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Lim, Ying Ying

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High Frequency Measurement of Dielectric Properties for Packaging Materials

Lim Ying Ying

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University in
fulfillment of the requirement for the degree of
Master of Engineering

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SUMMARY

In the packaging industry, dielectric materials have numerous significant functions. For instance, substrate materials employed in electronic circuits are widely used in the industry for applications such as printed wiring boards (PWB), low-temperature cofired ceramics (LTCC), and microwave components. Furthermore new emerging packaging technologies require substrates with low permittivity and low losses. Coupled with the increase in the complexity and functionality of electronic systems, the design stage of a product development cycle has become increasingly critical. As a result, the need for well characterized dielectric materials through reliable measurements is now more crucial since the measured data would be fundamental in circuit design, signal integrity analysis as well as electromagnetic interference/compatibility (EMI/EMC) analysis.

This thesis summarizes the purpose and milestones covered in the process of carrying out the project. Essentially, the project proposes to develop and implement a suitable method to measure the complex permittivity of thin (1-3mm) packaging materials used in the microelectronics industry. Apart from being able to measure over a large bandwidth from 0.1-20 GHz and be relatively low cost and easy to implement, the technique also has to be reasonably accurate as compared to other available methods.

After a detailed literature review of the available measurement techniques, the open ended coaxial probe technique was analyzed and improved. In this thesis, a lumped equivalent circuit model was proposed using the bilayer configuration. In the model, the

interaction between the probe, material under test (MUT) and the backing of the MUT is studied and the sample backing was found to have a strong effect over the measured results, especially over high frequencies. Consequently, an optimum dielectric backing was investigated for an accurate extraction of the complex permittivity of thin packaging materials. A frequency dependent term was also included in the lumped circuit model to account for variations of the static capacitance with increasing frequency for a given coaxial line dimension. In addition, an error network was incorporated for convenient probe calibration. The validity and accuracy of the model was verified through numerical modeling and high frequency measurements (0.1-20 GHz) on well characterized dielectric materials. Furthermore, a comparison between the simulation and experimental sets of results reveal a good correlation over broadband. For the experimental verification of the improved model, the dielectric parameters of FR4-epoxy based substrates commonly used in the packaging and microelectronics industry were also restored for. Consequently, different testing conditions were imposed on the packaging laminates and the effects of the testing environment on the dielectric properties of those packaging substrates were investigated and the implications for design engineers discussed.

In addition, a comparison was made of the developed model with that of another existing model and the results discussed. Generally, the developed model yielded improved accuracy as compared to the former, without requiring a reference material that has the same order as that of the sample and be within the expected range of the sample permittivity, in order to achieve good accuracy for the restored dielectric parameters.

Summary

Furthermore, the limitations of the improved model were also discussed, such as measurement uncertainties attributed by different sample transverse shapes, the range of sample dielectric constants and thickness for which the developed model is valid and the conditions in which sample preparation is required for the material under test. Consequently, guidelines were also provided for an effective deployment of the model.

Finally, a conclusion was made based on the results obtained, along with recommendations pertaining to the future work on the dielectric measurements of thin packaging materials.

CHAPTER 1 INTRODUCTION

1.1. General Background

The electronic systems of the future would combine computing, communication, consumer and biomedical signal processing functions all in one state-of-art product. While the systems of today mainly cater to specific functions, such as telecommunications providing voice-based communications and computers performing data computations, an emerging trend comprising of "convergent systems" is fast gaining popularity as the next generation system to tap into the automotive and biomedical industries. These include high speed processors (clock frequencies above 2 GHz), broadband wireless computer networks (802.11, HiperLAN 2) and multi-band RF/microwave devices which are ushering in the new era of computing with communications convergence. Other convergent systems currently available include smart watches with cell phone and smart medical implants incorporating computing, sensing and wireless communication functions.

The technologies required to achieve the convergence of such data, voice and other functions include digital, optical, RF and MEMS concepts and this in turn leads to a strong drive for the integration of RF, digital and passive components to cater for mixed-signal integration, multi-band functionality and high frequencies. The emerging applications and trends in the next generation consumer market require highly compact and multifunctional integrated modules with efficient low-cost packages. Consequently, a new wave of packaging technologies encompassing dramatic developments in electronic packaging materials and processes are in demand to support future requirements of the wireless and

portable electronic devices. Unlike the packaging technologies of the past which houses specific components in bulky compartments, the new packaging era require the integration of multiple system functions (RF, digital, optical and sensing) in the quest for higher operating frequencies, miniaturization, low cost and a high performance packaged system. The system on package (SOP) is one such concept emerging as a viable and inexpensive alternative to current conventional package designs. Akin to Moore's Law for systems by component integration [1], SOP seeks to bridge the gap existing in current packaging limitations to attain a miniaturized, convergent package with the needed system functions of the next decade.

Nevertheless key integration challenges arise in packaging technologies. For one, higher operating frequencies translate into more sophisticated modeling required in the design stage, which is possible only with more accurate and realistic material properties. While design engineers require reliable values of the material properties in electronic packaging in order to accurately predict the performance of the system module, the conventional material data provided by the vendor is far from adequate. Thus assumptions of the dielectric constant and loss tangent (dissipation factor) of the laminates in the various layers are often made during specific designs or selection of passive and active components. However, such assumptions may not be sufficiently adequate, for the final product may not function as desired if the actual material properties in the manufactured board are different from the values assumed.

For an electronic system, the performance, reliability and cost are dependent upon the packaging materials involved, apart from the electrical design used. In particular, the

interaction of the electromagnetic energy with matter is determined by the dielectric properties of the material. In high-speed digital design, two important electrical properties of dielectric materials affect the module performance - dielectric constant and loss tangent. In signal integrity design issues, the dielectric constant directly influences the speed of the signal and the amount of cross talk between traces while the loss tangent affects the signal attenuation. In general, this complex dielectric constant contributes towards signal distortion. Consequently, being able to predict the performance for a given set of material properties is essential when specifying the right material for each application. In particular, the importance of knowing the material properties is critical, since an accurate prediction of the module performance is possible only with dependable values of the material properties available. In the emerging packaging technologies arena catering for high frequency applications, this translates to being able to determine the dielectric constant and dissipation factor of the material to at least a few times the clock frequency in the microwave frequencies (gigahertz region). Despite the demand to restore the dielectric parameters of packaging materials with greater precision, very limited literature are available for extracting the complex permittivity of such samples [2, 3]. Paulter [2] introduced a novel technique based on the TDR method to determine the dielectric constant of printed wiring board (PWB) dielectrics. Though the technique allows for fast parameter extraction and long term repeatability, measurements for the technique were only made up to 10 GHz. Furthermore, sample insertion repeatability remains an issue, particularly for thin samples, since different sample positions within the sample holder introduced variations in the measurement results. On the other hand, Heinola *et al* [3] measured for the complex permittivity of printed circuit board laminates as a function of both frequency and

temperature using a microstrip resonator with a climate chamber. Nevertheless, the insertion loss of the coupling gaps increases with frequency for the microstrip structure, which in turn decrease the coupling between the feed lines and the ring resonator. Measurements were also obtained over a limited frequency range from 1-10 GHz.

Generally, several issues and challenges exist in extracting for the dielectric parameters of a material, as the dielectric properties of a sample may be dependent on several factors like frequency, homogeneity, temperature, surface roughness and anisotropy. Despite the plethora of techniques available for material characterization, no single technique suffices to accurately characterize the material property across all frequencies and temperature while satisfying other characterization requirements such as the cost of performing the measurements and the time required for the sample measurements. In addition, challenges are posed in the measurement of thin materials (1-3 mm) employed in the current packaging technology and this is particularly so for resonant methods where there must be enough material to yield a frequency shift.

1.2. Motivation

Apart from resonant methods, some of the other commonly used techniques include the transmission line methods, the free space technique and the open-ended coaxial probe. Resonant techniques are generally applicable within a limited bandwidth and only for measuring low loss samples, while transmission line methods are destructive in nature as the MUT has to be machined into a specific size and shape prior to measurements, thereby destroying the material integrity. On the other hand, free space techniques involve costly

experiment setups prior to performing dielectric measurements. In contrast, the open-ended coaxial probe is a nondestructive, broadband technique that allows for reasonably accurate measurements to be obtained while preserving the material integrity. Consequently, an impetus arises to employ the open-ended coaxial probe technique in restoring for the dielectric parameters of packaging substrates over microwave frequencies.

1.3. Main Contribution of the Thesis

The main contributions of this thesis are as follows. Firstly, an improved technique utilizing the open-ended coaxial probe has been implemented successfully to measure the complex permittivity of thin electronic packaging materials (FR4 substrates) over a large bandwidth from 0.1-20 GHz. In addition, the developed model is relatively cheap and easy to implement, as compared to other available methods such as the free space technique. This is since minimal sample preparation is required prior to measurements. For the measurements, readily available commercial connectors were modified to obtain open-ended coaxial probes, without fabricating custom made coaxial probes that would be costly. Furthermore, reasonable accuracy (2-3%) has been obtained compared to other methods, along with measurement repeatability, such that the technique allows for reproducibility, which is an important requirement in ensuring for the consistency of the measurement data. Lastly, the nature of the developed model, which is based on a lumped element approach, enables the dielectric parameters to be computed from the measurement data in a relatively short period of time (since no numerical codes are required), as compared to computationally intensive

solutions obtained using rigorous techniques. This in turn allows for a convenient restoration of the complex permittivity of packaging materials.

In addition, an analysis of the effect of different testing conditions on the dielectric parameters of packaging laminates was carried to account for the baking and moisture absorption process taking place in the intermediate handling stages of the laminates. To date, engineers have to rely on the data provided by the vendors, which are approximate since they refer to the initial values (prior to any intermediate processing) and are given only at a particular frequency. Thus it is important to obtain more accurate values for use during the design cycle such that the system performance can be better predicted. Lastly, a procedure to investigate the measurement repeatability of the developed model as a result of probe polishing and sample fabrication tolerances has been presented in the thesis.

1.4. Organization of Thesis

Chapter 2 provides an overview of the common measurement techniques available for dielectric measurements, which includes the various milestones achieved, along with the merits and demerits inherent for each technique. In particular, the open-ended coaxial probe technique is emphasized, along with the issues and challenges that arise in utilizing the technique for dielectric measurements of thin packaging materials.

Chapter 3 introduces the formulation behind the developed bilayer model. The theoretical verification of the developed model is presented with the aid of numerical modeling. The significance of some of the main parameters in the model is also discussed. In addition, a comparison of the developed model with another existing model was made and the results discussed. Due to the limitations of the model, guidelines are also given for an effective deployment of the model in restoring for the dielectric parameters of packaging substrates.

Chapter 4 presents the measurement results in an experimental verification of the developed model, through the restoration of the dielectric parameters of well-known samples. A comparison between the simulation and measurement results obtained were also made, where a good correlation was obtained. Furthermore, the dielectric properties of some of the packaging laminates were measured using the developed model, and the effects of introducing different testing conditions on their dielectric parameters were analyzed.

Chapter 5 details the procedure for investigating the measurement reproducibility of the developed model, which is affected by factors such as probe polishing and the sample fabrication tolerances. The results obtained and corresponding implications are discussed in the chapter.

Finally Chapter 6 provides a conclusion to the entire thesis by summarizing the contributions made to the project, in addition to recommendations for future work.

CHAPTER 2 OVERVIEW OF DIELECTRIC MEASUREMENT TECHNIQUES

2.1. Introduction

Generally, the measurement techniques available for material characterization can be classified under three broad categories. These are the broadband transmission line and free space methods for measuring medium to high loss materials, in addition to resonant techniques which utilize a high quality (Q) factor in the restoration of the dielectric properties of low loss materials ($\tan\delta < 0.001$). This chapter seeks to cover some of the most common techniques available for measuring the complex permittivity of dielectric materials, as well as the motivation behind selecting a particular technique for extracting the dielectric parameters of packaging materials. The figures of the main techniques are shown in Fig. 2.1 to Fig. 2.4 [6, 7] with a summary provided in Table 2.1.

2.2. Transmission Line Techniques

Transmission line methods are widely used due to their ability to provide broadband data. Furthermore, they are the simplest among other relatively accurate techniques used to extract the dielectric properties of materials. Transmission line techniques can be broadly divided into two categories - transmission/reflection methods employing rectangular waveguides [4-5], coaxial lines and short circuit methods [6]. In addition, there are also the

Figure 2.1: Waveguide transmission line

Figure 2.2: Coplanar waveguide fixture

Figure 2.3: TE_{01} split cylindrical resonator

Figure 2.4: Free space measurement setup

coplanar waveguide and microstrip line based techniques. The measurement setups for transmission line techniques can generally be categorized under two broad classifications - where the cross section of the sample and holder is constant and where the cross section of the sample holder or sample contains discontinuities [6]. In the former, the sample must be machined precisely to fit into the transmission line (with minimum air gaps) and is thus not suitable for liquid or power samples.

In the latter category, fixtures with discontinuities include the open-ended coaxial probes, shielded coaxial lines, open-ended waveguides and open striplines. Conventionally, the dominant mode is used in the first category to treat the electromagnetic modes that decouple. On the other hand, full modal solution accounting for the various mode generation at the interfaces is usually required for fixtures with discontinuities. In the

following two subsections, the main transmission/reflection methods would be described along with their inherent advantages and limitations.

2.2.1. Coaxial Line and Waveguides

Precise measurements in this category of transmission line techniques are typically achieved by employing waveguides such as rectangular waveguides with two port measurements (shown in Fig. 2.1), coaxial lines using one port measurements [6] or short-circuit line methods, all of which assume the TEM mode. In these methods, a precisely machined sample is placed in a section of the waveguide or coaxial line and the scattering parameters measured using a network analyzer. The scattering equations relate the scattering parameters to the dielectric parameters of the sample. Nevertheless, three main problems are encountered in these techniques - air gaps, half wavelength resonance and overmoding (excitation of higher order modes other than TEM) at higher frequencies. For instance, though coaxial lines can be used for restoring permittivity and permeability measurements over broadband frequencies under the TEM mode, the main drawback stems from the presence of air gaps around the centre conductor, which results in large measurement uncertainties from the degraded measurements.

The waveguide [4-5] and coaxial line techniques employ the reflection/transmission coefficient measurements for extracting the dielectric properties of materials and the procedures adopted for the bulk of these techniques are built upon the Nicholson, Ross [8] and Weir [9] method for obtaining two port broadband measurements of the complex

permittivity, which yields explicit expressions for the permittivity and permeability as a function of the S-parameters. Nevertheless, the measurements are not stable for low loss materials at frequencies corresponding to integral multiples of one-half wavelength in the sample. In addition, while these techniques are commonly employed for their broadband capability, they suffer from other inherent limitations. For instance, precisely machined bulk samples are required to be placed in a section of the waveguide or coaxial line. Consequently such sample preparation renders transmission line techniques as destructive testing methods. In addition, small samples are required for the testing which prevents the resonance behaviour of large sheets from being studied, since higher unwanted modes would resonate in the sample (TE or TM) when the sample electrical dimensions become sufficiently large compared to the operating wavelength.

In situations where two port measurements are not feasible, another transmission line method available is the short circuit line technique, which employs one port measurements to extract for the dielectric and magnetic properties of the material under test (MUT). First introduced by Roberts and Von Hippel [10], short circuit line methods can generally be classified into the two-position technique and two-sample technique. In the former, one port scattering parameters are measured for a sample at two different positions within the sample holder. For the case of the two-sample technique, one material is machined into two samples of different lengths and the each sample is pressed against the short circuit termination to measure for the scattering parameters. The measurements may use either a fixed or movable short-circuit device and the determination of the dielectric parameters is through the solving of transcendental equations incorporating the sample length, sample

position and reflection coefficient. However, a number of uncertainties are encountered in this technique. These include sample gaps, wall and reflection losses and the location of the sample reference planes. For instance, the presence of an air gap between the sample and conductor walls creates a significant discontinuity in the E-field component of the coaxial line or waveguide and unless it is being accounted for, the measured permittivity value obtained would always be in the lower range [11]. Furthermore, uncertainties also arise from the tolerances of the network analyzer. Typically, transmission line methods are unsuitable for low-loss materials and operates better for medium to high loss materials, particularly that of magnetic materials.

2.2.2. Microstrip and Coplanar Waveguides

For measurements where the transmission lines have open surfaces which undergo radiation, a more complicated field structure is obtained. For the case of thin materials that are possibly copper cladded, it would be difficult to perform dielectric measurements using closed transmission lines. Consequently, microstrips or coplanar fixtures of the material can be adopted to extract for the permittivity of the substrate or substrate-thin film composites in a nondestructive manner. For the case of microstrips, measurements can be obtained by employing the microstrip as a transmission line or a resonator, where the complex permittivity is extracted from the resonant frequency and Q factor [12]. In adopting coplanar waveguides, as shown in Fig.1.2, usually approximate conformal mapping techniques [13] are used in extracting the dielectric parameters from the transmission and reflection coefficients obtained. In all typical transmission line methods, the loss tangent for low loss substrates cannot be accurately determined for, since the fields are primarily

normal to metal conductors. Consequently the loss of the dielectric is hard to separate from the metal, since the conductor loss is higher compared to the substrate loss.

