Millimeter-Wave Printed Circuit Board Characterization Using Substrate Integrated Waveguide Resonators

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Abstract—This paper proposes a substrate integrated waveguide (SIW) cavity based method that is compliant with ground-signal-ground (GSG) probing technology for dielectric characterization of printed circuit board materials at mm-wavelengths. The paper presents the theory necessary to retrieve dielectric parameters from the resonant characteristics of SIW cavities with particular attention placed on the coupling scheme and means for obtaining the unloaded resonant frequency. Different sets of samples are designed and measured to address the influence of the manufacturing process on the method. Material parameters are extracted at V- and W-band from measured data with the effect of surface roughness of the circuit metallization taken into account.

Index Terms—mm-wave measurement, dielectric characterization, substrate integrated waveguide, resonator, GSG probing

I. INTRODUCTION

Millimeter wave technology paves the way to next generation wireless applications, e.g. high-bandwidth LAN at 60 GHz, vehicular collision avoidance at 77 GHz and 94 GHz imaging radar for security and defense are some prominent examples [1–3]. Volume manufacture requires either system-on-chip (SoC) or system-in-package (SiP) realization [2]. The SoC solution integrates passive and active components on a wafer, but yield suffers and real estate is increased, whereas SiP allows embedding active components into low-loss substrates. The latter approach presents an opportunity to mix active components manufactured with different processes, to reduce losses in passive components such as inductors and allow real estate intensive antennas to be patterned. There is a growing demand for using inexpensive printed circuit board substrates (PCBs) in SiP designs. The knowledge of the electromagnetic PCB substrate parameters is vital for accurate design, especially in the mm-wave range where dimensions based on electrical characteristic such as substrate permittivity are tolerance critical. Today, most substrate PCB manufacturers specify the dielectric parameters of their substrate materials at 10 GHz. Consequently in realized designs at mm-wavelengths, frequency shifts attributed to permittivity deviation commonly occur, e.g. [4]. This arises because there is no established accurate technique for on-board dielectric characterization of PCBs at millimeter wavelengths, despite the availability of a number of different material characterization strategies [5]. A comprehensive review of the techniques for dielectric characterization has been made in [6] and it has been concluded that waveguide cavity resonant methods remain the most accurate in terms of permittivity and loss tangent determination.

However, due to their form factor, waveguide characterization methods have classically been difficult to apply to microwave PCB materials. The full-sheet resonance method is generally used for PCB testing; yet this approach lacks accuracy for dielectric loss evaluation due to substantial radiation losses that are hard to quantify [7]. Other approaches for the PCB characterization are mostly restricted to microstrip resonators of various shapes [5], [8]. Since the parameters of microstrip resonators depend on the substrate thickness, multiple resonators designs are usually required for each nominal substrate thickness at a particular frequency. Besides, since the electric field is not homogeneous in the cross-section of a microstrip line, the method suits extraction of effective permittivity, e.g. [8], rather than absolute permittivity, and loss estimates also have uncertainty associated with them due to radiation and metal loss characterization complicated by etching tolerances, including irregularity of the strip cross-section (e.g. undercut) [5].

Ideally, a metallic waveguide cavity filled entirely with a dielectric would provide a fully screened test environment with near homogeneous field distribution and excellent loss tangent resolution potential. In order to synthesize such an
environment for a PCB material, some of the authors have earlier proposed the use of substrate integrated waveguide (SIW) resonators, [9] at microwave frequencies. However their utility at V- and W-bands has, to the authors’ knowledge, never been reported before.

SIW is a fully screened transmission line, which is compatible with standard PCB processing techniques [10], [11]. Electric field distributions in SIW are very close to those in metallic waveguides and there is a straightforward relation between the transmission lines. As a consequence, the field is uniform in the vertical direction for substrates thinner than a half of guided wavelength. The latter is a unique feature of SIW resonators in comparison with other planar resonators, and thus it allows direct measurement of the absolute value of relative permittivity as well as straightforward calculation of metal loss ensuring accurate characterization of loss tangent.

