        str2double(get(hObject,'String')) returns contents of rayleighrange as a double

    % --- Executes during object creation, after setting all properties.
    function rayleighrange_CreateFcn(hObject, eventdata, handles)
        % hObject    handle to rayleighrange (see GCBO)
        % eventdata  reserved - to be defined in a future version of MATLAB
        % handles    empty - handles not created until after all CreateFcns called

        % Hint: edit controls usually have a white background on Windows.
        %       See ISPC and COMPUTER.
        if ispc && isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
            set(hObject,'BackgroundColor','white');
        end

    % --- Executes during object creation, after setting all properties.
    function outputbeamwaist_Callback(hObject, eventdata, handles)
        % hObject    handle to outputbeamwaist (see GCBO)
        % eventdata  reserved - to be defined in a future version of MATLAB
        % handles    structure with handles and user data (see GUIDATA)

        % Hints: get(hObject,'String') returns contents of outputbeamwaist as text
        %        str2double(get(hObject,'String')) returns contents of outputbeamwaist as a double

        % --- Executes during object creation, after setting all properties.
        function outputbeamwaist_CreateFcn(hObject, eventdata, handles)
            % hObject    handle to outputbeamwaist (see GCBO)
            % eventdata  reserved - to be defined in a future version of MATLAB
            % handles    empty - handles not created until after all CreateFcns called
function location_Callback(hObject, eventdata, handles)
% hObject    handle to location (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of location as text
%        str2double(get(hObject,'String')) returns contents of location as a double

end

function power_Callback(hObject, eventdata, handles)
% hObject    handle to power (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)

% Hints: get(hObject,'String') returns contents of power as text
%        str2double(get(hObject,'String')) returns contents of power as a double

end

% --- Executes during object creation, after setting all properties.
function location_CreateFcn(hObject, eventdata, handles)
% hObject    handle to location (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

end

function power_CreateFcn(hObject, eventdata, handles)
% hObject    handle to power (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns called

end
if ispc && isequal(get(hObject,'BackgroundColor'), get(0,'defaultUicontrolBackgroundColor'))
    set(hObject,'BackgroundColor','white');
end

% --- Executes on button press in calculate.
function calculate_Callback(hObject, eventdata, handles)
% hObject    handle to calculate (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)
f=str2num(get(handles.lensfocallength,'string'));
freq=str2num(get(handles.frequency,'string'));
w_in=str2num(get(handles.beamwaist,'string'));
s_in=str2num(get(handles.antennatolens,'string'));
sample_size=str2num(get(handles.samplesize,'string'));
%frequency to wavelength conversion
freq_GHz=freq*1e9;
lambda= ((3e8)/freq_GHz).*1000;
%Rayleigh range before the lens
zr_in=(pi/lambda)*(sqrt(2)*w_in);
%Output beam waist location
s_out = (f.*(1+(s_in./f-1)./((s_in./f-1).^2+(zr_in./f).^2))).*(-1);
%Rayleigh Range for Output beam waist
zr_out= sqrt(((1/s_in)-(1/f))*((s_out^2)-(s_out*f)));
%Output Beam Waist Size
w_out= (zr_out*lambda)/(sqrt(2)*pi);
%Total power incident on sample
f_e = (1-exp(-2*(sample_size/w_out).^2)).*100;
%Minimum Sample Thickness
thickness= lambda/20;
%-------------------------------------%
%Number to string starts here
location=num2str(s_out);
rayleighrange=num2str(zr_out);
outputbeamwaist=num2str(w_out);
samplethickness=num2str(thickness);

function samplethickness_Callback(hObject, eventdata, handles)
% hObject    handle to samplethickness (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    structure with handles and user data (see GUIDATA)
% Hints: get(hObject,'String') returns contents of samplethickness as text
%        str2double(get(hObject,'String')) returns contents of
%        samplethickness as a double

% --- Executes during object creation, after setting all properties.
function samplethickness_CreateFcn(hObject, eventdata, handles)
% hObject    handle to samplethickness (see GCBO)
% eventdata  reserved - to be defined in a future version of MATLAB
% handles    empty - handles not created until after all CreateFcns
called

% Hint: edit controls usually have a white background on Windows.
%        See ISPC and COMPUTER.
if ispc && isequal(get(hObject,'BackgroundColor'),
    get(0,'defaultUicontrolBackgroundColor'))
    set(hObject,'BackgroundColor','white');
end
APPENDIX B

Completions of the Institutional Review Board (IRB) with Collaborative Institutional Training Institute (CITI)

COLLABORATIVE INSTITUTIONAL TRAINING INITIATIVE (CITI)
CITI CONFLICTS OF INTEREST CURRICULUM COMPLETION REPORT
Printed on 11/11/2013

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INSTITUTION
University of Colorado at Colorado Springs