2.3. Resonant Methods

Resonant measurement methods are among the most accurate ways of obtaining the dielectric parameters of thin materials. While some conventional resonator methods such as the re-entrant cavity [14] operate at microwave frequencies and are impractical at millimeter-wave frequencies, others such as the split-cylinder resonator, split-post resonator [14-15], whispering gallery mode resonator [16] and Fabry-Perot resonator [14] are useful for dielectric measurements in the high frequency region. An example is shown in Fig. 2.3. Resonator methods can either be classified under destructive or non-destructive measurements and in such techniques, the resonant frequency and Q factor determine the complex permittivity of the material under test (MUT). Consequently, the main drawback of the technique lies in that it is restricted to measuring samples in the low loss region. The split cylinder resonator introduced by Kent [17-18] offers nondestructive measurements of the complex permittivity of an unclad sample at microwave frequencies and is particularly useful for measuring the in plane permittivity and loss of substrates. This technique only requires a flat sample extending sufficiently far beyond the diameter of the two cylindrical waveguide section and no sample machining is needed, as shown in Fig. 2.4. Nevertheless, the technique involves a complicated theoretical model that considers the electric and magnetic fields extending into the sample region beyond the cylindrical waveguide regions. Previously the method was constrained by a limited frequency range, since only the

fundamental TE_{011} resonant mode was considered in the model. However Janezic *et al* [15] went on to present a full wave analysis for the split cylinder to attain improved permittivity measurements in 1999 and another rigorous model [19] was introduced in 2003 which adopted the mode-matching technique to extract for the dielectric parameters over a broadband frequency range. Despite being able to improve the frequency range of measurement, the main drawback lies in that the technique is computationally intensive. Furthermore, uncertainties in the sample thickness translate into uncertainties in the values of dielectric constant being restored for such resonator techniques. On the other hand, the whispering gallery method [16] is useful for measuring uniaxially anisotropic mediums apart from isotropic materials by considering higher order azimuthal modes called whispering gallery modes. In this technique, the conductor losses are eliminated and accurate loss tangent values can be achieved. Meanwhile, the Fabry-Perot resonator [14] consists of an open resonator setup that is useful for measuring the dielectric properties of low loss substrates in the millimeter wave region. A typical setup consists of two separated mirrors with a coupling aperture on one of the mirrors. Due to the open structure, such a technique suffers from a leakage of radiation, which is difficult to measure and account for physically during measurements.

Depending on the type of resonator employed, the accuracy of the measurements extracting for the complex permittivity are generally restricted by a few factors. These include uncertainties arising from the physical dimensions of the MUT and the resonator, and the presence of air gaps between the MUT and the conducting walls of the resonator [14]. Other factors such as inaccurate field solutions and the resonator wall conductivity losses

[14] also contribute to measurement inaccuracies. Generally, resonant techniques are not feasible for frequencies below 100 MHz, since the fixtures required would become large and expensive, rendering them impractical for dielectric measurements. Moreover, the samples often need to be accurately machined into a symmetrical geometry such as cylindrical shapes for techniques involving destructive measurements. Apart from destroying the material integrity, such form of sample preparation could be a costly process. Furthermore, the frequency range for which the dielectric parameters can be extracted is limited for most of the resonant techniques, since the data obtained corresponds to single or harmonically-related frequencies. Consequently increasing the number of frequency measurements would be at the expense of tuning the resonator fixtures mechanically. Furthermore there must be enough material to yield a frequency shift for resonant methods, where uncertainties in the sample thickness would translate into uncertainties in the restored permittivity. Generally, conventional resonator methods are rarely used for commercial dielectric measurement testing due to the disadvantages mentioned above.

2.4. Free Space Techniques

Free space measurements include antenna systems [20-21] and the coherent microwave transient spectroscopy [22], which uses picosecond transient radiation for optoelectronically pulsed antennas. Such approaches provide for contactless, broadband and nondestructive measurements using the transmission and reflection coefficients, which can be achieved under different temperature conditions [21]. One distinct advantage of the free space measurement system introduced by Ghodgaonkar *et al* [20] as shown in Fig. 2.4 lies in the

versatility offered for in situ characterization of non-planar [21] and cylindrical samples. Nevertheless such system utilizing an antenna setup [20-21], as shown in Fig. 2.4, can be costly. In addition, major drawbacks of such a technique include diffraction effects around the sample edges and multiple reflections from the horn lens antennas. To reduce the diffraction errors, the surface area of the MUT used during dielectric measurements have to be sufficiently large [23]. Furthermore, uncertainties arising from the sample thickness and surface roughness are not accounted for in the electromagnetic equations [24] used to compute for the dielectric parameters. The resulting systematic errors introduce could affect the measurement accuracy of the free-space technique [24].

2.5. Open-ended Coaxial Probe

Open-ended coaxial lines [25-27] have been widely used as nondestructive testing methods over broadband frequencies due to the convenience of using one port measurements to extract for the dielectric parameters, the minimal sample preparation required as well as the relatively simple setup. In using an open-ended coaxial probe for measurements, as shown in Fig. 2.5, an electrical signal is applied to a material medium via the coaxial aperture, following which a portion of the signal is reflected back due to the impedance mismatch at the junction discontinuity and measured by a vector network analyzer (VNA). The reflected signal contains the characteristics of the material that can be extracted by various approaches and models. Subsequently if the aperture admittance or reflection coefficient can be accurately represented, the coaxial probe can be adopted as a sensor for the characterization of dielectric materials [28-29]. In particular, the open-ended coaxial probe

has been commonly used to measure the dielectric properties of biological materials over broadband frequencies [27, 30]. This is since the technique does not require any tissue preparation, thus allowing for dielectric measurements to be incorporated among surgical protocols. Consequently, both *in vivo* and *ex vivo* measurements are feasible in using the open-ended coaxial probe. Apart from biomedical applications, the open-ended coaxial probe can also be adopted in process control in the form of a non contacting probe. In considering for such cases, Baker-Jarvis *et al* [31] have introduced a liftoff model based on an analytical method that considers for an air gap between the probe and the MUT.

Many approaches have been introduced to extract for the complex permittivity of the MUT, with the objective of describing the electromagnetic field at the probe aperture plane when the probe is in contact with the MUT. These include the static, quasi-static formulations and full wave solution, which employs numerical solution to an integral equation, from which the characteristics of the MUT can be found. Levine and Papas [32] pioneered the work in 1951 by considering a homogeneous medium where an air-filled co-axial line with an infinite conducting flange radiating into free space was studied. For such a model, an infinitely or semi-infinitely thick sample is considered such that the electric field at the far end of the sample would be at least two orders of magnitude smaller than that at the aperture [33]. Assuming a dominant TEM mode and that the field distribution at the aperture is proportional to $1/\rho$, where ρ is the radius, two expressions based on integral equations for the aperture electric and magnetic field were evaluated to obtain the probe aperture admittance [34]. In the 1970s, Harrison and Chang [35-36] were among a group of

Figure 2.5: Open-ended coaxial probe

authors offering complete solutions to the problem. The former employed equivalent magnetic currents and image theory to obtain an integral equation for the aperture electric field while the latter introduced a technique to solve for the integral equation through the extraction for the singularity of the kernel. In 1981, Mosig *et al* [37] used a mode-matching technique to solving an integral equation for the tangential electric field at the probe aperture.

The analysis expanded the aperture field to include terms of the dominant TEM and TM_{0n} waveguide modes. Nomograms, which are plots relating the values of the reflection coefficient to the relative permittivity of a MUT, were then generated to determine for the complex permittivity of the MUT corresponding to a particular set of reflection coefficients. Nevertheless, several nomograms would be required to account for different values of frequency as well as the various coaxial cable configurations. In addition, other rigorous techniques include the Method of Moments (MoM) and inverse point matching

presented by Nevels *et al* [38] and Grant *et al* [28] respectively. While many authors have considered only the fundamental TEM mode in their approaches, others have taken into account higher order modes. Stuchly *et al* [39] proposed a rational function model in 1994 to represent the aperture admittance of the open-ended waveguide, which is applicable to any waveguide geometry. The model is based on the moment method that includes the effects of radiation and evanescent waveguide modes. In their work, it was assumed that the probe admittance is a rational function of the square root of the permittivity frequency product, where the coefficients are determined by calculating the admittance for various dielectric materials. Nevertheless the coefficient values are valid only for limited geometries and the range of permittivity that can be computed using the rational function model is limited. Furthermore, the model is not well suited for materials with high conductivity values. In restoring for the complex permittivity, authors such as Langhe *et al* [40] and Delecki *et al* [41] have also considered formulations involving higher order modes other than the fundamental TEM mode in representing the incident field at the probe aperture.

While the abovementioned rigorous approaches are more general and yields very precise measurements, the main drawback of those techniques lies in the lengthy iterative calculations that are required, which render them unsuitable for real time dielectric measurements. In contrast to the rigorous techniques, the lumped admittance model for the open ended coaxial probe has the advantage of yielding closed form equations, by relating the aperture admittance of the open-ended coaxial line to the dielectric parameters of the MUT. In particular, Stuchly *et al* [27,42] were among the first to model the fringing fields

at the discontinuity as an equivalent lumped admittance resulting from the coaxial dielectric (independent of the material properties) and the sample fringing fields (directly proportional to the sample permittivity). A lumped capacitance model used to restore for the sample permittivity [44] can be described as shown in equation (2.1):

$$\epsilon^* = \frac{1 - \rho}{j\omega Z_0 C_0 (1 + \rho)} - \left(\frac{C_f}{C_0} \right) \quad (2.1)$$

where ϵ^* is the sample permittivity, ρ is the input reflection coefficient at the probe aperture plane, C_f represents the capacitance of the coaxial line while C_0 is the static capacitance of the open-ended probe in free space. The equivalent circuit parameters C_f and C_0 can be determined using experimental methods [43] via direct measurements or inferred from measurements on known dielectrics [28] so as to obtain the sample permittivity, along with the measured reflection coefficients. Numerical techniques employing MoM or FEM [39] have also been used to determine for the circuit parameters. On the other hand, quantitative evaluations on the experimental accuracy of the lumped circuit model have also been presented by few researchers [43-44].

Apart from using rigorous techniques and lumped admittance model, alternative approaches to restoring for the dielectric parameters using the open-ended coaxial probe include those by Burdette *et al* [30], where an antenna modeling theorem introduced by Deschamps [45] was considered for the open-ended coaxial probe. On the other hand, Misra *et al* [46] employed a variational formulation that was simplified to consider a quasi-static case. Time domain reflectometry (TDR) techniques introduced by Cole *et al* [47] were also considered where among the authors [2, 7], Garcia-Ruiz *et al* [7] introduced a calibration model based

on TDR principles which allows for the broadband dielectric constant extraction of well-known materials using the open-ended coaxial probe. Other variations to the model include those introduced by Baker-Jarvis *et al* [48], who employed a model to measure for thin films and materials by implementing two coaxial probes to form a two port test device. In that model, the MUT is placed between the probes. By approaching the problem in the same way as a typical one port coaxial probe technique, an exact solution was used to restore for the permittivity of the MUT from the data obtained.

To employ the open ended coaxial probe model experimentally, various design rules for the use of the open-ended coaxial line has been provided by Langhe *et al* [49], which includes the extent to which a conducting flange can be considered infinite in the radial direction. Other milestones include using the coaxial probe for high temperature measurements [50] following which the dielectric parameters of a material can be extracted with more versatility in terms of both frequency and temperature.

One main drawback of the coaxial probe technique arises from the presence of air gaps, which can lead to significant measurement errors. The electric fields at the probe aperture consist of both the normal and radial components and air gaps arising between the probe and MUT would result in a discontinuity of the normal electric field, thereby introducing errors in the measured dielectric parameters. For the case of non-planar solid samples, a single flat and smooth surface is required for contact with the probe to minimize the effect of air gaps. To achieve that, some form of sample preparation is required, such as machining the sample until it is coplanar with the probe face. Consequently, the coaxial

probe technique has been used more extensively for liquids and soft solid materials (biological materials and powders) where good contact can be assured for to avoid the introduction of air gaps. The different measurement techniques discussed are summarized in Table 2.1 with their important characteristics highlighted.

2.6. Issues and Challenges in Employing the Open-ended Coaxial Probe

As each technique has its own niche as well as inherent limitations, the optimal measurement method to be adopted should yield a broadband yet non-destructive extraction of the dielectric parameters of the MUT. In addition, the technique has to be reasonably accurate and allow for easily reproducible measurements, along with minimal sample preparation. Among the various methods available, the open-ended coaxial probe is an approach that allows for fast, reliable extraction of dielectric parameters. Another unique advantage of the technique lies in its open structure that facilitates non destructive testing, thus allowing the material integrity to be maintained. Apart from providing broadband data, as compared to the open-ended waveguide techniques the approach requires minimal sample preparation as compared to other measurement techniques such as resonant methods, and a simple measurement setup suffice since only the reflection coefficient of the coaxial probe in contact with the MUT is required for data interpretation. However three main challenges exist in adopting the technique to restore for the dielectric parameters of finite samples - determining a suitable model to restore for the sample dielectric parameters, considering for probe calibration to ensure measurement accuracy and alleviating for the air gaps inherent between the probe and the MUT.

Table 2.1: Comparison of dielectric measurement methods

Technique	Field	Advantages	Disadvantages
Transmission line methods	TEM, TE ₁₀ , TM ₀₁	Broadband, complex permittivity & permeability	Sample machining, air gaps
Cavity	TE ₁₀	Very accurate	Narrowband, low loss
Dielectric Resonator	TE ₁₀	Very accurate	Low loss samples
Open-ended probe	TEM, TM ₀₁	Nondestructive	Insufficient contact between probe and sample
Free space methods	Quasi-TEM	Nondestructive	Expensive, interfering signals from the surroundings.

There are two basic approaches for the determination of complex permittivity from the measurements of the coaxial line open-circuit reflection coefficient – a rigorous solution of the electromagnetic field equations and a lumped equivalent circuit representing the probe’s fringing fields. In particular, the latter has been particularly useful due to the convenience in dielectric parameters extraction being offered, in contrast to rigorous approaches that are generally computationally intensive. In restoring for the sample permittivity, a balance has to be achieved between obtaining accurate solution resolution while fulfilling the other characterization criteria. Subsequently, the relative simplicity of the equivalent admittance

circuit to represent the fringing fields as capacitances is preferred over the rigorous solution since a closed form expression for the sample permittivity can be obtained allowing for the dielectric parameters to be restored with on line measurements and minimal data processing time. Nevertheless, theoretical formulations for the open-ended coaxial probe assume that the MUT extends to infinity, which is impractical when considering for finite samples. Due to the need to measure for thin dielectric materials, the applicability of employing the open-ended coaxial probe for such samples requires further investigation. Available but limited literature had considered cases of open-ended coaxial lines radiating into layered media [25-26, 51]. Jiang *et al* [25] introduced an empirical formula based on an exponential fit for samples with finite thickness in a lumped element model. Consequently, the impedance of the coaxial probe was related to the dielectric parameters of the MUT. However the method depends on the availability of a range of thicknesses for a particular material, and dielectric measurements were reported within a limited frequency range of 4-8 GHz [25]. Similarly, Chen *et al* [51] considered an equivalent circuit for a bilayer dielectric configuration which allows for the determination of the permittivity of either layer without the knowledge of the first layer medium, where the analytical expression was determined using the conformal transformation method. Nevertheless, the radiation conductance was not accounted for in the simplified model and measurements only up to the low microwave frequencies were reported (0.6-2.6 GHz) using the equivalent circuit. Subsequently, motivations arise in utilizing the lumped element model to extract for the dielectric parameters of finite samples (1-3 mm) over broadband up to 20 GHz.

In measuring for solid samples, flat planar substrates are required such that good contact with the probe can be achieved with minimum air gap effect, which could result in significant measurement errors, particularly for thin samples. Jarvis *et al* [31] studied the lift-off effect for the open-ended coaxial probe while other authors [40, 52] sought to account for the influence of air gaps by considering multi-layered mediums. Apart from considering for the air gap effects through theoretical models, Arai *et al* [53] suggested performing sample preparation for the MUT in the form of sample polishing, such that air gaps arising from the inherent surface roughness of the MUT are alleviated.

In conclusion, the main objective of this project seeks to extract for the dielectric parameters of packaging substrates over a broadband frequency range by employing the open ended coaxial probe technique. Overall, there is an incentive to develop an improved open-ended coaxial probe model to maintain the merits of the technique allowing for non destructive dielectric measurements to be performed over broadband in a convenient manner. On the other hand, the existing disadvantages and constraints have to be minimized. This include applying the model for thin packaging materials, calibrating the measurement setup to account for the additional electrical length being introduced by the probe and minimizing the air gaps inherent between the probe and the MUT. These issues and challenges are addressed in this thesis, with the objective of achieving measurement accuracy and repeatability.

CHAPTER 3 SUGGESTED BILAYER MODEL

3.1. Introduction of the Developed Bilayer Model Based on a Lumped Element Technique

From Chapter 2, the impetus to perform dielectric measurements for thin packaging samples was discussed, and the open-ended coaxial probe technique was selected over other techniques as it is broadband, non-destructive, and offers measurement convenience in a simple setup. In particular, the lumped circuit approach [25-26, 51] was preferred over rigorous techniques due to the relative ease of computation for real time measurements in using the open-ended coaxial probe method (no transcendental equations involved). In considering a lumped element model to measure for thin samples, a sufficiently thick dielectric medium could be used to back the MUT, such that the electric field at the far end would be at least two orders of magnitude smaller compared to the field at the probe aperture plane [44]. Consequently, the underlying assumptions for an “infinite” MUT can then be satisfied using the open-ended coaxial probe technique. The general bilayer model along with the equivalent circuit being considered in the suggested model is shown in Fig.

3.1.