In this paper, the technique presented in [9] is extended to propose a simple and efficient method for mm-wave dielectric characterization with SIW resonators. Section II presents the underlying theory for extraction of the substrate dielectric properties from the resonant characteristics of an unloaded SIW cavity and discusses metal loss as well as a means to account for the effect of surface roughness. Section III concentrates on the design of practical resonators and discusses a GSG probe based coupling mechanism, equivalent schematic of the resonator and robust technique for extracting the unloaded resonant cavity frequency from the measured loaded one. Section IV presents experimental results obtained using the proposed method in V- and W-bands for a low-loss non-reinforced PTFE based PCB substrate suitable for mm-wave applications [12]. Main findings are discussed in the conclusions section of the paper.

II. THEORY

A. SIW Cavity Resonator

A substrate integrated waveguide rectangular cavity is shown in Fig. 1. It has been shown in [13], [14] that both propagation and attenuation constants of a SIW waveguide are equivalent to the quantities of a rectangular waveguide with effective width $a_{eff}$,

$$a_{eff} = a - 1.08 \frac{d^2}{s} + 0.1 \frac{d^2}{a}$$  \hspace{2cm} (1)

Additionally the resonant frequency of the TE$_{mn0}$ mode of a rectangular SIW resonator is obtained from:

$$f_{mn0} = \frac{c}{2\pi \sqrt{\varepsilon_r}} \sqrt{\left(\frac{m\pi}{a_{eff}}\right)^2 + \left(\frac{n\pi}{b_{eff}}\right)^2}$$  \hspace{2cm} (2)

where $c$ is the velocity of light in vacuum and $\varepsilon_r$ is the relative permittivity of the substrate.

As described in [13], there are three major loss mechanisms associated with SIW: dielectric loss, conductor loss and radiation loss. Therefore, the unloaded quality factor $Q_0$ of the cavity is:

$$\frac{1}{Q_0} = \frac{1}{Q_c} + \frac{1}{Q_d} + \frac{1}{Q_{rad}}$$  \hspace{2cm} (3)

Here $Q_c$ is the quality factor associated with the conductor loss, whereas $Q_d$ is the quality factor associated with the dielectric loss, whereas $Q_{rad}$ is the quality factor due to radiation loss, which is in case of SIW cavity is manifested by leakage through the via walls. The latter, as will be shown later in the paper by numerical simulations, can be designed as a negligible contributor subject to the sufficient via density ($s < 2d$)[15], [16]. $Q_c$ can be determined by the cavity geometry, permittivity of the dielectric and the surface resistance of the metal $R_s$, whereas $Q_d$ is reciprocal to the dielectric loss tangent $\tan \delta$ [17]. Conventional PCB metallization consists of copper or brass, whose electromagnetic properties are well-known.

The quality factor $Q_c$ for the TE$_{100}$ mode of a rectangular cavity is given as[17]:

$$Q_c = \frac{k_c^2 h \eta (2\pi^2 R_s)^{-1}}{(2n^2 b_{eff}^2 h + 2ha_{eff}^3 + n^2 b_{eff}^2 + a_{eff}^2)}$$  \hspace{2cm} (4)

where

$$k_c = \sqrt{\left(\frac{\pi}{a_{eff}}\right)^2 + \left(\frac{n\pi}{b_{eff}}\right)^2}$$  \hspace{2cm} (5)

$h$ is the substrate thickness, and

$$\eta = 120\pi \sqrt{\varepsilon_r} Ohm$$  \hspace{2cm} (6)

For different wall, top and bottom metallization with surface resistance $R_{sw}, R_{st}$ and $R_{sb}$ respectively, equation (4) has been revised in this paper as:

$$Q_{c,d} = \frac{k_c^2 h \eta \pi^{-2}}{2R_{sw}(2n^2 b_{eff}^2 h + 2ha_{eff}^3)} + \frac{k_c^2 h \eta \pi^{-2}}{(R_{st} + R_{sb})(n^2 b_{eff}^2 + a_{eff}^2)}$$  \hspace{2cm} (7)

This modification is necessary in order to accommodate the physical realization of the test structure, see below section III. We note that for a rectangular cavity with reduced height, conductor losses in the sidewalls (reciprocal of the first term of (7)) are smaller than those in the top and bottom walls (reciprocal of the second term of (7)).