EXPIRATION DATE
11/13/2017

CONFLICTS OF INTEREST
COURSE/STAGE: Stage 1/1
PASSED ON: 11/14/2013
REFERENCE ID: 11699877

REQUIRED MODULES
CITI Conflict of Interest Course - Introduction 11/11/13
Financial Conflicts of Interest: Overview, Investigator Responsibilities, and COI Rules 11/11/13
Institutional Responsibilities as They Affect Investigators 11/14/13

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Paul Braunschweiger Ph.D.
Professor, University of Miami
Director Office of Research Education
CITI Program Course Coordinator
# Collaborative Institutional Training Initiative (CITI)

## Responsible Conduct of Research for Engineers Curriculum Completion Report

Printed on 11/14/2013

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**EXPIRATION DATE**

<table>
<thead>
<tr>
<th>RESPONSIBLE CONDUCT OF RESEARCH FOR ENGINEERS</th>
</tr>
</thead>
<tbody>
<tr>
<td>COURSE/STAGE</td>
</tr>
<tr>
<td>PASSED ON</td>
</tr>
<tr>
<td>REFERENCE ID</td>
</tr>
<tr>
<td>REQUIRED MODULES</td>
</tr>
<tr>
<td>Introduction to RCR for Engineers</td>
</tr>
<tr>
<td>Completing the RCR for Engineers Course</td>
</tr>
<tr>
<td>Research Misconduct (RCR-Engineering)</td>
</tr>
<tr>
<td>Whistleblowing and the Obligation to Protect the Public (RCR-Engineering)</td>
</tr>
<tr>
<td>Authorship (RCR-Engineering)</td>
</tr>
<tr>
<td>Peer Review (RCR-Engineering)</td>
</tr>
<tr>
<td>Conflicts of Interest (RCR-Engineering)</td>
</tr>
<tr>
<td>Environmental and Social Dimensions of Engineering Research (RCR-Engineering)</td>
</tr>
<tr>
<td>Mentoring (RCR-Engineering)</td>
</tr>
<tr>
<td>Research Involving Human Subjects (RCR-Engineering)</td>
</tr>
<tr>
<td>Using Animal Subjects in Engineering Research</td>
</tr>
<tr>
<td>Data Management (RCR-Engineering)</td>
</tr>
<tr>
<td>Collaborative Research (RCR-Engineering)</td>
</tr>
<tr>
<td>Introduction to the Responsible Conduct of Research</td>
</tr>
<tr>
<td>Using Animal Subjects in Research (RCR-Interdisciplinary)</td>
</tr>
<tr>
<td>The CITI RCR Course Completion Page</td>
</tr>
</tbody>
</table>

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COLLABORATIVE INSTITUTIONAL TRAINING INITIATIVE (CITI)
HUMAN RESEARCH CURRICULUM COMPLETION REPORT
Printed on 11/17/2013

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EXPIRATION DATE
11/16/2018

SOCIAL AND BEHAVIORAL RESEARCH

COURSE/STAGE: Basic Course/1
PASSED ON: 11/17/2013
REFERENCE ID: 11693975

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<tr>
<th>REQUIRED MODULES</th>
<th>DATE COMPLETED</th>
</tr>
</thead>
<tbody>
<tr>
<td>Introduction</td>
<td>11/14/13</td>
</tr>
<tr>
<td>Belmont Report and CITI Course Introduction</td>
<td>11/15/13</td>
</tr>
<tr>
<td>Students in Research</td>
<td>11/15/13</td>
</tr>
<tr>
<td>History and Ethical Principles - SBE</td>
<td>11/15/13</td>
</tr>
<tr>
<td>Defining Research with Human Subjects - SBE</td>
<td>11/15/13</td>
</tr>
<tr>
<td>The Regulations - SBE</td>
<td>11/15/13</td>
</tr>
<tr>
<td>Assessing Risk - SBE</td>
<td>11/15/13</td>
</tr>
<tr>
<td>Informed Consent - SBE</td>
<td>11/16/13</td>
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<tr>
<td>Privacy and Confidentiality - SBE</td>
<td>11/16/13</td>
</tr>
<tr>
<td>Research with Prisoners - SBE</td>
<td>11/16/13</td>
</tr>
<tr>
<td>Research with Children - SBE</td>
<td>11/16/13</td>
</tr>
<tr>
<td>Research in Public Elementary and Secondary Schools - SBE</td>
<td>11/16/13</td>
</tr>
<tr>
<td>International Research - SBE</td>
<td>11/17/13</td>
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<tr>
<td>Internet Research - SBE</td>
<td>11/17/13</td>
</tr>
<tr>
<td>Avoiding Group Harms - U.S. Research Perspectives</td>
<td>11/17/13</td>
</tr>
<tr>
<td>Research and HIPAA Privacy Protections</td>
<td>11/17/13</td>
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<tr>
<td>Vulnerable Subjects - Research involving Workers/Employees</td>
<td>11/17/13</td>
</tr>
<tr>
<td>Hot Topics</td>
<td>11/17/13</td>
</tr>
<tr>
<td>Conflicts of Interest in Research involving Human Subjects</td>
<td>11/17/13</td>
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