Figure 3.1: Bilayer lumped circuit model and the equivalent circuit

With reference to Fig. 3.1, the MUT is backed by an arbitrary medium of semi-infinite thickness. The coaxial line is considered to have an infinite flange extending in the radial direction, while the MUT is considered to be linear, isotropic, homogeneous and non-magnetic in nature. It is further assumed that only TEM mode fields exist at the probe aperture. Z_0 is the characteristic impedance of the coaxial line and C_f is the capacitance inside a Teflon-filled coaxial line. C_{01} represents the capacitance due to the fringing field outside the coaxial line into the finite sample while C_{02} represents the capacitance of the fringing field into the infinitely thick medium that is used to back the sample. Previously,

Jiang *et al* [25-26] introduced an empirical equation relating the dependence of the sample dielectric constant on the finite thickness in a bilayered model, where [25]:

$$\varepsilon_{eff} = \varepsilon_1 + (\varepsilon_2 - \varepsilon_1)e^{-x/D} \quad (3.1)$$

ε_1 and ε_2 are the dielectric constant values of a low loss MUT (loss tangent $\varepsilon_1'' \approx 0$) and the infinite medium (dielectric backing) in Eqn. (3.1). x is the thickness of the MUT and D represents an empirical parameter with dimensions of length [25]. To determine the unknown parameters ε_1 and D , the values of ε_{eff} , x and ε_2 have to be first ascertained in the model by Jiang *et al* [25]. This is achieved by plotting the quantity $\ln(\varepsilon_{eff} - \varepsilon_1)$ against x from Eqn. (3.1), where ε_1 is considered as an adjustable parameter such that the best linear fit can be obtained. Consequently, ε_1 can be found using Eqn. (3.1) after D is computed from the slope of the linear plot. The sample dielectric parameters were extracted over a limited frequency band of 4-8 GHz by Jiang *et al* [25]. A significant drawback of the technique lies in that it is dependent upon the availability of a range of thickness in using a particular material to generate the appropriate dependence of ε_{eff} on x which may not always be possible. In such instances, it could be costly to obtain a range of varying thicknesses from the vendors. Another disadvantage is that the MUT may have to be backed with three or four materials of known dielectric constants such that they are thick enough to be considered a semi-infinite sample assumed for the open-ended coaxial probe model. Consequently, a challenge lies in obtaining several materials with well known dielectric constants which would be stable over the frequency range of operation considered. Overall, these requirements introduce restrictions in performing dielectric measurements apart from increasing the measurement costs.

In another common model [27,44,54] using the lumped element approach, Athey *et al* [44] suggested incorporating frequency dependent terms for the capacitance model to be valid at higher frequencies, since the capacitance of the open-ended coaxial line in air, C_0 , would increase as a result of evanescent modes excited at the probe discontinuity. In this suggested model, the frequency dependent term Af^2 [44], is included in the lumped circuit to account for the frequency dependent variations of C_0 for a given coaxial line dimension. The corresponding expression is of the form:

$$C(f) = C_0 + Af^2 \quad (3.2)$$

where A is a constant dependent on the coaxial line dimensions and f is the signal frequency. In the suggested model, equation (3.2) was considered together with the model by Jiang *et al* [25]. By incorporating a frequency dependent term for C_0 into equation (3.1), the model by Jiang *et al* can then be extrapolated into the high frequency range, while accounting for the evanescent modes excited at higher frequencies, where:

$$C_{01} = (C_0 + Af^2)(1 - e^{-x/D})\epsilon_1 \quad (3.3)$$

$$C_{02} = (C_0 + Af^2)\epsilon_2 e^{-x/D} \quad (3.4)$$

$$C_T = C_f + C_{01} + C_{02} \quad (3.5)$$

C_T is the total terminal capacitance. From the expressions, when $x \rightarrow \infty$, $C_{01} \rightarrow \epsilon_1 C_0$ and $C_{02} \rightarrow 0$. Similarly when $x \rightarrow 0$, $C_{01} \rightarrow 0$ and $C_{02} \rightarrow \epsilon_2 C_0$, thus satisfying the requisite boundary conditions for the case when the thickness of the MUT tends to semi-finite and when it approaches zero, which results in the fringing fields being concentrated in the second layer. The suggested model is valid at frequencies for which the line dimensions are small $a/\lambda \ll 1$ (more specifically $a/\lambda < 0.04$) [33] compared to the wavelength, such

that a parallel conductance term to account for losses due to radiation can be considered negligible. In the condition, a represents the inner conductor radius of the coaxial line. From the lumped element model, the impedance of the terminal load Z_T as shown in Fig. 3.1 can be represented as:

$$Z_T = \frac{-j}{\omega C_T} \quad (3.6)$$

By substituting Eqn. (3.5) into Eqn. (3.6), and expressing the reflection coefficient of the open-ended coaxial Γ in terms of its normalized admittance Y_T , where:

$$\Gamma = \frac{1 - Y_T}{1 + Y_T} \quad (3.7)$$

the impedance of the probe can be related to the permittivity of the MUT with some manipulation, where:

$$\varepsilon_1 = \left(\frac{1}{1 - e^{-x/D}} \right) \left[\left(\frac{1}{j\omega Z_0 (C_0 + Af^2)} \right) Y_T - \left(\frac{C_0}{C_0 + Af^2} + e^{-x/D} \right) \right] \quad (3.8)$$

However, to maximize the accuracy of the impedance measurements from the reflection coefficients using the VNA, the probe needs to be calibrated from the end of the coaxial line to the probe aperture plane, which is in contact with the MUT. Consequently, a two port error network was included in the suggested model to account for the de-embedding process from the end of the coaxial line to the probe aperture plane. The true reflection coefficient Γ can be related to the measured reflection coefficient by the relationship:

$$\Gamma = \frac{\rho - E_{11}}{\rho E_{22} - (E_{11}E_{22} - E_{12}E_{21})} \quad (3.9)$$

where E_{11}, E_{12}, E_{21} and E_{22} belong to the error scattering matrix $\mathbf{E}$ and ρ is the measured reflection coefficient. By substituting Eqn. (3.9) into Eqn. (3.8), the final expression for

the permittivity of the MUT incorporating the error network in the suggested model can then be expressed as:

$$\varepsilon_1 = \frac{g_1(f, x, D, \varepsilon_2)\rho + g_2(f, x, D, \varepsilon_2)}{g_3(f, x, D, \varepsilon_2)\rho + 1} \quad (3.10)$$

$$\varepsilon_1 = \frac{a\rho + b}{c\rho + 1} \quad (3.11)$$

where a , b and c are complex coefficients that are functions of frequency, sample thickness, empirical parameter D and the dielectric backing used, as shown in Eqn. (3.10). In particular, the significance of the dielectric backing on the extracted dielectric parameters would be investigated in section 3.3.1., and an emphasis on the use of an optimum dielectric backing would be subsequently highlighted for achieving accurate dielectric measurements. To extract for ε_1 from Eqn. (3.11), three standards are required to determine for a , b and c by measuring three materials with known dielectric properties and solving for the corresponding simultaneous equations. Compared to the model by Athey *et al* [44] presented in Eqn. (2.1), where:

$$\varepsilon^* = \frac{1 - \rho}{j\omega Z_0 C_0 (1 + \rho)} - \left(\frac{C_f}{C_0} \right)$$

no determination of the intermediate parameters C_f or C_0 is required, which in turn facilitates the implementation of real time measurements. In their work, C_f and C_0 were calculated by Athey *et al* [44] through two methods of experimental evaluation. One of the techniques involves placing dielectrics with well-known properties in contact with the coaxial probe. However, a drawback lies in that the permittivity of the medium also has to satisfy the optimum capacitance condition [44], otherwise large errors would occur.

Furthermore, the determination of C_f requires the use of a low dielectric constant material, which would degrade the measurement accuracy [44]. Moreover, an iterative procedure is required to obtain the final values of C_0 and C_f from the initial first approximations, which increases the overall computation time to extract for the permittivity of the MUT. By comparison, not only does the developed model not require the values of C_0 and C_f to be determined, no specific condition is required relative to Athey's model [44]. In addition, the procedure of extracting for the permittivity of the MUT is relatively fast and convenient, as described above and in Appendix I.

To demonstrate the robustness of the model, a comparison of the developed model with a commonly referenced model would be made in Section 3.4. The model would also be validated through an experimental implementation and the results reported in Chapter 4.

3.2. Numerical Modeling using High Frequency Structure Simulator (*HFSS*)

Prior to fabricating the measurement setup, performing a theoretical verification is an important prerequisite in emulating for laboratory measurements. For electromagnetic models, numerical modeling using Ansoft *HFSS* software serves as a useful prototyping tool in validating the effectiveness of the model while generating for the theoretical results. *HFSS* is a full wave electromagnetic field simulator that employs the Finite Element Method (FEM) to solve for Maxwell's equations from which S parameters can be computed. The FEM solution generated by *HFSS* also enables the visualization evaluation and analysis of field components during the post processing stage, thereby allowing one to

gain more insight into the electromagnetic behaviour of the geometrical structure modeled. Furthermore, such 3D simulators offer versatility in fine-tuning the model within a faster cycle time, at no extra costs as compared to modifications made in the post fabrication stage. The steps involved in the modeling process are shown in Fig. 3.2 and described as follows:

1. The physical structure used in the numerical modeling is drawn using geometrical 3D objects, as shown in Fig. 3.3(a). In addition, an airbox is drawn to enclose the entire structure and define for the problem region, indicating the region in which the structure is to be solved for and the meshes be refined. Defining an airbox reduces the problem boundary and results in a more efficient computation process particularly since the FEM technique involves a rigorous matrix approach, leading to a computationally intensive solution process. Furthermore, the actual full model was split at the XZ and YZ plane as shown in Fig.3.3 (a), such that a quarter model shown in Fig. 3.3(b) remains, which reduces the complexity of the overall structure and improves the solution time with fewer computer resources. Consequently, the quarter model is used throughout the entire simulation processes and such a model suffices, since the discrepancy of the S parameters obtained between simulating for a full model against that of a quarter model is negligible, with a deviation of within 1%, as shown in Fig. 3.3(c).
2. Next, material properties (electrical properties, permittivity and permeability) are assigned to the objects in the modeled 3D structure in Fig, 3.3(b). Each object is assigned material properties representing the actual material being modeled from an

existing database in *HFSS*. Alternatively, new materials can also be created and added to the list of materials available.

3. To allow energy to flow into the structure, a wave port is defined on a planar surface in the structure. The ensuing structure from the port that functions as a waveguide is assumed to have the same cross section and material properties as the port defined. The 2D field patterns generated in the port serve as boundary conditions for the full 3D problem. Consequently generalized S parameters can be obtained for further analysis. For the open-ended coaxial probe model, only the reflection coefficients (S_{11}) are computed for, since the structure is considered as a one-port network. For the quarter structure being modeled, perfect H (magnetic field) symmetry boundaries were defined, which forces the electric field to be tangential to the surface in considering for a TEM mode in the structure in Fig. 3.3(b). Radiation surfaces were defined at the remaining sides of the airbox to act as the outer radiation boundaries to allow emanating waves from the structure to radiate infinitely into space, without reflections from the boundaries.
4. Next, the adaptive meshing criteria are defined. In adaptive meshing, the mesh is automatically refined in the critical regions to yield the most efficient mesh. By generating a solution based on a coarse initial mesh, the mesh in areas of high error density is refined and a new solution generated. Generally the number of tetrahedras added during each round of iteration in the adaptive refinement process is determined by the tetrahedra refinement, which results in the mesh growing by a certain percentage

Figure 3.2: Numerical modeling process using *HFSS*

in each pass. This is to ensure that the mesh is sufficiently perturbed and the desired convergence attained. After each adaptive pass, the S_{11} values of the current mesh with that of the previous mesh would be compared to determine if the convergence criteria has been achieved. The default value of tetrahedra refinement used is 20% and the maximum delta S defined is 0.5%, where delta S represents the change in the magnitude of S_{11} between two consecutive passes and serves as a stopping criterion for the adaptive process. Once the magnitude and phase of S_{11} change by an amount less than 0.5% value between the current iteration to the next, the adaptive analysis ends. Otherwise, the adaptive cycle continues until the requested number of passes is completed.

5. Upon attaining the specified adaptive criteria, a discrete frequency sweep was performed to obtain a set of S_{11} within the frequency range of interest from 0.1-20 GHz. *HFSS* uses the finite element mesh refined during the adaptive solution in the previous stage as the solution frequency. For the suggested model, the solution frequency was set at the upper frequency of 18 GHz. The frequency sweep was performed using the current adaptive mesh, yielding a full solution at every frequency point. Consequently the total solution time required is the time taken for a single frequency solution multiplied by the number of frequency points.
6. In the post processing stage, the simulation results can be analyzed through plots of the various field overlays, or the plot animation of the field quantities. The results of the S parameters generated can also be plotted for further analysis.

Figure 3.3: (a) Full model in *HFSS* (b) Quarter model simulated in *HFSS*
(c) Comparison of S_{11} plots after simulating the full and quarter model

3.3. Analysis of Results from Numerical Modeling

3.3.1. Verification of the Suggested Bilayer Model and the Significance of the Dielectric Backing

Using the lumped circuit approach, the open-ended coaxial line technique is analyzed and a bilayer model for a finite sample was developed as detailed in Section 3.1. In particular, the interaction between the probe, MUT and the medium backing the MUT is studied and described in this section, since the type of dielectric backing employed is a main factor in the restoration of the sample complex permittivity, as observed from Eqn. (3.10). Due to the finite thickness of the MUT, the significance of the medium backing in the second layer cannot be undermined, as a significant proportion of the maximal electric field from the probe aperture would be present in the dielectric backing. Consequently, the effect of employing different dielectric mediums as backings is investigated in this section. In accordance to the design rules set out by Langhe *et al* [49] that a sufficiently infinite MUT fulfill the criteria $A/\pi b^2 > 2$, where A is the sample surface area and b is the outer radius of the coaxial probe. The transverse dimensions adopted for both the MUT and sample backing in the numerical modeling was 25 x 25 mm, which follows the typical dimensions of actual samples. The probe diameter was 4.12 mm corresponding to the probe configuration of a typical SMA connector. The sample dimensions were found to suffice as slight deviations of within 5% were observed when the transverse dimensions were doubled to 50x50 mm. In addition, the thickness of the sample backing is 10 mm, such that it can be considered infinitely thick when the electric field at the bottom of the sample is at least two orders of magnitude smaller than that at the probe aperture [33].

With reference to the simulated S_{11} plots in Fig. 3.4, different mediums of varying dielectric constants ($\epsilon' \sim 1-10$) were considered as sample backings for a MUT with a predefined dielectric constant ($\epsilon' = 3$) and zero loss tangent. Fig. 3.5(a) and Fig. 3.5(b) further illustrate the simulation results obtained in using Eqn. (3.11) to extract for the dielectric parameters of the MUT. From Fig. 3.5(a), restored dielectric constants close to the value of 3 were obtained using Eqn. (3.11) of the developed model, with an accuracy of 3% up to 8 GHz. Consequently, it can be observed from the S_{11} plot that any dielectric backing below 8 GHz suffices to restore for the dielectric parameters of the MUT. Beyond the cutoff frequency of 8 GHz, however, the type of dielectric backing employed affects the measured reflection coefficients sensed by the coaxial probe significantly. Consequently, the emphasis is on the use of an optimum sample backing for a reliable extraction of the of the finite MUT dielectric properties over a broadband range from 0.1-20 GHz. From the results, foam, which has the lowest dielectric constant ($\epsilon' \sim 1.03$), was found to provide for the optimum backing.

Foam ($\epsilon' \sim 1.03$), serves as the ideal medium for backing the MUT since its dielectric constant approximates that of air. Consequently, the electromagnetic energy emanating from the probe aperture would be dissipated into free space and no reflections would result from the finite wall boundaries of the mediums in the two layers. In comparison with other dielectric backings, significant fluctuations were observed in the dielectric properties of the MUT over increased frequencies when materials with higher dielectric constants are used to back the thin sample. Essentially, the sample backing was found to have a strong effect over the measured results, especially over high frequencies. At lower frequencies, the effect

Figure 3.4: Simulation results of S_{11} (Mag) plots obtained using different dielectric backings

Figure 3.5: Simulation results of (a) Dielectric constant ϵ' (b) Loss tangent ($\tan \delta$) obtained using different dielectric backings.

of surface waves is not significant, since the total electromagnetic field is localized around the probe aperture, following which the power carried away by surface waves are negligible. Over higher frequencies of operation, however, surface waves arising from radial modes would be introduced at the probe discontinuity plane. Apart from the number

of modes being excited at a particular frequency, the propagation of surface waves is also strongly dependent on the physical parameters of the layered material. In particular, the propagation of such higher order modes are influenced by the dielectric constant values of the mediums which determine the interaction of the signal with the material, the thickness of the mediums as well as the boundary conditions. This is also evident from the results in Fig. 3.5 (a) and Fig. 3.5 (b), where significant fluctuations of the dielectric parameters were observed when sample backings with higher dielectric constants were used. In addition, the semi-infinite thickness of the sample backings at 10 mm exacerbated the propagation of the surface waves over higher frequencies as the leaky waves are kept contained within the dielectric medium. Consequently, these factors contribute to measurement inaccuracies in restoring for the dielectric parameters.

3.3.2. Significance of Sample Transverse Dimension

Having investigated the significance of using different mediums to back the MUT on the dielectric properties extracted, this section focuses on the importance of the sample transverse dimension with respect to the propagation of surface waves and how effective characterization could ensure for a reliable extraction of the dielectric parameters. For the open-ended coaxial probe, there are generally three ways in which the incident wave can be distributed. While most of the energy would be reflected, part of it would be radiated with the remaining energy expended on the excitation of surface waves, if the conditions for the propagation of such waves are fulfilled as discussed in the earlier section.