B. Extraction of Dielectric Properties

Once the resonant frequency and the quality factor of the resonator have been measured, the substrate permittivity $\varepsilon_r$ can be recovered from (2). Subsequently, the loss tangent $\tan \delta$ can be obtained from (7) combined with (5) and (6).

$$\varepsilon_{r,rect} = \left(\frac{c}{2\pi f_{mn0}}\right)^2 \left(\left(\frac{m\pi}{a_{eff}}\right)^2 + \left(\frac{n\pi}{b_{eff}}\right)^2\right)^2$$  \hspace{2cm} (8)

$$\tan \delta = \frac{1}{Q_d} - \frac{1}{Q_{c,d}}$$

Importantly, it follows from (8) that the extracted permittivity does not depend on the thickness of the substrate and the metal loss is defined by closed-form accurate expression (7). This makes the presented approach preferable.
to other planar techniques like microstrip or stripline resonator measurement, where individual designs and subsequent extractions are necessary for each substrate thickness. The radiation loss is discarded in (8) as the proposed method assumes design with negligible leakage.

C. Surface roughness effect

It is important to mention that the calculation of $Q_{cp}$ is carried out using the conductivity of bulk metal. However the metal in the real structure is rough on the inside in order to ensure proper adhesion to the dielectric of the substrate. It has been previously shown that metal roughness leads to enhanced attenuation usually described by a loss enhancement factor, [18]:

$$K_{SR} = \frac{P_{a, rough}}{P_{a, smooth}}$$

which is the ratio of power absorbed in a rough metal surface $P_{a, rough}$ to a power absorbed in the smooth metal $P_{a, smooth}$. This ratio is always greater than unity and thus the tan $\delta$ extracted from (8) without roughness effect considered, in fact sets an upper limit on the dielectric loss tangent as it is calculated for the lowest possible metal loss. For the quality factor calculations following substitution [18] is made in (7):

$$R_{s, rough} = R_{s, smooth} K_{SR}$$

which is equivalent to $\sigma_{rough} = \sigma_{smooth} K_{SR}$

III. SAMPLE DESIGN

We now study PTFE-based Taclamplus substrate whose specified permittivity is 2.1 and loss tangent is 0.0008 at 50GHz [12]. The thickness of the substrate was 0.1mm. The top metalized layer is made of 18$\mu$m thick copper with electrical conductivity 5.8e+7 S/m. The bottom ground plane is made of 3mm thick 63-37 brass with conductivity 1.6e+7 S/m.

A. SIW Cavity Design

The design of a practical SIW cavity requires an appropriate feeding mechanism. Most mm-wave circuits either have a waveguide interface to a measuring instrument or are directly probed.

In this paper we propose the use of small U-shaped aperture for probing, as shown in Fig. 1. The aperture is fed by a standard GSG probe thus removing any uncertainty with respect to the reference plane [21], which arises when a length of transmission line is inserted between the generator and the coupling element, e.g. coaxial connector [21] or a length of microstrip line feeding ring resonator trough coupling gap [5]. The probe acts as a current source feeding the resonator through inductive coupling. The U-shaped slot acts as a current source and placed close to the magnetic field maximum to ensure effective coupling. The resonator is represented as a series RLC circuit at the equivalent circuit in Fig. 2. This is the second Foster’s form in [21]. As discussed in [21] for the second Foster’s form the relationship between loaded $f_l$ and unloaded $f_u$ resonant frequencies reads as