With respect to the MUT in the open-ended coaxial probe, the transition region from the probe aperture into the MUT is assumed to extend infinitely in the radial direction such that evanescent modes excited at the junction discontinuity would decay significantly, resulting in a dominant TEM mode incident at open end of the probe. The evanescent modes attenuate exponentially with distance from the point of excitation at the probe aperture, and can be considered negligible at a distance comparable to the transverse dimension of the coaxial line [55]. This implies that the sample transverse dimension should be large enough such that reflections from the boundaries arising from evanescent modes are sufficiently negligible to be detected by the probe. However in a practical situation, the finite samples to be measured may not satisfy the above prerequisite. Furthermore the reflections at the edge of a finite material layer may be neglected only if the material is sufficiently lossy that the surface waves decay sufficiently before reaching the edge of the material layer, which may not be applicable for low and medium loss materials. In such cases, there exists a need to investigate the effects of varying transverse dimensions and sample thicknesses on the propagation of evanescent modes for the open-ended coaxial probe model. In the suggested model considering for finite samples, the lateral dimensions of the media acting as the sample backing define the boundary conditions. These finite transverse dimensions could support the propagation of surface waves over higher frequencies, as the waves are reflected from the finite sample sidewalls back to the probe aperture, resulting in a perturbation of the measured reflection coefficient and measurement inaccuracies.

Consequently, the effects of varying transverse dimensions and thickness of the sample backing on the propagation of surface waves is investigated in this section to ascertain the

effective frequency range for which the dielectric parameters of the MUT can be extracted. Samples ($\epsilon'=6$, $\tan \delta=0$) having varying square transverse dimensions were studied first and the investigation was later extended to samples with rectangular transverse dimensions. Numerical modeling using *HFSS* was performed for the characterization. The sample thickness considered ranged from 2mm to 5mm while the transverse dimensions were varied according to (a) 8x8 mm (b) 16x16mm and (c) 24x24 mm. Foam backings were used to back the MUT in the numerical modeling and the results are shown in Fig. 3.6 (a)-(c). The values of the S_{11} parameter are analyzed here with respect to a certain cutoff frequency (frequency beyond which higher order modes are excited), since the sample complex permittivity can be restored from the input impedance (Z_{in}) in the open-ended coaxial probe technique, where Z_{in} is a function of S_{11} from Eqn. (3.7).

From Fig. 3.6(a), it can be observed that using samples that are too small (8x8mm) could perturb the measured reflection coefficient due to reflections at the finite sample boundaries. Subsequently higher reflection coefficients would be obtained, due to the stronger reflections arising from the smaller sample boundaries, which are detected by the probe sensor. As the sample thickness is increased from 2 mm to 5 mm, the corresponding cutoff frequency declined from 14 GHz to 9 GHz and it can be deduced that an increase in the sample thickness exacerbates the support of propagating modes, following which the cutoff frequency for the dominant TEM mode declines. Below the cutoff frequency, any transverse dimension suffices for performing dielectric measurements. Beyond the cutoff

(a)

(b)

-(c)

Figure 3.6: S_{11} plots for sample thickness of (a) 2mm (b) 3mm (c) 5mm (square) (d) 5mm (rectangular) with varying transverse dimensions.

frequency however, the sample transverse dimension becomes a significant factor in determining the frequency limit for which the developed model can be used to extract for the dielectric parameters. Consequently, the sample thickness has to be reduced with an increase in the operating frequency, so as to prevent the onset of propagating evanescent modes in the coaxial probe structure.

In addition, rectangular samples were studied with transverse dimensions of (d) 8x30 mm (e) 16x30 mm and (f) 24x30 mm. Comparing Fig. 3.6(c) with Fig. 3.6(d), no significant increase in the cutoff frequency was observed despite using a sample with a larger lateral dimension on one side. Consequently, it is deduced from the S_{11} measurements that the open-ended coaxial probe can be used for dielectric measurements for any sample configuration, using either the rigorous or lumped element technique, up to a frequency of 8 GHz, beyond which propagating modes arise. On the other hand, the inclusion of a frequency dependent term in Eqn. (3.2) for the developed model considered the evanescent

modes introduced across higher frequencies, thus enabling the dielectric parameters to be extracted up to 20 GHz, as shown in Fig. 3.5.

In conclusion, a full wave simulator was employed to ascertain conditions under which the lateral dimensions of the MUT can be disregarded, following which measurement uncertainties can be accounted for and removed in cases where samples of different lateral dimensions are considered.

3.4. Comparison of the Developed Model with an Existing Model

With the developed model verified through simulation results and the significance of different dielectric backings investigated, a comparison was subsequently made between the developed model and that of the bilinear calibration model by Garcia-Ruiz *et al.* [7] through numerical modeling in *HFSS*. Having been widely referenced in literature [56-60], the bilinear calibration model serves as a reasonable benchmark for comparison with the developed model. Originating from the work by Cole *et al.* [47], the bilinear calibration model is a robust technique and has been frequently used to obtain reflection measurements in open-ended coaxial cells [47,61] and probes [56-60]. The model arises from an extension of the bilinear transformation in microwave circuit theory, where [47]:

$$\varepsilon^* = \frac{A\rho + \varepsilon_{ref}^*}{1 - B\rho} \quad (3.12)$$

and

$$\rho = \frac{\Gamma_{ref} - \Gamma_x}{\Gamma_{ref} + \Gamma_x} \quad (3.13)$$

A and B are calibration constants that are calculated using two reference materials while ϵ_{ref}^* and Γ_{ref} are the permittivity and reflection coefficient of a known reference material. In addition, ϵ^* and Γ_x are the reflection coefficient and permittivity of the MUT. While linear equations in Eqn.(3.12) and Eqn.(3.13) are obtained that allows for a convenient restoration of the dielectric parameters, the main drawback of the model lies in the accuracy of the permittivity that can be restored, as it is influenced significantly by the value of the reference dielectric. To achieve good accuracy, the dielectric constant of the reference material employed must be the same order as that of the sample and be within the expected range of the sample permittivity.

In the comparison, the dielectric parameters of a well-known material, Rexolite ($\epsilon'=2.55$, $\tan \delta=0.001$), were restored for using both models. The model by Garcia-Ruiz *et al.* [7] considered solid samples prepared in 20x20x10 mm dices while for the case of the developed model, samples of 20x20x1 mm were modeled along with foam backing of dimension 20x20x10 mm to emulate for a semi-infinite medium. In both models, air, Teflon and Chlorofoam were used [7], with Teflon ($\epsilon'=2.1$) as the reference materials for the bilinear calibration model and the three mediums used as calibration materials for the developed model. In the numerical modeling, the dimensions of a standard SMA (4-hole flange mount) connector were used, with $2a = 1.27$ mm and $2b = 4.12$ mm where a is the radius of the inner conductor and b is the radius of the outer conductor.

The dielectric parameters obtained from both models are shown in Fig. 3.7 and Fig. 3.8. The significant variations observed between 0-2 GHz in the dielectric parameters in Fig. 3.7

and Fig. 3.8 are attributed to the low fringing fields present at the probe aperture at the low frequencies. This results in an inaccurate measurement of the dielectric parameters using the developed model, which is based on a lumped element approach. Using the bilinear calibration model [7], the mean dielectric constant obtained for Rexolite yielded a 4% deviation from the manufacturer's value of $\epsilon' = 2.55$, as shown in Fig. 3.8. This is in contrast to the 1% deviation obtained from the developed model, which is significantly smaller by comparison. In addition, Garcia-Ruiz *et al.* [7] reported dielectric measurement up to a maximum frequency of 18 GHz, as observed from the simulation results in Fig. 3.7 and Fig. 3.8. Nevertheless, the loss tangent values for the Garcia-Ruiz *et al.* model start to deviate beyond 18 GHz from Fig. 3.8, as significant radiation loss occur with the increase in the operating frequency and the probe simulates an antenna radiating out electromagnetic energy. In contrast, the developed model incorporating a frequency dependent term yielded a relatively stable response for the dielectric parameters extracted over the frequency range of 0.1-20 GHz. In particular, the results were obtained with the aid of foam as the optimum dielectric backing.

In conclusion, a comparison was made using a full wave simulator between the developed model and the bilinear calibration model, for which the former has been shown to be more effective for dielectric measurements compared to the latter, in terms of improved accuracy and a wider frequency range.

Figure 3.7: Simulation results showing the extracted dielectric constant for Rexolite ($\epsilon' = 2.55$)

Figure 3.8: Simulation results showing the extracted loss tangent for Rexolite ($\tan \delta = 0.001$)

3.5. Limitations of the Developed Model

This section provides a discussion of the limitations of the developed model, along with recommendations for an effective deployment of the model.

3.5.1. Range of Dielectric Parameters and Substrate Thickness

Apart from determining the measurement uncertainties attributed by different sample transverse shapes, this section seeks to investigate the range of sample dielectric constants and sample thicknesses for which the developed model is valid in extracting for the dielectric parameters of the MUT.

In this study, a sample of relatively high dielectric constant ($\epsilon'=6$) was used, since the value represents the upper limit of the typical dielectric range of packaging substrates. In addition, an expression relating the aperture admittance of an open-ended coaxial line to the complex permittivity of the MUT was adopted [52] to consider for finite samples supported by an infinitely thick foam backing, assuming the TEM mode. For the generic case of a lossy sample, this expression [52] can be represented by:

$$y = \frac{\epsilon}{\sqrt{\epsilon_{rc} \ln\left(\frac{b}{a}\right)}} \int_0^{\infty} \frac{[J_0(k_0 \zeta b) - J_0(k_0 \zeta a)]^2}{\zeta} F(\zeta) d\zeta \quad (3.14)$$

where

$$F(\zeta) = \frac{1}{\sqrt{\varepsilon - \zeta^2}} \left(\frac{1 + \rho_1}{1 - \rho_1} \right) \quad (3.15)$$

$$\rho_1 = \frac{\kappa_1 - 1}{\kappa_1 + 1} e^{-j2k_0 d_1 \sqrt{\varepsilon - \zeta^2}}$$

and

$$\kappa_1 = \frac{\sqrt{\varepsilon - \zeta^2}}{\varepsilon \sqrt{1 - \zeta^2}}$$

With reference to Eqn. (3.14), y represents the normalized admittance and R is the complex reflection coefficient. ε_{rc} represents the permittivity of the dielectric filling the coaxial line, k_0 is the free-space wave number and $\varepsilon = \varepsilon' - j\varepsilon''$ is the complex dielectric constant of the MUT with thickness d_1 . ρ_1 and κ_1 are recurrence functions. From Eqn. (3.15), there exist singularities in the integrand when $1 - \rho_1 = 0$ for $F(\zeta)$, which occurs when $|\rho_1| = 1$ and $\angle \rho_1 = 0^\circ$. These singularities indicate the excitation of surface waves and radiative waves [62] for the developed bilayer model employing foam backing. To investigate the relationship between the presence of singularities on the effective range of dielectric constants and sample thickness that can be extracted, a program was written (Appendix IV) to obtain the values of ρ_1 satisfying the limiting condition in Eqn. (3.15), where $\angle \rho_1 \leq 0.0001^\circ$ was set as the criteria to approximate $\angle \rho_1 = 0^\circ$.

Consequently, nomograms covering a range of sample thickness and dielectric constants were constructed across the frequency range of operation (0.1-20 GHz). The nomograms enable the sample thickness corresponding to a particular dielectric constant to be determined at a certain frequency, and vice versa. Obtained based on the probe dimensions of a SMA connector (4-hole square flange), they are useful in providing the upper bounds

(a)

(b)

Figure 3.9: Nomograms for a specified range of (a) dielectric constants
(b) sample thickness

when employing the developed model for dielectric measurements without the influence of surface waves, since the model considers only the dominant TEM mode. The dielectric constants considered ranged from that of FR4 samples ($\epsilon' \sim 4-6$) to ceramic substrates ($\epsilon' \sim 10-200$) and the sample thickness considered is from 1-5 mm. The respective nomograms are shown in Fig. 3.9(a) and Fig. 3.9(b). Generally, the sample thickness is inversely proportional to the dielectric constant with the increase in frequency. To avoid the surface wave phenomena, the maximum sample thicknesses measurable for FR4 substrates at 20 GHz is between 2-3 mm, and the value was observed to reduce significantly to 0.9 mm for a ceramic substrate with a dielectric constant of 200. This inverse relationship between the value of the dielectric constant and sample thickness is also evident from Fig. 3.10(a) where the useful range of dielectric constants measurable for a sample over 0.1-20 GHz declined from 20-1431 to 2.5-61 as the sample thickness increased from 1 mm to 3 mm. In conclusion, reduced sample thickness are required to perform dielectric measurements for mediums of increasing dielectric constants in order to avoid the propagation of surface waves and the decreased electrical lengths of the lumped elements across higher frequencies, which in turn invalidate the developed model considering only the TEM mode.

3.5.2. Uncertainty Analysis

Apart from surface waves resulting from evanescent TM modes excited at higher frequencies, dimensional and scattering measurement ambiguities could also result in measurement uncertainties for the developed model. Consequently, an uncertainty determination is crucial in obtaining knowledge of the degree of accuracy that can be

achieved with the developed model. Despite the availability of dielectric materials with well-known properties to determine a , b and c in Eqn. (3.11), the respective parameters obtained from the manufacturer's data are often only provided at specific frequencies and could undergo dispersion across much higher frequencies. However, difficulties arise in ascertaining the particular frequency range over which such dispersion phenomena could occur. Thus, the advantage of obtaining a closed form expression relating the sample complex permittivity to the measured reflection coefficient in Eqn. (3.11) lies in that an analytical investigation of the measurement uncertainty can be achieved where the sample permittivity ε_1 can be expressed as:

$$\varepsilon_1 = f(\rho)$$

where

$$\rho = \rho_0 \pm \Delta\rho$$

ρ_0 is the true reflection coefficient while $\Delta\rho$ represents the deviation in the reflection coefficient. The final expression obtained is:

$$\varepsilon_1 = f(\rho_0) + \frac{\partial f}{\partial \rho} \Delta\rho$$

where

$$\Delta\varepsilon_1 = \frac{a-bc}{1+c|\rho|^2} \times |\Delta\rho| \quad (3.16)$$

a , b and c are determined by the use of 3 suitable calibration materials with well known dielectric properties. By considering the worst case errors for a MUT within extreme loss tangent regions (ranging from low loss to high loss) and substituting them into Eqn. (3.16), the maximum deviation obtained compared to the actual dielectric constant was around 4%. Similar to other transmission line methods such as the coplanar waveguide technique, the developed model using the open-ended coaxial probe is found to be unsuitable in

determining the loss tangents for very low loss samples ($\tan \delta < 0.001$), though it is able to serve as a gauge for low loss samples from the loss tangent values obtained. This is since the fields are primarily normal to metal conductors and consequently the loss of the dielectric cannot be effectively distinguished from the metal losses, since the metal losses is much greater compared to the dielectric loss. Rather, the developed model was found to operate better for medium loss to high loss ($\tan \delta > 0.1$) materials. Numerical modeling was done to investigate the effects of a probe displacement on the dielectric parameters obtained. The results did not yield any significant deviation even for a probe offset distance that is 20% the sample transverse dimension, displaced from the centre of the MUT. This suggests that measurement errors arising from minor probe misalignment can be considered insignificant for a given sample transverse dimension that is much larger than the offset distance.

3.5.3. Conditions for Sample Preparation

While samples for dielectric measurements may have different transverse dimensions, cases where the MUT are copper cladded could also arise. For the case of samples that are cladded on both sides, their dielectric parameters cannot be accurately extracted for, since a short circuit would result as the samples are placed flush with the conducting flange of the open-ended probe, resulting in erroneous results. In this section, the case of a MUT backed by copper claddings was investigated using the developed model, to ascertain if the dielectric parameters of the MUT could still be effectively restored for such samples.

Numerical modeling was performed to restore for the dielectric parameters of RT5880 ($\epsilon' = 2.2$), for the case of an uncladded sample (with foam backing) and that of a sample backed by copper claddings. Eqn.(3.11) of the developed model was used, with air, Teflon and Fused Quartz as the calibration materials to extract for unknown parameters a , b and c . From the plot obtained in Fig. 3.11, significant deviations were observed for the restored dielectric constant of the copper cladded RT5880 across the entire frequency range (particularly beyond 12 GHz), as compared to the manufacturer's value of 2.2. A comparison between the phase plots for the two cases in Fig. 3.10 reveal a strong influence of the conductor backing in terms of a change of the slope in the phase variation across 13-15 GHz, which could be attributed to the strong reflections existing at the interface between the dielectric backing and the surface of the conducting plane. Consequently, the sharp phase change across the higher frequencies would correspond to electrical lengths of $n\lambda_m/4$ (n is an integer and λ_m represents the wavelength in the MUT), as the dielectric configuration changes from an open-circuit case (foam backing) to a short-circuit case. Compared to foam ($\epsilon' \sim 1.03$) with a dielectric constant close to that of air, a conductive backing beneath the finite MUT could be viewed as a short circuit case, when the probe is placed against a material layer shorted by a metallic plate, with $\sigma \rightarrow \infty$, where σ represents the conductivity of the metal.