$$f_l = f_u \left(1 + \frac{b_e}{2Q_e}\right) = f_u \left(1 + k \frac{b_e}{2Q_0}\right)$$

where $b_e$ is the coupling susceptance, $Q_e$ the external quality factor, $Q_0$ the unloaded quality factor, $k = d/(2-d)$ the coupling coefficient and $d$ the diameter of the resonant circle on the Smith chart. The coupling is found from the detuned reflection coefficient $\Gamma_d = (1-jb_e)/(1+jb_e)$ representing the reflection coefficient at a frequency far away from resonance. Correct choice of the Foster’s form is crucial for proper determination of the unloaded frequency, as both
positive and negative detuning is possible in (9) [21]. For the dielectric characterization, the incorrect choice of the equivalent circuit can lead to an error in the extracted dielectric parameters.

In order to verify the choice of the Foster’s form of the SIW resonator proposed here, a parametric analysis of the coupling circuit was performed using CST Microwave Studio. A resonator with \( a=2.35\text{mm}, \, b=2.8\text{mm}, \, d=0.2\text{mm}, \, s=0.34\text{mm} \) has been simulated at the TE\(_{120}\) resonance. A coupling aperture with \( w_u = w_d =0.1\text{mm}, \, w_c = 0.07\text{mm}, \, L_p = w_t = 0.08\text{mm}, \) and variable aperture length \( D_p \) is placed at \( y_j = 0.46a \). The position \( y_j \) has been chosen close to the magnetic field maximum for the selected mode in order to get sufficient coupling.

The results of the simulations are presented in Fig. 3. As one can observe from the Smith chart, an inductive coupling is present. This type of coupling results in \( f_L < f_u \) for the series resonator (the second form), whereas for the parallel resonator the opposite holds true [21]. One can see that the coupling coefficient \( k \) grows (i.e. the resonant circle widens at the Smith chart) with \( D_p \). Furthermore, from the absolute value plot of \( S_{11} \), it follows that \( f_L \) decreases with \( D_p \) increased. Hence, according to (9) \( b_c < 0 \). All these factors indicate that the SIW circuit of Fig. 1 indeed corresponds to a series resonator with inductive coupling as in Fig. 2, c.f. Table I in [21].

![Fig. 3. Effect of the coupling slot length on the reflection coefficient of the resonator.](image)

The parameters \( f_L, Q_u, Q_0 \) and \( \Gamma_u \) can be retrieved from the measurements. In this paper Matlab Q0REFL and Q0TRAN programs from [23] have been used for this purpose. The unloaded frequency was extracted from measured parameters with the aid of (9).

### C. Samples description

The experimental samples were designed in order to establish the method sensitivity to various geometrical features of the cavities. Namely, the effect of via diameter and separation is studied as this can help to assess the effect of manufacturing tolerances and potential scalability of the technique to frequencies higher than W-band. Also, the effect of perturbation of the coupling aperture geometry is considered as its shape can be affected by etching tolerances.

Three different sets of 13 resonator geometries prepared for the experiment with all of them having two additional replicas, see Fig. 4. Each of 13 resonators in all sets has cavity dimensions \( a = 2.25\text{mm}, \, 2.35\text{mm} + n*0.2\text{mm}, \, n=0,1,...,11 \) and \( b = 2.8\text{mm} \), designed to resonate at the TE\(_{120}\) mode at a certain frequency within the 60-110GHz frequency range. The resonators with equal \( a \) are gathered in the rows of nine samples on the test board, see Fig. 4. The first and second sets were designed for one-port measurements and have different via dimensions and apertures. For all resonators the excitation point is placed at \( y_j = 0.46a \). The third set was designed for two-port weakly excited resonators measurements, in which case the loaded resonant frequency is expected to be very close to the unloaded one. Most of the geometry parameters for the resonators of the third set are retained from the first set except for two apertures shifted to yield weak coupling, see inset in Fig. 4.