Furthermore, radial guided waves are excited [62] at higher frequencies and the excitation of these guided electromagnetic waves would affect the evaluation of the dielectric properties of the MUT. This is since the coaxial probe setup comprising of the MUT sandwiched between the conducting flange and the conductive backing resembles that of a

Figure 3.10: S_{11} (phase) for cladded and uncladded RT5880 sample

Figure 3.11: Dielectric constant extracted for RT5880 sample

Fabry-Perot resonator at higher frequencies, with its metal-dielectric-metal configuration, and waves are bounced back and forth along the radial direction. The resonant phenomenon was generally characterized by a sharp decline of the S_{11} values and a sudden increase in the dielectric constant at around 14 GHz, as shown in Fig. 3.11. Due to the onset of resonating phenomena in the coaxial probe structure, the relation between the sample permittivity and the measured reflection coefficient becomes more complicated, and invalidates the quasi-static analysis of the developed model based on the lumped element technique, since the probe now operates at frequencies where the effects of radial guided waves have become significant and cannot be neglected. Thus a more comprehensive model such as that of a full wave analysis taking into account evanescent modes would have to be considered for metal-cladded samples, such as that introduced by Xu *et al* [63]. In conclusion, the developed model is more suited for restoring the complex permittivity of unclad substrates. To restore for the dielectric parameters of cladded samples, the copper claddings would have to be removed using Ferric Anhydrous solutions prior to performing dielectric measurements.

3.6. Influence of Air Gaps and the Effects on the Developed Model

The coaxial probe has been widely used for liquids and biological materials where excellent contact can be easily achieved. However this may not be the case for solid samples, since surface roughness and waviness are inherent in such samples. Nevertheless unless excellent contact between the probe and MUT is achieved in extracting for the dielectric parameters, the measurements obtained would be degraded by the presence of air gaps. This implies that for solid materials, the sample face has to be machined at least as flat as the probe face,

otherwise which air gaps present between the probe and sample would result in a significant source of error. In particular, the introduction of air gaps would decrease the probe's sensitivity, and the effect of small air gaps on the measured reflection response have been investigated both theoretically and experimentally [25-26, 31, 52-53]. Authors such as Jiang *et al* [25-26], and Arai *et al* [53] adopted a lumped capacitance model, with the latter employing the model to relate the sample surface roughness to the measured permittivity. On the other hand, Ganchev *et al* [52] considered the effect of air gaps by incorporating a preceding air gap layer between the MUT and the coaxial probe in a dielectric composite configuration. By accounting for the measurement inaccuracies arising from air gaps in that theoretical model, the permittivity of the MUT could then be restored with greater precision through the use of a rigorous solution [52]. The results of lift-off effects for the open-ended coaxial probe that was considered by Baker-Jarvis *et al* [31] suggest that air gaps in the order of fractions of a millimeter can significantly affect the reflection coefficient for samples with medium to high permittivity as well as probes with electrically small dimensions [31].

As air gaps contribute to a major source of error in the open-ended coaxial probe technique, an impetus arises to investigate the effects of air gaps between the coaxial probe and the MUT for the developed model. From Fig. 3.12(a) and Fig. 3.12(b), the sensitivity of the current SMA probes is affected by the presence of air gaps, where no distinction is observed in the S_{11} magnitude for differing dielectric constant values up to 7 GHz (Fig. 3.13(b)), as compared to 4 GHz (Fig. 3.13(a)) without the presence of any air gaps. This

Figure 3.12: $S_{11}(\text{Mag})$ vs. frequency plots for (a) perfect contact
(b) 0.25 mm air gap between coaxial probe and MUT.

can be attributed to the fact that the SMA probe dimensions does not suffice to measure for mediums with low dielectric constants at the lower frequencies [40]. To improve the probe sensitivity at the lower frequencies in the presence of air gaps, coaxial probes with larger dimensions could be used such that more fringing fields are present in the MUT during measurements, following which more accurate evaluations of the sample dielectric parameters can be obtained.

One of the problems encountered in using the coaxial probe technique lies in ensuring for the measurement repeatability using the same probe. This arises from the imperfection of the sample surface and the probe, resulting in air gaps between the probe and MUT. Consequently significant variations of the input impedance would arise in measuring for the same sample. Thus the developed bilayer model was used to investigate the influence of air gaps and their effects on the input impedance and dielectric constants (ϵ') for a MUT within a dielectric constant range of 2-6 (printed circuit board FR4 substrates). The study

made use of numerical modeling in *HFSS* and the simulation results obtained are for uniform air gaps of 0.25 to 1 mm (in increments of 0.25 mm), between the probe and the surface of the MUT. The thickness of the MUT is 1 mm and the loss tangent range considered is from the low to medium loss region ($0.001 < \tan\delta < 0.03$) corresponding to the FR4 substrates.

In computing the results, the percent error in the input impedance with the introduction of air gaps were determined at the upper limit of 20 GHz, while the mean percent error for the dielectric constant was computed for within a range of 4-16 GHz. The percent error in the lower frequency range (0.1 - 4 GHz) for the dielectric constant was neglected, since the probe sensitivity was low across the regions. On the other hand, the upper frequency limit considered was 16 GHz since beyond that value, the dielectric constant values would fluctuate due to the excitation of higher order modes. The percent error corresponding to the mean dielectric constant is obtained from Eqn. (3.17), where:

$$\begin{aligned}\varepsilon &= \frac{a\rho + b}{1 + c\rho} \\ \Rightarrow \frac{\partial \varepsilon}{\partial \rho} &= \frac{a - bc}{(1 + c\rho)^2} \\ \therefore \Delta \varepsilon &= \Delta \rho \times \frac{a - bc}{(1 + c\rho)^2}\end{aligned}\tag{3.17}$$

ε is the permittivity of the MUT and ρ is the measured reflection coefficient. Generally, the probe's insensitivity in measuring for low loss materials can be observed from Fig. 3.13(a), where slight variations are observed for the input impedance error obtained for low loss substrates relative to medium loss samples. Consequently, accurate dielectric

measurements of substrates that are low loss in nature would be constrained in using the open-ended coaxial probe technique.

With reference to Fig. 3.13(b), it was found that an air gap on the order of 0.25 mm can result in a significant error in the input impedance above 30% for the range of samples, with an error in the dielectric constant (ϵ') exceeding 15% for both types of loss regions. The results illustrate the potential of small air gaps in introducing drastic errors in the measurement results obtained, and the importance of alleviating for air gaps either by taking it into account in computations or through the experimental setup. In particular, the errors become significant when measuring for samples of high dielectric constant, since the small pockets of air gaps are present in the region of the highest electric field near the probe aperture. In the case of a sample with $\epsilon'=6$, a 0.25 mm air gap results in a significant deviation of 37% in the dielectric constant restored. For air gap sizes beyond 0.25 mm, the probe sensitivity decline significantly. In particular, it is noted that as the proportion of the maximal electric field from the probe aperture contained within the air gap layer increases with the increase in air gap from 0.5 mm to 0.75 mm, variations in the dielectric constant are less easily detected by the probe. This can be observed from the decreased slope in the Z_{in} error and percentage error plots in Fig. 3.13(b)-(c), where slight deviations were obtained beyond an air gap size of 0.5 mm. As the thickness of the MUT is increased, the corresponding errors introduced by existing air gaps are reduced, due to the smaller aspect ratio of the air gap size relative to the thickness of the MUT, and it can be inferred that the air gap effect is rendered less significant on such samples.

(a)

(b)

Figure 3.13: Percent error in (a) input impedance for low vs medium loss substrates (b) input impedance at 20 GHz due to air gaps (c) dielectric constant from 4-16 GHz.

In conclusion, the implication of air gaps on measurement reproducibility is that smaller measurement uncertainties are allowed in restoring for the dielectric parameters of the MUT with a specified accuracy of 5%, particularly as the air gap size increases. In turn, this poses a challenge in performing actual measurements, since uncertainties arising from fabrication and dimensional tolerances would also contribute to additional measurement inaccuracies. Consequently, using solid reference materials [7] in the developed model would aid in alleviating errors introduced by imperfect contact. This is since the MUT measured has a solid surface with inherent surface roughness that has to be taken into account, in this case through the use of solid reference materials [7]. In addition, sample

polishing could also be performed prior to taking measurements or an adjustable knob could be used in the measurement setup to exert pressure and improve the contact between the probe and the MUT, as detailed in Chapter 4. Although uniform air layers are not encountered in practical situations, considering uniform air layers in numerical modeling provides an insight into the measurement uncertainties arising from air gaps in an experimental setup. This section also discusses the importance of alleviating for such air gaps which could result in significant measurement errors if not duly considered or taken into account in the theoretical model or experimental setup.

3.7 Conclusion

In this section, the open-ended coaxial line technique is analyzed and an improved bilayer model that is applicable for measuring thin packaging samples was introduced using a lumped equivalent approach. The interaction between the probe, material under test (MUT) and the backing of the MUT is studied, and the sample backing was found to have a strong effect over the measured results, especially over high frequencies. Consequently, an optimum dielectric backing (foam) is required for an accurate extraction of the complex permittivity of thin packaging materials. A frequency dependent term was also included in the lumped circuit model to account for variations of the static capacitance with increasing frequency for a given coaxial line dimension. The validity of the model has been verified with simulation results obtained using a 3D full-wave simulator. Consequently, a comparison of the developed model with an existing model was made, where an improved accuracy and a wider frequency range of measurement was attained. The influence of

uncertainties in the sample transverse dimension on the extracted dielectric parameters was also investigated. In addition, the limitations present in the model are discussed, with an uncertainty analysis provided. As the open-ended coaxial probe technique is susceptible to air gaps, their influence on the extracted dielectric parameters using the developed model was also studied.

CHAPTER 4 EXPERIMENTAL REALIZATION OF THE DEVELOPED MODEL

4.1 Introduction

This chapter seeks to verify the simulation results obtained in Chapter 3 through experimental results obtained from a measurement setup. In the experimental verification, the results from dielectric measurements on substrates with well-characterized properties were obtained using the developed model. The effects of using different sample transverse dimensions and probe configurations on the restoration of the dielectric parameters were also investigated. Lastly, dielectric measurements on unknown FR4 laminates using the developed model are detailed in this chapter. The results were analyzed and the effects of imposing different testing conditions on their dielectric properties were also studied, with the corresponding implications discussed.

4.2 Measurement (Probe) Calibration

VNA calibration was then performed to calibrate the measurement setup by using a standard 3.5 mm calibration kit consisting of open, short and load standards. With the aid of the VNA internal software, the calibration coefficients could then be determined, thus allowing for measurements to be obtained at the reference plane by removing the systematic errors of the VNA and coaxial cable. In this case, the reference plane coincides with the end of the coaxial cable.

Nevertheless, upon the attachment of a subminiature A (SMA) probe to the coaxial cable, the measured reflection coefficient would differ from the actual value at the probe aperture due to impedance mismatches between the microwave components (VNA, coaxial cable and coaxial probe) as well as the phase shift arising from the electrical length of the respective components calibration, which references the reflection coefficient from the end of the cable. Though a VNA calibration resolves part of the systematic errors due to the VNA and the coaxial cable, there exists a need to calibrate the probe from the end of the coaxial cable (reference plane) to the probe aperture plane, upon attaching the probe to the coaxial cable. This would account for practical limitations such as the impedance mismatch between the probe and the coaxial line, the electrical length of the probe and the attenuation of the signal. Typically, the calibration procedure adopted involving S_{11} should be simple to facilitate real time measurements. The relationship between the measured reflection coefficient and the actual reflection coefficient referenced at the probe aperture can be described by Eqn.(3.9). To solve for the unknown coefficients in the equation, three known terminations with dissimilar reflection coefficients should be employed to achieve an accurate calibration. A few calibration models for the open-ended coaxial probe have been used in literature and most of them employ an Open, Short, Load (OSL) as terminations [27-28, 30], where a matched load was connected at the probe aperture plane rather than the end of the coaxial cable. However matched loads for non-standard cables are uncommon as compared to those available for standard coaxial cables. In particular, a matched termination would not be easily attached to the aperture plane and difficulties arise in designing a special matched load for the probe aperture plane that would be repeatable

while yielding an accurately known reflection coefficient. Furthermore, such a component would be expensive to fabricate.

Kraszewski *et al* [64] suggested the use of polar liquids of known dielectric properties as terminations for the probe. Consequently the elements of a two port error network can be determined provided that the equivalent circuit parameters of the probe are known. Such a technique using reference liquids has the advantage of allowing for calibration to be performed at the probe interface with readily available and well-defined calibration materials. While the relevant circuit parameters can be estimated from the probe geometry, such estimations may not be accurate since it is not representative of the actual experimental conditions. For instance, the conductors in real semi-rigid cables may not be perfectly coaxial. On the other hand, Blackham *et al* [65] used the properties of air, short and water as standards and mathematically removed the system errors from the measurements obtained by using the difference between the predicted and actual values. However, difficulties arise in obtaining a perfect repeatable short circuit over at higher frequencies. Li *et al* [61] used three fabricated shorting stubs of varying lengths to fit in the space between the inner and outer conductors of a coaxial air line for measurements over 0.3-4 GHz. A drawback of such an approach lies in that a good calibration could not be obtained below a certain cutoff frequency due to the large wavelengths, which resulted in the shorting stubs yielding phases that could not be well distinguished [61].

For the developed model, three reference materials with known permittivities were adopted for the calibration procedure, rather than employing the typical OSL due to the difficulties

Figure 4.1: Flowchart showing the calibration procedure used in performing dielectric measurements

that would arise. Essentially, the model incorporates a two port error network that takes into account the de-embedding procedure from the cable end to the coaxial probe aperture, which allows the probe to be calibrated using materials of known properties while performing dielectric measurements. The calibration materials employed in the developed model have stable and well known dielectric properties, such as Teflon®PTFE [7, 66], Rexolite [7, 66] and Fused Quartz. Two other commercially available microwave substrate materials (RT5880, RT6006) with well characterized dielectric properties were also selected as calibration materials. The samples were prepared in dices and measured using the setup shown in Fig. 4.4. The measurement calibration procedure performed is summarized in Fig. 4.1. During the calibration, the S_{11} of 3 well-characterized materials are obtained through VNA measurements using the test setup. By substituting the S parameters into Eqn. (3.11), three simultaneous equations can be obtained, following which the unknown calibration coefficients a , b and c in Eqn. (3.11) are determined. Altogether 401 points were obtained for the linear frequency sweep measurements over a bandwidth of 0.1-20 GHz using the VNA. After determining the 3 calibration coefficients, the S_{11} of the MUT is then obtained. The calculation procedure used to extract for the dielectric parameters of the MUT using Eqn. (3.11) is described in detail in Appendix I.

4.3. Experimental Verification of the Developed Model Through the Restoration of the Dielectric Parameters for Well-Known Materials

4.3.1. Using Different Transverse Dimensions

In this section, measurements were performed to investigate the minimum transverse dimension required in the experimental realization of the open-ended coaxial probe technique. This was done using well characterized materials with dielectric constants within a range of 2-10. From the analysis in Section 3.2.2, samples with transverse dimensions of 8x8 mm were found to perturb the measured reflection coefficients. Consequently, four transverse dimensions starting from 10x10 mm to 25x25 mm were investigated, in steps of 5mm, backed by Styrofoam of 10 mm thickness. The sample thickness is 1 mm and the dielectric measurements were conducted using SMA connectors (4-hole square flange mount) with extended Teflon insulators and defined with 50 Ω impedance. The flange dimension of the SMA connector is 12.7x12.7x1 mm. The connectors were modified by making a cut of the centre conductor prior to polishing the probe termination face, such that a smooth flat interface is obtained for which the conducting flange (acting as ground), signal and coaxial dielectric (Teflon) are coplanar. The MUT is then placed flushed against the probe termination face, backed by Styrofoam, which was found to offer the optimum sample backing as detailed in Section 3.2.1.

The probe was attached to a precision microwave coaxial cable that was connected to the test port of a HP 8720D vector network analyzer. To alleviate the measurement errors, cable flexure between performing calibration and taking measurements would have to be

minimized. Consequently, the measurement setup was designed with a coaxial holder to securely clamp the coaxial cable and probe in a vertical position, as shown in Fig. 4.4. To improve the contact at the probe sample interface, an adjustable knob was utilized. By adjusting the knob, sufficient pressure can be exerted on the MUT, such that it is effectively sandwiched between the probe interface and the Styrofoam backing underneath. In addition, the adjustable metal platform was raised to the maximum height such that the distance to be closed between the coaxial probe and MUT is minimized, thus reducing the cable flexure.

In this section, the parameter being analyzed for investigating the minimum transverse dimension is the normalized aperture admittance Y_{in} , which is a function of the measured reflection coefficients and calculated using Eqn. (3.7). From the equation, the values of Y_{in} serve as a general guideline in determining the extent to which the open-ended coaxial probe is applicable for varying transverse dimensions, as the restored sample complex permittivity is a function of Y_{in} . Prior to computing for the input impedance, the S_{11} were first de-embedded to remove the additional electrical length introduced by the SMA connector. The de-embedding procedure is described in Appendix III and the results obtained are shown in Fig. 4.2(a)(i)-(b)(i) and Fig. 4.2(a)(ii)-(b)(ii). From Fig. 4.2(a)(i)-(b)(i), it can be inferred that a minimum transverse dimension of 10x10 mm suffices for measuring substrates of 1 mm thickness with dielectric constants in the range of 2-10, up to a maximum frequency of 13 GHz. Furthermore, the results seem to suggest that the transverse dimension of the MUT need not exceed those of the flange, as long as the condition $A/\pi b^2 > 2$ [49] is fulfilled.

From Fig. 4.2(b)(i), it is observed that as substrates with increasing dielectric constants are measured, radial modes are excited at the higher frequencies (beyond the cut-off frequency), following which the size of the sample transverse dimensions become significant in determining the surface waves propagation as discussed in Section 3.2.2. Generally the undesired higher order modes strongly excited at the probe aperture would resonate in the sample when the sample electrical dimensions become sufficiently large. In addition, as the developed model is based on a lumped circuit configuration, the sample transverse dimension has to be below that of the operating wavelength in the MUT, such that the condition $l \ll \lambda_{\text{MUT}}$ is satisfied, where l is the electrical length corresponding to the dimensions of the MUT. As the wavelengths in the media are dependent upon the sample dimensions (and dielectric constant), this indicate that the electrical length of the MUT would become comparable to the shorter wavelength of the signal as the operating frequency increases, which is exacerbated if the MUT has a high dielectric constant.