The parameters common for all sets of the apertures are \( w_u = w_d = 0.1\text{mm}, \, w_c = 0.07\text{mm}, \, w_t = 0.08\text{mm}, \, D_p = 0.25\text{mm} \). The parameters that differ between the sets are gathered in Table I.

![Fig. 4. Photograph of the test samples. Columns 1,4,7 –set 1, columns 2,5,8 – set 3, columns 3,6,9 –set 2 ](image)

<table>
<thead>
<tr>
<th>Table I. Parameters of the measurement sets</th>
</tr>
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<tbody>
<tr>
<td>Set</td>
</tr>
<tr>
<td>-----</td>
</tr>
<tr>
<td>#1</td>
</tr>
<tr>
<td>#2</td>
</tr>
<tr>
<td>#3</td>
</tr>
</tbody>
</table>
apertures remain constant. At higher frequencies the slot becomes larger with respect to the cavity dimensions and stronger coupling occurs, similar to the effect demonstrated in Fig. 3. The first two sets are designed to have coupling coefficient within the 0.1-1 range, which is sufficiently low to minimize the uncertainties of reflection type resonator measurements [24]. Likewise, the coupling for the two-port measurements is deliberately kept low for straightforward recovery of the unloaded resonant frequency and quality factor directly from the S21 curve and to reduce measurement uncertainty [25].

Further design constraints stem from the available measurement instrument dimensions (probe pitch of 150 μm) as well as cost efficient PCB etching tolerances, (which generally impose etched slot or track to be wider than 0.07 mm and diameter of the via greater than that of the substrate height with 0.05 mm increment in the drill bit diameter).

IV. EXPERIMENTAL RESULTS

A. Measurement Setup

The structures were measured using a Cascade millimeter-wave probe station with 50 ohm ground-signal-ground (GSG) probes. The pitch of the probes is 150 μm. The probes before each measurement were calibrated using an automated LLRM-procedure at the probe station and the calibration error was below 0.1 dB for a frequency band of 2 GHz in the vicinity of the resonance of each sample. This error level is acceptable as the mentioned Matlab programs apply curve-fitting algorithms and are capable of retrieving necessary characteristics from noisy data [23], [24]. A microphotograph of the probed samples from the first set is shown in Fig. 5.

B. Resonant frequency and permittivity measurements

The measured resonant frequencies of the samples are gathered in Table II. The natural resonant frequency f120 is calculated using equation (2) using the manufacturers nominal value for permittivity measured at 50 GHz εr = 2.1 [12]. The unloaded resonant frequency fu is extracted from the measurement and averaged over sample replicas. The unloaded resonant frequency was extracted with the frequency detuning due to loading as defined in (9) taken into account; the details of the effect will be discussed later in the paper.

The relative difference of the two resonant frequencies Δf incorporates both the difference between the declared and measured permittivity as well as the measured unloaded frequency uncertainty due to manufacturing tolerances and the measurement error. Thus, the characteristic Δf imposes an upper limit on uncertainty of the permittivity and is employed as a figure of merit for comparative analysis between different sets of samples.

It follows from Table II that resonator sets 1 and 2 exhibit similar uncertainty in the resonant frequency estimation whereas the uncertainty of set 3 is approximately 3 times higher. These results confirm that the two-port measurement technique results in higher uncertainty than the one-port measurement. Analysis of the measured results further reveals that modest variations in via diameter and spacing as well as aperture shape do not noticeably affect the quality of the extraction of unloaded resonant frequency.

The unloaded resonant frequency for each measured sample has been used on the expression for the relative permittivity (8) and the resultant data in all the diagrams below is plotted against the nominal natural resonant frequency from (2). As expected due to manufacturing tolerances there is spread of the measured permittivities, which are shown with error bars in Fig. 6.

![Microphotograph of probed one-port resonator](image)

**Fig. 5. Microphotograph of probed one-port resonator.**

Since the measured permittivity values do not vary significantly with frequency within the measured range, the values have been averaged in order to obtain a single value for

| Table II. Measured resonant frequencies for each set of samples. |
|-----------------|-----------------|-----------------|-----------------|-----------------|-----------------|
| a, mm | f120, GHz | f_u, GHz (mean) | Δf (f_u/f120) % |
| #1 | #2 | #3 | #1 | #2 | #3 | #1 | #2 | #3 |
| 2.25 | 104.76 | 106.75 | 104.76 | 104.94 | 107.18 | 106.44 | 0.17 | 0.40 | 1.60 |
| 2.35 | 100.71 | 102.54 | 100.71 | 101.00 | 102.93 | 102.29 | 0.29 | 0.38 | 1.57 |
| 2.55 | 93.68 | 95.26 | 93.68 | 93.80 | 95.64 | 94.91 | 0.13 | 0.40 | 1.31 |
| 2.75 | 87.80 | 89.18 | 87.80 | 88.10 | 89.48 | 89.01 | 0.34 | 0.33 | 1.38 |
| 2.95 | 82.83 | 84.06 | 82.83 | 83.07 | 84.35 | 83.86 | 0.29 | 0.34 | 1.24 |
| 3.15 | 78.58 | 79.69 | 78.58 | 78.70 | 79.86 | 79.38 | 0.15 | 0.21 | 1.01 |
| 3.35 | 74.92 | 75.93 | 74.92 | 74.99 | 76.16 | 75.68 | 0.10 | 0.29 | 1.01 |
| 3.55 | 71.73 | 72.67 | 71.73 | 71.92 | 72.89 | 72.39 | 0.26 | 0.29 | 0.92 |
| 3.75 | 68.95 | 69.83 | 68.95 | 69.10 | 69.98 | 69.67 | 0.22 | 0.22 | 1.04 |
| 3.95 | 66.50 | 67.33 | 66.50 | 66.73 | 67.55 | 67.21 | 0.34 | 0.34 | 1.07 |
| 4.15 | 64.33 | 65.12 | 64.33 | 64.50 | 65.41 | 65.01 | 0.27 | 0.45 | 1.04 |
| 4.35 | 62.40 | 63.15 | 62.40 | 62.53 | 63.44 | 62.91 | 0.21 | 0.45 | 0.82 |
| 4.55 | 60.68 | 61.40 | 60.68 | 60.93 | 61.67 | 61.12 | 0.40 | 0.43 | 0.72 |
Taclamplus permittivity at mm-waves. Table III contains the data obtained. For single-port experimental tests both the average value and the error are very similar. For the two-port measurements, made on set 3, the error is about 2 to 3 times higher in comparison to single port.

TABLE III. COMPARISON OF AVERAGED MEASURED PERMITTIVITY TO THE ONE SPECIFIED BY MANUFACTURER.

<table>
<thead>
<tr>
<th>Set</th>
<th>Meas.</th>
<th>Spec</th>
<th>ε_r</th>
<th>Error, %</th>
</tr>
</thead>
<tbody>
<tr>
<td>#1</td>
<td>2.0824</td>
<td>2.1</td>
<td>0.84</td>
<td></td>
</tr>
<tr>
<td>#2</td>
<td>2.073</td>
<td>2.1</td>
<td>1.29</td>
<td></td>
</tr>
<tr>
<td>#3</td>
<td>2.0533</td>
<td>2.1</td>
<td>2.23</td>
<td></td>
</tr>
</tbody>
</table>

C. Quality factor

The measured unloaded quality factor and coupling coefficients are plotted in Fig. 7. One can see that for most of the cavity designs the spread between replica measurements is not large. However, some measurements, e.g. sets 1 and 2 around 95GHz and set 3 around 87GHz and 72GHz reveal a large spread, which is attributed to measurement error and manufacturing tolerances. The calibration error was generally kept below 0.1dB, yet for the specified cases it was as high as 0.5dB. Naturally, this has greater impact on the quality factor than on the resonant frequency.