Consequently the MUT would be considered as a distributed element when the dimensions exceed $\lambda_{\text{MUT}}/10$. This in turn renders the bilayer lumped circuit model based on a TEM mode invalid, as illustrated by the significant variations of Y_{in} observed over higher frequencies in Fig. 4.2(b)(i). The constraints in using a lumped circuit model can also be observed from the S_{11} phase plots of a MUT with a high dielectric constant ($\epsilon' = 10.2$), as shown in Fig. 4.2(b)(ii), where significant deviations occur with increased transverse dimensions at the higher frequencies.

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Figure 4.2: (i) Input admittance and (ii) S_{11} phase (before de-embedding) for substrates of (a) $\epsilon' = 2.2$ (b) $\epsilon' = 10.2$ with varying transverse dimensions.

Apart from studying the general effects of varying transverse dimensions on Y_{in} , the investigation was extended to studying the effects of the transverse dimensions on the developed model, in terms of restoring for the dielectric parameters of the MUT. The transverse dimensions used are as shown in Table 4.1 and the MUT (RT5880) had a dielectric constant of $\epsilon' = 2.2$ with a thickness of 1 mm. The corresponding calibration materials used for restoring the dielectric constant in Eqn.(3.11) are air, Rexolite ($\epsilon' = 2.55$) and RT6006 ($\epsilon' = 6.15$) with dimensions of 25x25x10 mm. The results are shown in Fig. 4.3,

Figure 4.3: Extracted dielectric constant using the developed model (manufacturer’s data $\epsilon'=2.2$)

Table 4.1: Table showing the accuracy of the restored dielectric constant corresponding to each transverse dimension

Transverse dimension (mm)	10x10	15x15	20x20	25x25
Accuracy (%)	3.5	1.8	1.6	0.8

where an accuracy of within 3.5% was obtained for the restored dielectric constants using the developed model. For MUTs with relatively low dielectric constant (eg. RT5880), the cutoff frequency for which the wavelength in the MUT approximates that of the transverse dimension is higher. Consequently, a larger transverse dimension could be used for dielectric measurements, which would allow higher order modes to attenuate sufficiently in

the radial direction. This is also illustrated in Fig. 4.3 and Table 4.1, where improved accuracy was obtained as the transverse dimension of the MUT was increased from 10x10 mm to 25x25 mm. However if a MUT with a higher dielectric constant is measured, smaller transverse dimensions should be used, so as to avoid the surface wave phenomena and the limitations of the lumped circuit model.

When the sample thicknesses were increased to 2 mm and 3 mm, the corresponding cutoff frequency for which the range of dielectric constants can be effectively restored declined to 10 GHz and 8 GHz respectively, due to the surface wave phenomenon described in Section 3.2.1, as well as the constraints of the lumped circuit model ($l \ll \lambda_{\text{MUT}}$). With an accuracy below 4% obtained in restoring for the dielectric constant of RT5880, any of the 4 transverse dimensions considered in this study should suffice for dielectric measurements, where a minimum sample transverse dimension of 10x10 mm is required. In cases where it is not possible or where the material is only available in small quantities, erroneous results could result from the restoration of the dielectric parameters, particularly in using rigorous techniques where the complex permittivity is expressed as a function of the input admittance and extracted through an inversion methodology.

4.3.2. Using Different Probes

In this section, two different probes modified from a BNC Receptacle ($2a \sim 2\text{mm}$, $2b \sim 7\text{mm}$) and a SMA connector ($2a = 1.27\text{mm}$, $2b = 4.12\text{mm}$) are used for dielectric measurements. Different probe dimensions affect the electromagnetic fields fringing into the material to

different extents, thus indirectly influencing the reflection coefficient obtained. Some authors such as Gershon *et al.* [67] optimized the open-ended probe by varying the probe dimension. The resulting probe dimensions corresponded to non-50 Ω impedance, which is impractical for a 50 Ω coaxial line, since the efficient transmission of the incident and reflected signal requires the impedance of the probe to match the 50 Ω impedance of the VNA and coaxial cable. For the experimental verification, commercial connectors (BNC Receptacle and SMA connector) were used instead as they could be readily purchased, thereby saving on the fabrication costs.

For the dielectric measurements in this section, the dielectric constants of two well characterized materials with stable dielectric properties over broadband - Fused quartz and RT5880, were restored through utilizing three different dielectric backings of Foam ($\epsilon' \sim 1.03$), Teflon ($\epsilon' = 2.1$) and Soda Lime glass ($\epsilon' = 6$), so as to verify the theoretical results in Section 3.2.1. As both materials are low loss in nature, the loss tangent values are not being reflected in this section. Following the analysis in Section 4.2.1, the MUT and dielectric backing were prepared in dimensions of 25x25x1mm and 25x25x10mm respectively. When RT5880 was tested, air, Fused quartz ($\epsilon' = 3.75$) and another Rogers substrate - RT6006 ($\epsilon' = 6.15$) were used as the calibration references. Similarly, air, RT5880 and RT6006 were used in testing for Fused Quartz. The calibration materials were chosen such that the calibration range of values would be evenly distributed. Consequently, Teflon was not used as a reference material since it would yield a similar reflection coefficient response compared to air, following which the possibility of obtaining poor calibration may arise.

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Figure 4.4: Measurement system for the developed model

From the results obtained (Fig. 4.5 to Fig. 4.8) using the developed model, it can be observed that the dielectric constants of the MUTs backed by the three different dielectric mediums were restored with a reasonable accuracy of around 10% both connector types, up to a frequency of 8 GHz. However, SMA probes modified from 4-hole square flange mount connectors are preferred for measurements as they are relatively small compared to the BNC Receptacles, which is an advantage when testing for samples of small dimensions, since the contact areas can be considered as approximately flat. Furthermore, the smaller probe dimensions allow for dielectric measurements to be conducted over a wider frequency range before radiation conductance becomes significant when $(b-a) < \lambda_m/15$ [68], where λ_m is the wavelength in the MUT.

From Fig. 4.5 and Fig. 4.7, results obtained using the BNC Receptacle yielded larger deviations of 7-10% for the dielectric constant (ϵ') restored, relative to those of the 4-hole flange. In particular, it is observed that the values of ϵ' start to fluctuate beyond 10 GHz for the BNC Receptacle, which can be attributed to the introduction of significant radiation conductance in relation to the bigger probe dimensions. In contrast, the SMA connector yielded relatively flat and stable frequency responses when performing dielectric measurements on Fused Quartz and RT5880, as shown in Fig. 4.6 and Fig. 4.8. Using the SMA connector, mean values of $\epsilon'=3.66$ and $\epsilon'=2.27$ were extracted for the Fused Quartz and RT5880 samples backed by foam as the optimum dielectric, which represent a 2-3% difference compared to the manufacturer's values of 3.75 and 2.20 respectively (Fig. 4.6 and Fig. 4.8).

Figure 4.5: Extracting for Fused Quartz using BNC Receptacle
(manufacturer's value: $\epsilon'=3.75$)

Figure 4.6: Extracting for Fused Quartz using 4-hole flange SMA
(manufacturer's value: $\epsilon'=3.75$)

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Figure 4.7: Extracting for RT5880 using BNC Receptacle (manufacturer's value=2.2)

Figure 4.8: Extracting for RT5880 using 4-hole flange SMA (manufacturer's value=2.2)

As observed from all the figures, the resulting standard deviations increased significantly when Soda lime glass of a higher dielectric constant was used to back the MUT, relative to using foam backing, which validate the theoretical results suggesting the importance of using an optimal dielectric backing to extract for the dielectric parameters of the MUT within an accuracy of 5%.

4.4. Comparison between Simulation and Experimental Results using Three Different Dielectric Backings

After having determined an appropriate open-ended coaxial probe for use in dielectric measurements, this section seeks to compare the simulation results obtained using *HFSS* with the experimental results from the measurement setup (shown in Fig. 4.4) in demonstrating the validity of the developed model. The comparison between the two sets of results is shown in Fig. 4.9 to Fig. 4.14. The data shown in the figures are obtained for the MUT backed by three different backings – Foam, Teflon and Soda lime glass, to restore for the dielectric constants of Fused Quartz and RT5880. The material parameters and testing conditions are as described in Section 4.2.1.

From Fig. 4.9 and Fig. 4.10, a good agreement is obtained between the simulation and experimental results when using foam backing to extract for the dielectric parameters, with deviations of within 1% between the two sets of results. The significant variations observed within 0.1-2 GHz is attributed to the minimal probe fringing fields at the aperture across the lower frequencies, leading to an inaccurate evaluation of the dielectric constant using the

developed model. For the case of Teflon and Soda lime glass backings (shown in Fig. 4.11 to Fig. 4.14), good correlation between the simulation and experimental results are also obtained for both Fused Quartz and RT5880. It is observed that significant fluctuations arise beyond 11 GHz in Fig. 4.13 and Fig. 4.14 for the restored dielectric constant of the MUTs, when the sample backing is changed from foam to Soda lime glass with a higher dielectric constant. This can in turn be attributed to the onset of the surface wave phenomena at higher frequencies, as discussed in Section 3.3.1.

4.4.1. Effect of Surface Condition on One of the Sample Backing

For the case of Teflon backing, investigations were carried out to ascertain if the backings with different surface conditions would affect the restored dielectric constant of the MUT. Due to dimensional constraints, two Teflon pieces of dimension 25x25x9 mm and another two of dimension 25x25x1 mm were prepared from one piece of bulk Teflon (dimension 70x70x9 mm).

One of the Teflon pieces (25x25x9 mm) was wax polished while the remaining pieces were manually polished to achieve a smoothened surface, as the bulk Teflon piece displayed roughened surfaces from possible wear and tear. Investigations were then conducted for different combinations of dielectric backing, as displayed in Table 4.2. The dimension of the MUT is 25x25x1 mm. For test cases 3 and 4 in Table 4.2, the 1 mm thick piece of Teflon was added to obtain the typical 10 mm thickness for the dielectric backing described in

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Figure 4.9: Comparison of simulation and measured results for RT5880 (Foam backing)

Figure 4.10: Comparison of simulation and measured results for Fused Quartz (Foam backing)

Figure 4.11: Comparison of simulation and measured results for RT5880 (Teflon backing)

Figure 4.12: Comparison of simulation and measured results for Fused Quartz (Teflon backing)

Figure 4.13: Comparison of simulation and measured results for RT5880 (Soda lime glass backing)

Figure 4.14: Comparison of simulation and measured results for Fused Quartz (Soda lime glass backing)

Table 4.2: Test cases showing the different combinations of Teflon backing used

Test Case	MUT	Teflon dielectric backing
1	Fused Quartz	Wax polished
2		Original piece (9 mm thick)
3		Wax polished stacked by 1 mm thick Teflon
4		Original piece (9 mm thick) stacked by 1 mm thick Teflon

Section 3.2.1, and for subsequent comparison with the simulation results which are obtained by considering a 10 mm backing. The results obtained are shown in Fig. 4.15.

From Fig. 4.15, discrepancies were observed between the restored dielectric constant obtained in test case 1 compared to test case 2. Test case 2 (9 mm thick Teflon backing) yielded dielectric constant values that are close to the *HFSS* results obtained for a 10 mm Teflon backing. Nevertheless, the results for test case 1 yielded a closer match to *HFSS* after the addition of a 1 mm Teflon piece placed on top of the wax polished Teflon. This suggests that the surface condition of the dielectric backing cannot be neglected, since the interaction of the wax layer with the signal at the MUT-dielectric backing interface resulted in a discrepancy of the dielectric constant evaluated using Eqn.(3.11), compared to using the Teflon backing in test case 2, as shown in Fig. 4.15. However, the restored dielectric

Figure 4.15: Comparison of test cases results against *HFSS* results

constant using wax polished Teflon backing (test case 1) displayed a more stable trend across 2-14 GHz compared to the Teflon backing in test case 2. This could be attributed to the flat surface attained from wax polishing, in contrast to the relatively uneven surface obtained from manual polishing the Teflon pieces, particularly for test cases 2 and 4, as observed in Fig. 4.15.

Apart from demonstrating a good correlation between the simulation and experimental results, this section seeks to highlight the effects of different surface conditions of the dielectric backing on an effective restoration of the permittivity of the MUT. This is in addition to the use of an optimum dielectric backing for dielectric measurements, as detailed in Section 3.2.1. On the other hand, the slight discrepancies observed between the

simulation and experimental results from Fig. 4.9 to Fig. 4.14 can be attributed to the presence of modeling and measurement errors, particularly since the simulated results from the developed model would not take into account the entire physical phenomenon arising from the measurement setup.

4.5. Material Characterization of Unknown Packaging Substrates

After having verified the validity of the developed model on samples with well known dielectric properties, with a good correlation obtained between theoretical and experimental results, dielectric measurements on packaging substrates (FR4 laminates) with unknown dielectric parameters were performed. In addition, a comparison was made between the preliminary and actual experimental results. The preliminary results were obtained prior to the fabrication of the measurement setup shown in Fig. 4.4. In attempting to obtain the measurement data in the absence of an adjustable clamp, hand pressure was vertically applied instead on the coaxial probe to sandwich the MUT (25x25x1mm) between the probe and Styrofoam backing (25x25x10mm) placed on a flat table, such that good contact can be achieved and the presence of air gaps alleviated. Precautions taken to minimize cable flexure during the preliminary measurements include fastening the mid-section of the coaxial cable to a fixed structure with tape. A VNA calibration was then performed prior to conducting the preliminary measurements. In contrast, the actual experimental results were obtained using the measurement setup shown in Fig. 4.4. In the setup, an adjustable knob was used to exert a more consistent pressure from the probe to the MUT, such that the contact at the probe-MUT interface is improved and existing air gaps are minimized.

The packaging substrates measured were FR4 organic materials, which are by far one of the most commonly used materials in printed wiring board industry as electrical interconnection substrates, PC board materials and for other microelectronics packaging applications. The advantages of FR4 laminates include ease of processing, machinability, low cost and availability. As they are made of woven mesh of glass fiber which is filled with organic filler, generally many variations of FR4-epoxy materials can exist depending on the types of fiber reinforcement and filler (epoxy) used. Consequently, factors affecting the dielectric properties of a FR4 laminate include the laminate constituent resin and fiber materials, fiber volume fraction and the reinforcement micro-structure. For instance, variations could arise in the resin volume fraction of a laminate, depending on the reinforcement weave style and processing conditions and the effect of resin volume fraction on laminate dielectric properties would be manifested in variations of the permittivity of the materials, where dielectric constants within a range of 3.8-5.5 are expected for such materials, according to the manufacturer's data. In this section, the preliminary and actual experimental results for two different sets of FR4 substrates were obtained, with the results shown in Fig. 4.16 to Fig. 4.21. The thickness of the FR4 substrates was around 0.8-1 mm. The corresponding values of dielectric parameters are displayed in Table 4.3 to Table 4.5.

From the results, the overall trend obtained from the preliminary results for the first set of FR4 samples are similar to that of the actual results, as observed in Fig.4.16 and Fig.4.17, where deviations ranging between 1-7 % are obtained for the dielectric constant (ϵ'). One of the possible experimental limitations accounting for the discrepancies could arise from the presence of air gaps between the coaxial probe and the MUT, which in turn lead to

inconsistent results due to the insufficient contact exerted. This is particularly so when considering solid samples with inherent surface roughness. Furthermore, the introduction of system noise in the network analyzer or the coaxial cable connecting them could have interfered with the measurement data, thus degrading the quality of the output data and the corresponding dielectric parameters extracted from the dataset. In particular, certain sections of the coaxial cable connected to the probe were flexed considerably during preliminary measurements, when hand pressure was applied on the coaxial probe against the MUT without the availability of a coaxial holder. Furthermore, using a wideband VNA for measurements could have resulted in a tradeoff in sensitivity.

For the second set of FR4 laminates, a good agreement between the preliminary and actual results was obtained, as shown in Table 4.5, Fig. 4.20 and Fig. 4.21, with deviations below 5%. The sharp decline observed in ϵ' for the EMC sample over 6-8 GHz in Fig. 4.20 could be attributed to inconsistent pressure being exerted on the probe in contact with the MUT during preliminary testing. Comparing Fig. 4.16 and Fig. 4.20 with Fig. 4.17 and Fig. 4.21, the introduction of system noise was observed to be more significant in the preliminary results, which could have been due to the flexing of the cable during measurements when bringing it into contact with the MUT. A further comparison between Fig. 4.17 and Fig. 4.21 suggest that the second batch of FR4 laminates are more suitable for high frequency applications as compared to the first set of FR4 substrates in Fig. 4.17. This can be inferred from the trend of the dielectric constants in Fig. 4.17, which start to decline beyond 4 GHz,

Figure 4.16: Dielectric constants of FR4 samples
(Preliminary results)

Figure 4.17: Dielectric constants of FR4 samples
(Actual results)

Figure 4.18: Loss tangents of FR4 samples (Preliminary results)

Figure 4.19: Loss tangents of FR4 samples (Actual results)

Figure 4.20: Dielectric constants of other FR4 samples
(Preliminary results)

Figure 4.21: Dielectric constants of other FR4 samples
(Actual results)

Table 4.3: Preliminary and actual results of dielectric constants – FR4 set 1

Sample	N1	N3	P1L1	P2L1	P3L1	P4L1
ϵ' (Prelim)	3.89	4.17	4.45	4.28	4.10	3.92
ϵ' (Actual)	4.17	4.50	4.51	4.38	4.29	4.21

Table 4.4: Preliminary and actual results of loss tangents – FR4 set 1

Sample	N1	N3	P1L1	P2L1	P3L1	P4L1
$\tan \delta$ (Prelim)	0.012	0.010	0.025	0.015	0.015	0.015
$\tan \delta$ (Actual)	0.014	0.012	0.024	0.016	0.015	0.014

Table 4.5: Preliminary and actual results of dielectric constants – FR4 set 2

Sample	EMC	M1L1	M1	M4	N2	N4
ϵ' (Prelim)	4.22	5.07	5.04	3.81	3.88	3.81
ϵ' (Actual)	4.40	4.95	4.94	3.99	3.93	3.79

in contrast to the relatively stable values displayed by the second set of FR4 laminates across the entire bandwidth from 0.1-20 GHz, as shown in Fig. 4.21.