Analysis of the single-port resonator measurements showed that b_e is within the range -0.5...-1 and the experimentally obtained (f_u - f_0)/f_u is plotted in Fig. 7c. It can be observed that detuning is less than 0.25%. Such detuning if not taken into account would result in 0.5% error in permittivity estimation.

D. Loss tangent and metal roughness effect

The extracted quality factor was substituted into (8) along with the calculated Q_D and the loss tangent was calculated. This calculation takes into account that two different metals copper and brass are used to form the top and bottom of the cavity, and the side wall posts metallization is copper. First, the calculation has been performed for smooth metallization. The extracted data were averaged over frequency and the mean values of all the test sets are close to 0.0011, see Table IV.

The results are larger than the specification by 34-40%. As discussed above, this figure is an upper limit on the loss tangent value and to further improve accuracy of the extraction the surface roughness effect has to be considered.

A comprehensive discussion of the loss enhancement factor calculation is carried out in [18]. It is proposed that for proper characterization metal surface profile should be measured and statistical distribution parameters established. Yet, this would require destruction of the samples which is unacceptable in many cases.

TABLE IV. COMPARISON OF MEASURED LOSS TANGENT TO THE ONES SPECIFIED BY MANUFACTURER.

<table>
<thead>
<tr>
<th>Set</th>
<th>Spec</th>
<th>Meas., smooth</th>
<th>tan δ, 10^{-3}</th>
<th>Error, %</th>
<th>Meas., rough</th>
<th>Error, %</th>
</tr>
</thead>
<tbody>
<tr>
<td>#1</td>
<td>0.8</td>
<td>1.12</td>
<td>40</td>
<td>0.734</td>
<td>8</td>
<td></td>
</tr>
<tr>
<td>#2</td>
<td>0.8</td>
<td>1.12</td>
<td>40</td>
<td>0.738</td>
<td>8</td>
<td></td>
</tr>
<tr>
<td>#3</td>
<td>0.8</td>
<td>1.07</td>
<td>34</td>
<td>0.684</td>
<td>14</td>
<td></td>
</tr>
</tbody>
</table>
Fig. 8. Quality factor for metal loss in smooth and rough surfaces with $h=0.7\,\mu m$, $l=3\,\mu m$. Loss enhancement factor for copper and brass is plotted at right y-axis.

However, one can use average parameters declared by manufacturers to estimate the effect. For $18\,\mu m$ electrodeposited copper used in Taclamplus the roughness height is given as $h=0.7\,\mu m$. The correlation length $l$ is not specified and a typical value of $3\,\mu m$ is taken [18]. Details of the calculations are given in the Appendix.

The calculated loss enhancement factor and quality factor $Q_{CD}$ calculated for the samples of set 2 with and without surface roughness effect presented in Fig. 8. It is demonstrated that metal loss increase due to the roughness by 10% for both copper and brass surfaces. Larger values of $h$ or lower values of $l$ lead to even higher metal loss, see Appendix. The quality factor is reduced for the rough surface and the reduction increases with the frequency as $K_{SR}$ grows.

The effect of surface roughness on the loss tangent is demonstrated for all the samples in Fig. 9. It results in a reduction of the loss tangent by almost 40%. The updated average values of loss tangent are presented in Table IV alongside with the ones calculated for the smooth metallization. The error has been reduced to 8-14%, which presents a significant improvement on the previous result when compared with the specifications.

E. Leakage

Despite number of publications addressing the issue of leakage in the SIW and cavities and proposing clear design rules how to render the effect negligible there is a concern that the effect can be crucial for loss tangent measurement.

Indeed, in order to test our samples for leakage the $Q_{rad}$ has been calculated by simulating the resonators in CST Microwave Studio with lossless dielectric and perfect electric conductor and further extracting the unloaded quality factor which is equal to $Q_{rad}$ as there is no other sources of energy dissipation.