Overall, the results obtained from the preliminary measurements are susceptible to systematic errors arising from the experimental setup such as air gaps and the flexing of the coaxial cable, which in turn introduces measurement errors leading to inconsistent results. With the availability of a proper experimental setup, the systematic errors were alleviated significantly, and the results obtained for the mean dielectric constants of the FR4 laminates are consistent with the dielectric range of 3.8-5.5 expected for FR4-epoxy based substrates,

based on the manufacturer's data. Furthermore, reasonable values of loss tangents were obtained, which are typical of the range of values for medium loss FR4 samples [69-70].

4.6. Effects of Different Testing Conditions on the FR4 Laminates

Among the challenges in the characterization of electronic packaging materials lie in the sensitivity of organic packaging materials to both temperature and moisture conditions, which add to the complexity of characterization. Consequently, the dielectric properties of electronic packaging materials are of interest as an indication of the robustness of the materials in surviving manufacturing, storage, handling and operating environments. In this section, the effects of the testing environment (temperature and moisture level) on the dielectric properties of a few classes of FR4 laminates are studied and a discussion of the results obtained is provided in this section.

The testing procedure was divided into 3 stages as shown in Fig. 4.22. Prior to stage 1, eight samples were measured for their dielectric properties and the results shown in Fig. 4.23 and Fig. 4.24. The first stage involved placing the samples in a baking oven for 96 hours (staggered into consecutive sessions of 48 hours and 24 hours at 125°C) to remove excess moisture present in the FR4 laminates. This was followed by conditioning in a humidity chamber for 48 hours (staggered into 2 sessions of 24 hours at 85°C) to simulate for an accelerated environment condition such that moisture uptake can take place.

Figure 4.22: Flowchart showing the 3 stages of testing conditions

Figure 4.23: Measured dielectric constants for FR4 laminates

Figure 4.24: Measured loss tangents for FR4 laminates

Moisture uptake (absorption) is defined as the increase in weight of a material after immersion in water for a specified duration at a specified temperature, and is dependent upon the chemical nature of the material as well as the presence of voids. Finally the samples were baked again for 96 hours as a post-conditioning stage. Dielectric measurements were performed for the FR4 laminates after each of the three processes, with the chief objective of investigating the post processing effects of various testing conditions on the dielectric properties of the FR4 laminates. Consequently, the implications of such post processing effects on the overall system performance could then be inferred from a macro-level perspective, which could in turn serve as guidelines for engineers involved in the design and development cycle.

4.6.1. Results and Discussion

In stage 1, the dielectric constants and loss tangents decreased after each round of baking, as shown in Fig. 4.25 to Fig. 4.29. Generally, the decline was 8-15% for the former and 10-30% for the latter. While the trend of the dielectric constant after 48 hours was similar to the pre-baking stage, significant variations were observed after an additional 24 hours (total 96 hours) of baking. In particular, the dielectric constant of P5 was observed to decline beyond 14 GHz, as shown in Fig. 4.28, though stable values were obtained prior to baking. This suggests the degradation of the dielectric properties upon prolonged exposure to high temperatures.

In the second stage, the samples were placed in a humidity chamber for 48 hours of conditioning. The electrical properties of the laminates changed after moisture uptake, where a partial restoration of the dielectric parameters was obtained within a range 2-8% for the dielectric constant and 8-20% for the loss tangent, compared to the post baking values. This is since water possesses a very high dielectric constant exceeding that of the laminates. Consequently, the dielectric constants of the laminates increased as the level of moisture absorption increased. While a longer period of conditioning could have resulted in a greater degree of restoration, the results suggest that a full restoration is not possible, since the various FR4 laminates have differing degrees of moisture uptake as a result of their permeability to moisture dependent on the respective material composition. Consequently, the limitations of the material properties results in an irreversible process. Furthermore, the laminates would undergo moisture saturation beyond a certain period of time. Comparing the 2 sessions of staggered conditioning, the second session reported increased loss tangent values within a range of 3-23% prior to those after the first session of conditioning.

For the final stage, the FR4 samples were baked again for another 96 hours. Due to the different material composition, the dielectric constants of the FR4 laminates studied were found to be only partially reversible after the 3 stages. For instance, the dielectric constant for 2 of the samples (EMC, M1) declined to similar values obtained during the first baking process of 96 hours, with deviations of within 1%, one of which is shown in Fig. 4.25. For the majority of the substrates (M4, N1-N4), however, the dielectric constants obtained after the last stage were comparable to the post conditioning values (within 1-2% deviation),

which suggests that further treatment of baking or conditioning would not significantly affect their dielectric constant, particularly for the N-class of FR4 substrates, shown in Fig. 4.26 and Figure 4.27. For these laminates with material properties which are less robust, permanent damage could have already existed at the post conditioning stage. Overall, the percentage weight of the laminates varied within 1% between consecutive stages of processing.

In conclusion, the dielectric properties of the laminates were affected by the testing conditions to different extents, due to the varying material composition. In particular, the dielectric parameters for the bulk of the FR4 samples underwent possible permanent damage either at the post baking or post conditioning stages, for which the final dielectric constants obtained were comparable to either the post baking or post conditioning values. However, P5 displayed significant fluctuations in the dielectric constant and loss tangent after the last stage of baking (Figure 4.28 and Fig. 4.29), which suggests a possible permanent damage to the dielectric properties that was exacerbated by the baking and conditioning processes. Such a behaviour has serious implications for packaging and design engineers, as the irreversible dielectric properties of substrates similar to P5 would adversely affect the module performance should the material be employed for design in the post processing stages. On the other hand, the trend obtained for the loss tangents of the FR4 laminates after the three stages were inconclusive and more investigations are required to study the behaviour of the loss tangents for such substrates.

4.7. Conclusion

In this chapter, the developed model was verified experimentally by restoring for the dielectric parameters of materials with known dielectric properties. In addition, the influence of varying transverse dimensions and sample thickness on the dielectric properties extracted were studied. The effect of using different probe sizes and sample backings were also considered. Lastly, the characterization of some unknown FR4 laminates was performed. In particular, the effect of temperature and moisture on the extracted dielectric properties of the packaging laminates was investigated. The motivation for the study arose as engineers frequently have to rely on the data provided by vendor in their design process. However, such values of the dielectric parameters are at best approximate (since they are often the initial values obtained at a particular frequency) and are not representative of the actual values of the dielectric parameters after subsequent processing stages. As discussed in Section 4.5., the baking and moisture absorption introduced during the intermediate storage and handling process could affect the dielectric parameters of packaging materials significantly, and adversely affect the system module performance. Consequently, it is important to take into account the effect of moisture on the packaging materials and obtain more accurate values of their dielectric parameters such that the design cycle becomes more effective.

Figure 4.25: Dielectric constant for EMC

Figure 4.26: Dielectric constant for N1

Figure 4.27: Dielectric constant for N2

Figure 4.28: Dielectric constant for P5

Figure 4.29: Loss tangent for P5

CHAPTER 5 REPEATABILITY OF MEASUREMENTS

5.1. Motivation

A main concern in employing the open-ended coaxial probe for dielectric measurements lies in ensuring for measurement repeatability of the results from the experimental setup. In particular, the measuring system capability of the setup has to be determined such that the reliability of the measured values can be accounted for. For the open-ended coaxial probe, the factors which may affect repeatability include the use of polished probes (with different polishing extents) and the fabrication tolerances of the MUT. For instance, imperfections of the sample surface and the probe would result in air gaps between the probe and MUT. Consequently significant variations in the input admittance Y_{in} (aperture admittance) would arise which result in discrepancies in the restored dielectric parameters. This chapter investigates the effects of above mentioned factors on measurement repeatability, and discusses how the measurement uncertainty arising from the factors could be minimized such that measurement reproducibility can be accounted for.

5.2. Effect of Polished Probes on Measurement Repeatability

5.2.1. Measurement Setup

In this section, two categories of SMA probes (Probe Type A and B) were used for the measurements. The probes were fabricated with different materials composition used for the inner and outer conductors. Their specifications are shown in Table 5.1. The

experimental setup employed in the investigation is shown in Fig. 5.1 where a flexible low loss coaxial cable (1m) is mounted onto the coaxial holder. A 3.5mm calibration kit was used to calibrate the measurement system using open, short and load standards, which references the reflection coefficient from the end of the cable. Next, the SMA probe is attached to the cable for measurements. The probe is left in the open and the resulting reflection coefficient was measured. Air was used as the MUT since it has stable dielectric properties with a reference dielectric constant of 1 and zero loss tangent, thus simplifying subsequent analysis of the measurement data.

5.2.2. De-embedding Procedure

In restoring for Y_{in} , the results obtained from the full wave simulator (*HFSS*) in modeling for the two different probes were used to compare against those obtained from actual measurements. The results from *HFSS* would serve to be the benchmark of comparison between the two sets of results and the degree to which they be matched. Prior to that, the S_{11} values had to be de-embedded. The de-embedding procedure (Appendix III) involved an exponential phase shift used to account for the additional electrical length of the transmission line, which corresponds to the difference between the calibration and sample measurement planes where:

$$S_{11}' = S_{11}e^{2\gamma l} \quad (5.1)$$

where γ represents the complex propagation constant, α is the attenuation constant and β is the propagation constant. S_{11}' represents the de-embedded S parameter while S_{11} corresponds to the measured S parameter value. In addition, l is the electrical length

Figure 5.1: The measurement setup

Table 5.1: Specifications of the SMA probes used

Probe Type	Probe Dimensions (mm)	Inner conductor material	Outer conductor material	Max operating frequency(GHz)	Cost/pc (S\$)
A	2a = 1.27, 2b = 4.12	Copper	Brass	18	4.50
B		Gold	Gold	18	10

Note: Costs are based on a small quantity basis.

introduced by the probe, which is measured using a micrometer screw gauge. In the actual computation, the values obtained for the attenuation constant α were found to be negligible compared to the propagation constant β , which implies that α can be neglected in the computation of γ .

In the measurements, four similar probes from Probe Type A (Probe 1–4 in Fig. 5.2) and five similar probes from Probe Type B (Probe 1-5 in Fig. 5.3) were obtained by cutting the centre conductor and polishing the probe such that the signal, coaxial dielectric and flange are coplanar. Using the setup in Fig. 5.1, the measured S_{11} were obtained. Consequently, plots of Y_{in} against frequency are obtained, as shown in Fig. 5.2 and Fig. 5.3.

5.2.3. Results and Discussion

From Fig. 5.2, it can be observed that similar probes yielded values of Y_{in} with different degrees of deviation, compared to the theoretical results obtained from *HFSS*. In particular, Probe 2 yielded a closer correlation with the *HFSS* results relative to the other three probes, with a maximum deviation of 25% (for Y_{in}) within 5-16 GHz. Such discrepancies could be attributed to the mechanical imperfections present in the open end of the Teflon-filled SMA probe. Firstly, the polishing of the probe aperture could have stretched and pulled the Teflon coaxial dielectric, in turn causing the metal conductors at the probe end to be covered with Teflon and be partially isolated. As each probe was polished to different extents due to the use of manual polishing, significant variations in Y_{in} obtained from the experimental results could also have arose from the different surface roughness of the probe

aperture, which cannot be captured using numerical modeling. This is since the coaxial probes are represented as flawless in the full wave simulator, though such is not the case in reality. In addition, inferior manufacturing of the coaxial probes which are of a lower quality would be undetectable by visual inspection, and could also have contributed to the discrepancies obtained. Consequently, the probe yielding the closest match to *HFSS* results from Fig. 5.2 and Fig. 5.3 could be selected for dielectric measurements, among the other polished probes. In particular, it is recommended that the values of Y_{in} be extracted from the experimental results within the range of 5 – 16 GHz in using Probe Type A, particularly if rigorous techniques are used for dielectric measurements, as the permittivity of the MUT are restored from Y_{in} through an inverse methodology. This is since large deviations of 50% and 40% are encountered at frequencies of 4 GHz and 20 GHz respectively. Despite Probe Type A having a cutoff frequency of 18 GHz in the manufacturer's specifications, the lower quality of the relatively cheaper probes could have resulted in a tradeoff in the maximum operating frequency allowable, for which 16 GHz is the upper frequency limit, as shown in Fig. 5.2.

On the other hand, a maximum deviation of 30% was obtained for the optimal probe (Probe 1) in Probe Type B within 0.1-18 GHz, as shown in Fig. 5.3. Compared to Probe Type A, the effective operating frequency range for these probes is wider, with an upper frequency limit of 18 GHz, beyond which significant deviations are observed. Consequently, Probe Type B are relatively more suitable for use in dielectric measurements using the developed model, considering that comparable deviations Y_{in} are obtained for Probe Type B relative to that of Probe Type A.

Figure 5.2: Plot of $\text{Im}(Y_{in})$ vs frequency for Probe Type A

Figure 5.3: Plot of $\text{Im}(Y_{in})$ vs frequency for Probe Type B

Overall, the results illustrate the importance of selecting a suitable probe type to minimize the deviations in the input impedance that arise from the effects of polishing. Generally, the results are applicable in the restoration of dielectric parameters using any approach in the open-ended coaxial probe method, though results seem to suggest the implications of such polishing effects are minimal for the developed model, since the model has is found to be relatively robust against the frequency constraints arising from the polished probes, with relatively stable results restored across 0.1-20 GHz, as shown in Section 4.2. Though such effects cannot be captured through a full wave simulator, different extents of polishing could affect the measurement data significantly, in turn creating difficulties in ensuring for measurement repeatability.

5.3. Effect of Fabrication Tolerances on Measurement Repeatability

After demonstrating the effects of probe polishing on the input admittance and the implications for measurement repeatability, another objective lies in investigating the effects of fabrication or dimensional tolerances on the measurement repeatability for the same probe.

Two similar polished probes of Probe Type A were used to restore the dielectric constant of Rexolite 1422 ($\epsilon' = 2.55$) using Eqn. (3.11) of the developed model. The same set of calibration materials (air, RT5880 and RT6006) were used for both probes. On the other hand, to investigate the effects of fabrication tolerances on the measurement repeatability,

(a)

(b)

Figure 5.4: Dielectric constant restored for Rexolite using (a) Probe 1
(b) Probe 2 (manufacturer's value: $\epsilon'=2.55$)

two sets of Rexolite with similar dimensions of 20x25 mm were used as the MUT. The results are shown in Fig. 5.4(a) and Fig. 5.4(b). Compared to the manufacturer's data, the deviation obtained from measuring Sample 1 and 2 was around 1% using Probe 1 and 2-3% using Probe 2 respectively. This suggests that the uncertainties introduced from the dimensional tolerances of the samples can be neglected, as the discrepancies obtained between the 2 Rexolite samples are negligible and within 5% accuracy.

On the other hand, while the effects of polished probes result in significant deviations of Y_{in} with deviations beyond 20% for Probe Type A (discussed in Section 5.2.3), the impact on the final restored dielectric parameters using the developed model is minimal, as an accuracy of within 3% was obtained for the Rexolite samples in this section. Consequently, this demonstrates the robustness of the developed model in restoring for the dielectric parameters over broadband (0.1-20 GHz). Nevertheless, the accuracy of the measured results could be further improved through the use of an optimum probe, as discussed in Section 5.2.3, since Probe 1 and Probe 2 yielded slightly different deviations for the restored values.

5.4. Effect of the Nature of the Calibration Materials on Measurement Repeatability

After having established the influence of fabrication tolerances on the accuracy of results, another objective of measurement repeatability is to investigate the effects of employing calibration materials with different nature (Fused Quartz against RT5880) on the restoration of the dielectric parameters. In this section, two sets of calibration materials were used to

restore for the dielectric constant of RT5880. Calibration set 1 comprises of dielectric materials air, Fused Quartz and RT6006 as shown in Table 5.1. On the other hand, calibration set 2 consists of dielectric materials air, Rexolite and RT6006. The experiments were carried out as described in Section 4.2.1, and the results shown in Fig. 5.5.

While measurement reproducibility can be obtained in using Fused Quartz as one of the calibration materials, Fused Quartz may not be an ideal calibration material for performing dielectric measurements when backed by foam, since the brittle nature of the material renders it difficult to exert pressure on the Fused Quartz against the foam backing. Consequently pockets of air gaps could arise between the probe and the MUT, thereby yielding inaccurate values of dielectric constants. The discrepancies obtained in the restored dielectric constants using the two calibration sets are shown in Fig. 5.5, where the deviation obtained compared to the manufacturer's data was 6% using calibration set 1, in contrast to 2% using calibration sets 2.

In conclusion, the effects of probe polishing and sample fabrication tolerances on the measurement repeatability were investigated. While the effects of polishing could lead to significant discrepancies in the aperture admittance, such effects have minimal impact on the developed model, for which good accuracy have been obtained for the restored dielectric constants. In addition, the effects of fabrication tolerances on the measured results were found to be negligible. On the other hand, the significance of selecting a suitable calibration set of materials in dielectric measurements to minimize the introduction of air gaps between the probe and the MUT has also been demonstrated in this chapter. In

particular, the presence of small air gaps could affect the measurement repeatability significantly, as discussed in Section 3.6.

Table 5.2: Specifications of the calibration materials

	Materials used	Transverse Dimension (mm x mm)	Thickness (mm)
Calibration Set 1	Air	NA	NA
	Fused Quartz ($\epsilon' = 3.75$)	25 x 25	1
	RT6006 ($\epsilon' = 6.15$)		1.27
Calibration Set 2	Air	NA	NA
	Rexolite 1422 ($\epsilon' = 2.55$)	30 x 25	1
	RT6006	25 x 25	1.27

Figure 5.5: Comparison of restoration for RT5880 using 2 different calibration sets (manufacturer’s data: $\epsilon'=2.2$)

CHAPTER 6 CONCLUSION AND FUTURE WORK RECOMMENDATIONS

6.1. Conclusion

In this thesis, a bilayer model was introduced based on the open-ended coaxial probe technique. The improved model was utilized to extract for the dielectric parameters of thin packaging substrates over a broadband frequency of 0.1-20 GHz. An accuracy of 2-3% was obtained for samples with well characterized dielectric properties, and a good correlation was achieved between the simulation and measured results over broadband. One of the advantages of the developed model is the inclusion of a two port error network to account for the additional electrical length introduced by the open-ended coaxial probe (using a modified SMA connector). By adopting a simple calibration procedure using three calibration materials with well characterized dielectric properties, the improved model allows for a convenient extraction of the dielectric parameters for the MUT, without having to specially fabricate a matched load attached to the probe aperture for measurement calibration. Furthermore the fabrication of custom made probes was circumvented through the use of inexpensive coaxial connectors, which were modified from commercially purchased SMA connectors and reduce the overall fabrication costs for the measurement setup. In addition, the presence of air gaps between the probe and the MUT was minimized, through the use of an adjustable knob in the setup to effectively sandwich the MUT between the probe and the foam backing during dielectric measurements, thus enhancing the contact between the probe and the MUT.

Consequently, measurement repeatability in the developed model was accounted for, since the various measurement uncertainties arising from imperfections between the probe and the samples were investigated and found to have minimal effect on the developed model.

In addition, a comparison between the developed model and another existing model was done using a full wave simulator. Generally, the developed model offers more versatility compared to the model by Garcia-Ruiz *et al.* [7], since no reference material with dielectric properties close to that of the MUT is required to attain good accuracy. With current components capable of operating up to significantly higher frequencies (VNA, coaxial cable, coaxial connectors), it is optimistic that the improved model would work over a wider frequency range, since the developed model has to be reconsidered for higher order modes only at frequencies where $(b-a)/\lambda_m > 2.5$ [49].

6.2. Recommendations for Future Research

From the results obtained, the following recommendations are made pertaining to the future work on dielectric measurements of thin packaging materials.

Firstly, being a transmission line technique, the developed model is able to extract for the dielectric parameters of medium to high loss materials. Nonetheless, low loss substrates are increasingly being used in the emerging packaging technologies. Consequently, a motivation lies in developing a suitable technique to accurately determine the loss

tangents of such electronic packaging materials. Resonator techniques are one of the most accurate methods in extracting for the dielectric parameters of low loss substrates, with precise accuracy. However, the main drawback of such techniques lies in the computationally intensive solution time, since rigorous methods are involved. To minimize the solution time and achieve improved accuracy for the restored dielectric parameters, a combination of the coplanar waveguide model and resonator technique could be considered. By leveraging on the broadband capability offered by the coplanar waveguide and the precise solution yielded by the resonator technique, broadband measurements (>20 GHz) for low loss samples which are less computationally intensive could be realizable by maximizing the merits of the two techniques.

Furthermore, the developed model only considers samples that are homogeneous in nature. Consequently, a more comprehensive measurement technique that allows the dielectric components of anisotropic or ferrite materials in the orthogonal in-plane and out-of-plane directions to be separated could be considered. Such a method could be established upon cavity-based or resonator techniques. In addition, the resonator technique would also be useful in the measurement of very thin substrates (<1 mm), which is currently not possible using the developed model where the minimum sample thickness measurable is around 0.8 mm.

AUTHOR'S PUBLICATIONS

Y.Y. Lim, A. Alphones, M.D. Rotaru and A.P. Popov, "Improved Calibration Model for Dielectric Constant Measurements on Thin Packaging Materials at Microwave Frequencies", *Symposium of Microelectronics*, Institute of Microelectronics, 4 June 2004.

Y.Y Lim, M.D Rotaru, A.Alphones, A.P Popov and R.Rajoo, "Improved and Fast Extraction of Dielectric Parameters of Thin Organic Packaging Materials Using Open-ended Coaxial Line Technique", *6th Electronics Packaging Technology Conference*, Singapore, 8-10 December 2004.

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APPENDIX I

This appendix summarizes the procedure involved in performing the dielectric measurements using a network analyzer (HP8720D) and how the data can be imported into Agilent ADS for further computation of the dielectric parameters of the MUT.

Prior to performing measurements, the coaxial cable and connector probes should be visually inspected. If they are dirty, it is recommended that they be cleaned using swaps with isopropyl alcohol (IPA).

Defining the Format and Storage of Data

1. Insert a floppy disk into the HP 8720D network analyzer
2. Press the SAVE/RECALL key; select INTERNAL DISK; then press
SELECT DISK
INTERNAL DISK
RETURN
DEFINE DISK SAVE (Select DATA ARRAY ON, DATA ONLY ON)
SAVE USING ASCII
RETURN

Setting the Frequency Range and Frequency Sweep

1. Press the START button and key in the starting frequency.
2. Press the STOP button and enter the end frequency.
3. Press MENU key, select NUMBER OF POINTS (enter 401).

Performing VNA Calibration

1. Press the CAL key; then press

CALIBRATION KIT

SELECT CAL KIT (select 3.5mmD, HP850520)

RETURN

CALIBRATE MENU (Select S11 1-port or S22 1-port)

2. Attach the Open standard to the coaxial cable and press OPEN.

(Care should be taken in ensuring a smooth fitting of the standard to the coaxial cable by turning the metal connector piece at the end of the coaxial cable to fit into the calibration standards during fitting. Otherwise, the mechanism of the standards may be degraded).

3. Attach the Short standard to the coaxial cable and press SHORT

4. Attach the Load standard to the coaxial cable and press LOAD; then press

BROADBAND

RETURN

DONE CAL

Establishing for a Good Contact between the SMA probe and the MUT

1. Press the FORMAT key and select PHASE.
2. Attach the SMA connector to the coaxial cable (Open). Note the phase on the display screen.
3. Press the SMA connector onto a metal foil (eg. Aluminium) backed by a soft material such as foam.

4. A good contact between the probe and the foil is established when the resulting phase observed changes by 180° .

Performing the Dielectric Measurements

1. Place a thin piece of foam (~ 0.5 cm) under the 1 cm Styrofoam which is used to back the MUT.
2. Upon contact of the probe with the MUT, visually inspect that the thin piece of foam is compressed. This would serve as an indication that sufficient pressure has been exerted on the sample.
3. From Section 4.2, three calibration standards are required to restore for the sample dielectric constant and loss tangent. With air (Open) used as one of the standards, the other standards available are RT5880 ($\epsilon' = 2.2, \tan \delta = 0.0009$), RT6006 ($\epsilon' = 6.15, \tan \delta = 0.0019$), Fused Quartz ($\epsilon' = 3.75, \tan \delta \sim 0.001$) and Rexolite ($\epsilon' = 2.55, \tan \delta \sim 0.001$).

Recording the Measured S_{11}

1. Press the FORMAT key and select LOG MAG.
2. Using the adjustable knob on the testing jig, press the probe against the MUT and fix it in position.
3. Press the SAVE/RECALL key and select SAVE DATA to save the data onto disk.

Transferring and Reading the Network Analyzer Data to Agilent ADS

1. Click on the ADS2003C icon.
2. In the main window under 'File Browser', create or open a project in ADS as shown in Fig. 1.

Fig. 1

3. Next, click on the 'New Data Display' icon from the top menu bar shown in Fig. 1.
4. In the data display window, select the 'Instrument Server' icon to set up the reading of the citifiles, as shown in Fig. 2.

Fig. 2

- 5. A window will pop up as shown in Fig. 3. Check 'READ' radio button. Select 'File' in 'Read From'. Then, browse to :/A drive where the citifiles are saved. Give Dataset Name or an appropriate name. Next, click on 'Read File'. This will read the data from the specified citifile into the dataset file.
- 6. Check on the 'Display Status Log' box underneath to view the data transfer process, as shown in Fig. 3.

Fig. 3

7. If the file is successfully read, the *Display Status Log* should give the message “File read was successful”, as shown in Fig. 4.

Fig. 4

Computing for the Dielectric Parameters of the MUT

1. After having transferred all the *S* parameter data from VNA to ADS, select the yellow ‘EQN’ icon from the menu sidebar (Fig. 5) to insert an equation. Variables can be entered into the pop up window as shown in Fig. 6 to compute for the dielectric parameters. From the formulations in Section 4.2, 3 simultaneous equations are obtained, where:

$$C\rho_{ref1}\epsilon_{ref1} + \epsilon_{ref1} = A\rho_{ref1} + B \text{-----} (1)$$

$$C\rho_{ref2}\epsilon_{ref2} + \epsilon_{ref2} = A\rho_{ref2} + B \text{-----} (2)$$

$$C\rho_{ref3}\epsilon_{ref3} + \epsilon_{ref3} = A\rho_{ref3} + B \text{-----} (3)$$

Solving for (1)-(3), we have:

Appendices

$$A = \frac{C(\rho_{ref1}\varepsilon_{ref1} - \rho_{ref2}\varepsilon_{ref2}) + (\varepsilon_{ref1} - \varepsilon_{ref2})}{(\rho_{ref1} - \rho_{ref2})}$$

$$B = C\rho_{ref1}\varepsilon_{ref1} + \varepsilon_{ref1} - A\rho_{ref1}$$

$$C = \frac{(\varepsilon_{ref1} - \varepsilon_{ref2})(\rho_{ref1} - \rho_{ref3}) - (\varepsilon_{ref1} - \varepsilon_{ref3})(\rho_{ref1} - \rho_{ref2})}{(\rho_{ref1} - \rho_{ref2})(\rho_{ref1}\varepsilon_{ref1} - \rho_{ref3}\varepsilon_{ref3}) - (\rho_{ref1} - \rho_{ref3})(\rho_{ref1}\varepsilon_{ref1} - \rho_{ref2}\varepsilon_{ref2})}$$

where

$\rho_{ref1,2,3}$ are the S_{11} values measured for the 3 calibration standards.

$\varepsilon_{ref1,2,3}$ are the permittivity of the 3 calibration standards from the manufacturer's data sheets. In this case, the loss tangents for all the calibration materials available can be considered negligible since they all lie in the low loss region.

When A , B and C has been computed for in ADS, the dielectric parameters of the sample can be calculated using:

$$\varepsilon_{test} = \frac{A\rho_{test} + B}{C\rho_{test} + 1} \quad (4)$$

where ρ_{test} is the measured S_{11} value for the MUT and ε_{test} is the permittivity of the MUT. The mean value of the restored dielectric constant can be computed for by using the 'mean' command in ADS, eg. *mean(real(epstest)*, where *epstest* represents ε_{test} in equation (4). Consequently, the average loss tangent of the sample can also be computed for using the same command. The command 'stddev' can be used to obtain the standard deviation of the dielectric parameters.

Appendices

Fig. 5

Fig. 6

Note: The yellow icons by the menu sidebar in the 'New Data Display' window also allow for several display formats of the variables in the form of Smith Charts, polar plots, rectangular plots and stacked rectangular plots. One can also choose to list the variables in a table format.

APPENDIX II

(Referenced from "Microwave Engineering" by D.M. Pozar, John Wiley & Sons, 2nd Edition, 1998)

(a) Calculating for the conductor loss α_c

$$\alpha_c = \frac{R}{2Z_0}$$

$$= \frac{R_s}{4\pi Z_0} \left(\frac{1}{R_i} + \sqrt{\frac{\sigma_i}{\sigma_o}} \frac{1}{R_o} \right)$$

where

R is the series resistance per unit length for the inner and outer conductors in a transmission line.

Z_0 is the characteristic impedance of the line

$R_s = \sqrt{\frac{\rho f \mu}{\pi}}$ is the surface resistivity and ρ is the bulk resistivity of metal.

R_i and R_o are the radius of the inner and outer conductors.

σ_i and σ_o are the inner and outer conductor material conductivity.

(b) Calculating for the dielectric loss α_d

$$\alpha_d = \frac{G}{2Z_0}$$

$$\gamma = \sqrt{k_c^2 - k^2} = \sqrt{k_c^2 - \omega^2 \mu_0 \mu_r \epsilon_0 \epsilon_r (1 - j \tan \delta)}$$

Most dielectric materials are low loss (in this case Teflon) where $\tan \delta \ll 1$

Appendices

Using $\sqrt{a^2 + x^2} \approx a + 0.5 \left(\frac{x^2}{a} \right); x \ll a$

$$\begin{aligned}\gamma &= \sqrt{k_c^2 - k^2 + jk^2 \tan \delta} \\ &\approx \sqrt{k_c^2 - k^2} + \frac{jk^2 \tan \delta}{2\sqrt{k_c^2 - k^2}} = \frac{k^2 \tan \delta}{2\beta} + j\beta \\ \therefore \alpha_d &= \frac{k \tan \delta}{2} = \pi f \sqrt{\mu_0 \varepsilon_o \varepsilon_r} \tan \delta = \frac{\pi f \tan \delta}{c_0}\end{aligned}$$

where:

c_0 is the speed of light.

G is the shunt conductance per unit length of the transmission line.

k_c is the cutoff wave number.

APPENDIX III

Derivation of the De-embedding Procedure for a 1 Port Network

$$V_1 = V_1^+ + V_1^- = \sqrt{Z_0}(a_1 + b_1)$$

$$I_1 = \frac{1}{\sqrt{Z_0}}(a_1 - b_1)$$

$$\begin{cases} a_1 = \frac{1}{2\sqrt{Z_0}}(V_1 + I_1 Z_0) \\ b_1 = \frac{1}{2\sqrt{Z_0}}(V_1 - I_1 Z_0) \end{cases}$$

$$\begin{bmatrix} V_1^{+'} \\ V_1^{-'} \end{bmatrix} = \begin{bmatrix} e^{-\gamma l} & 0 \\ 0 & e^{\gamma l} \end{bmatrix} \begin{bmatrix} V_1^+ \\ V_1^- \end{bmatrix} \Rightarrow \begin{aligned} a_1' &= e^{-\gamma l} a_1 \quad \text{----- (1)} \\ b_1' &= e^{\gamma l} b_1 \quad \text{----- (2)} \end{aligned}$$

Then substituting (2) into (1)

$$b_1 = e^{-\gamma l} b_1' = e^{-\gamma l} S_{11}' a_1' = e^{-2\gamma l} a_1' S_{11}'$$

$$\Rightarrow S_{11}' = S_{11} e^{2\gamma l}$$

The positive value l represents the length of transmission line removed from the structure.

APPENDIX IV

/* This program is written to find the maximum thickness measurable for a MUT at a particular frequency. By fixing the value of the sample thickness and varying the values of the dielectric constant, the program is also applicable for finding the maximum dielectric constant measurable for the MUT. The range of dielectric constant considered is from 4-200, while the range of thickness considered is from 1-5 mm. */

/* Consider a sample with dielectric constant =4, loss tangent=0 */

/* Range of x: $1 < x < \epsilon_{sr}$, where x represents the variable ζ in Eqn.(3.14) */

f=10e9; % Particular frequency considered

d=8e-3; % Max sample thickness measurable, vary through trial and error

c0=3e8; % Speed of light

k0=2*pi*f/c0; % Free space propagation const

epsr=4; % max dk considered

x=1.19:0.001:1.2; % Current range considered, to be varied until singularities are

% encountered, as described in Section 3.5.1.

format short e

num=[sqrt(epsr-x.^2)./(epsr*sqrt(1-x.^2))]-1; % Represents (κ_1-1) in Eqn.(3.15)

den=[sqrt(epsr-x.^2)./(epsr*sqrt(1-x.^2))]+1; % Represents (κ_1+1) in Eqn.(3.15)

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```
rho=(num./den) .* exp(-i*2*k0*d.*sqrt(epsr-x.^2)); % Represents the expression for  $\rho_1$  in
% Eqn.(3.15).
```

```
/* For singularity to exist, magrho=1 and phrho=0 */
```

```
magrho=abs(rho)
```

```
plot(x,magrho,'-b*')
```

```
grid on
```

```
hold on
```

```
phrho=angle(rho)*pi/180 %in degrees
```

```
plot(x,phrho,'-rx')
```

```
grid on
```

```
/* Sample thicknesses measurable at particular frequencies of 10 GHz, 15 GHz and 20
GHz without the occurrence of singularities */
```

```
%f=20GHz, max d=4 mm, x=1.196, phrho=1.3444e-003
```

```
%f=15GHz, max d=5.3 mm, x=1.197, phrho=1.8021e-003
```

```
%f=10GHz, max d=8 mm, x=1.196-1.197, phrho=1.3444e-003
```