Fig. 10. Quality factor associated with radiation loss for resonators of set#1.

Comparing the quality factor plotted in Fig. 10 with the values of measured loss tangent in Fig. 9 one may conclude that for frequencies below 90 GHz the correction for leakage yields contribution to the loss tangent below 1%. For higher frequencies this goes up to 4%. The effect can be further diminished by careful design.

Bearing in mind the contribution of the uncertainties of the measurements and the effect of metal roughness one may as well discard the correction for leakage as insignificant, thus avoiding any 3D electromagnetic simulations in the method itself.
V. Conclusion

Mm-wave dielectric characterization of printed circuit board materials using a new substrate integrated waveguide resonator technique has been demonstrated in the paper. A launching mechanism using GSG probing of a u-shaped slot is proposed to facilitate measurement and accurate deembedding at millimeter-wave frequencies. Special attention has been paid to the accurate extraction of the unloaded resonant frequency for single-port resonators.

A comprehensive set of test samples was designed and measured over the 60 to 110 GHz frequency range. It has been found that moderate variation of via diameter and via separation does not lead to extracted parameter deviation as long as it is properly accounted for in the effective width calculation of the resonator. The method is also robust with respect to the aperture dimension deviations, making the method less sensitive to etching tolerances hence more readily extendable to deployment at higher frequencies.

The retrieved substrate permittivity is in good agreement with the known nominal value specified by manufacturers at 50 GHz.

It has been shown that the metal roughness effect is crucial for the estimation of the loss tangent. Thus for the smooth metallization assumption the loss tangent extracted is by 34-40% larger than the specified, i.e. 0.0011 c.f. 0.0008. However, once the surface roughness is properly taken into account the mean value loss tangent has become about 0.0007 or only 8-14% lower than specified.

The leakage loss has been shown to have no significant effect on the calculation of the loss tangent provided special care is exercised at the design stage.

Based on these results and the fact that the same geometrical configuration of a cavity can be used for characterization of different substrate thicknesses the authors propose this method for mm-wave testing of printed circuit board substrates as an alternative to the more restrictive microstrip and stripline resonator methods.

APPENDIX

It has been found in [18] that for the case when plane wave is incident on a rough 3D metal surface the loss enhancement factor can be calculated as

\[ K_{SR} = 1 + \frac{2h^2}{\delta^2} \int_{-\infty}^{\infty} dk_x \int_{-\infty}^{\infty} dk_y W(k_x, k_y) Re \left[ \frac{-2j}{\delta^2} - k_x^2 - k_y^2 \right] \]

where \( h \) is rms roughness height, \( \delta = \frac{1}{\sqrt{\sigma \mu \sigma}} \) is skin depth, and \( W(k_x, k_y) \) is spectral density.

Depending on the random profile of the surface two spectral density functions are proposed in [18]. Here we discuss the Gaussian distribution with the correlation length \( l \):

\[ W(k_x, k_y) = \frac{h^2 l^2}{4\pi} \exp \left( -\frac{k_x^2 + k_y^2}{4l^2} \right) \]

The integral can be converted to a one-dimensional one by switching to polar coordinates with the substitution \( r = (k_x^2 + k_y^2)^{1/2} \):

\[ K_{SR} = 1 + \frac{2h^2}{\delta^2} \left( \frac{\delta}{4l} \right) \int_{0}^{\infty} dr \exp \left( -\frac{r}{4} \right) Re \left[ \frac{-2j}{\delta^2} - r \right] \]

The resultant integral is a fast converging one and can be easily calculated numerically. The authors used quadgk procedure from Matlab for the simulations presented in the paper.

In our case the ratio between the substrate thickness and the surface roughness height is more than 100 and the formula above is valid. However, if the ratio is less than 40 more involved expression for the loss enhancement factor in parallel-plate waveguide should be employed [18].

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