Measurement of the Complex Permittivity of Low-Loss Planar Microwave Substrates Using Aperture-Coupled Microstrip Resonators

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Abstract—**This paper describes a technique for the determination of the complex permittivity of low-loss dielectric substrates at microwave frequencies. The technique utilizes an aperture-coupled microstrip resonator fed using a microstrip line in a two layer configuration. The ends of the resonator are shorted in order to avoid radiation. The technique can also be used for the measurement of the complex permittivity of other electronic materials such as thin and thick film materials at microwave frequencies. Nonresonant modes and conductor losses are taken into account in the analysis to improve the accuracy of the results. Analysis procedure as well as experimental results are presented.**

I. Introduction

KNOWLEDGE of the complex permittivity of microwave materials is of prime importance to designers of microwave circuits built using such materials. Any errors in the complex permittivity could lead to significant changes in the overall performance of the designed circuit. Many techniques are available for the measurement of the real part of the complex permittivity (dielectric constant) [1]–[8], however, each technique is useful for certain kinds of materials and certain frequency ranges. Much fewer techniques are available for the accurate measurements of the loss tangent of low-loss materials and almost all of them utilize resonant cavities or other resonators [9]–[12]. In such techniques, the measured resonance frequency and quality factor of the resonator are used to determine the complex permittivity. Most of the resonator techniques used for characterization of microwave substrates use some form of gap-coupled stripline or microstrip line resonators. Gap-coupled microstrip ring resonators were used to determine the effective dielectric constant of planar substrates [13], [14] while gap-coupled stripline resonators [11], [12] were used for the measurement of the dielectric constant and loss tangent. In these techniques, neglecting radiation losses as well as errors in modeling the coupling gaps and end effects introduce errors in the measured permittivity. Furthermore, several fabrication iterations may be necessary in order to obtain the proper gap size to provide acceptable coupling of the resonator to compromise between very strong coupling where radiation loss becomes significant and very weak coupling which leads to difficulty in detecting resonance.

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This paper presents a new technique for the simultaneous measurement of both the dielectric constant and loss tangent. The proposed technique uses an aperture-coupled microstrip resonator with shorted ends in a two-layer configuration, as shown in Fig. 1. The conductivity of the resonator's conductor walls is assumed known. A microstrip feed line is built on the bottom substrate whose dielectric properties are known while the shorted resonator is built on the top substrate whose dielectric constant and loss tangent are to be measured. The two substrates are separated by a ground plane which contains a small circular coupling aperture. The ground plane with the aperture is part of the feed line substrate while the substrate under test has no ground plane, as shown in Fig. 2. The purpose for this arrangement is to make it possible to slide the substrate under test back and forth to improve coupling, if necessary. In effect, this process is equivalent to moving the aperture location underneath the resonator to find the proper location for reasonable coupling. Radiation losses are not present since both ends (radiating edges) are shorted, however, radiation from the opening of the feed line, even though it is small, is taken into account. Losses in the conductor walls of the resonator are also taken into account, assuming their conductivity is known. This method can also be used for characterizing electronic materials such as thin film and thick film materials. In this case, the feed line with the ground plane and aperture are built on a substrate with known properties; the material to be characterized along with the shorted resonator is then printed or deposited on the back of the substrate.

The proposed technique is analyzed taking into account higher order modes excited in the resonator. First, the input impedance and return loss are derived for given resonator dimensions, aperture diameter and location, substrate properties, and feed-line extension. Then, an optimization (iteration) procedure is described which enables the calculation of the aperture diameter, location, and feed line extension for a desired input impedance given substrate properties. This analysis is presented in Section II. The procedure for determination of dielectric constant and loss tangent given resonance frequency and return loss or input impedance along with experimental results is described in Section III. Finally, some conclusions and discussions are presented in Section IV.

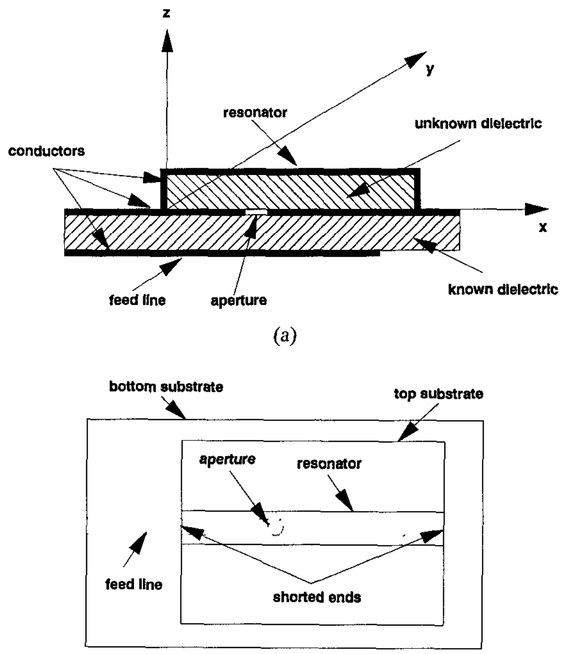

Fig. 1. Aperture-coupled microstrip resonator with shorted ends: (a) Side view. (b) Top view.

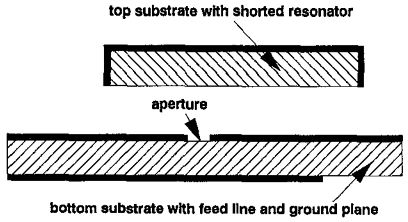

Fig. 2. Illustration of the separate feed line substrate and substrate under test.

II. Analysis Assuming Material Properties are Known

The analysis of the structure in Fig. 1 uses the theory of coupling through small apertures originally proposed by Bethe [15]. Coupling through small apertures can be accounted for by using an equivalent electric current dipole $\vec{J}$, and a magnetic current dipole, $\vec{M}$, located at the center of the aperture. The aperture is then replaced by a perfect conductor, thus splitting the structure into two regions: the feed-line region (region a) and the resonator region (region b). The current dipoles are related to the fields in the feed line region as follows:

$$\vec{J} = j\omega\epsilon_{re}^{a}\epsilon_o\alpha_e\vec{E}_{an}\delta(x-x_o)\delta(y-y_o)\delta(z) \quad (1)$$
$$\vec{M} = -j\omega\mu_o\alpha_m\vec{H}_{at}\delta(x-x_o)\delta(y-y_o)\delta(z) \quad (2)$$

where $\vec{E}_{an}$, $\vec{H}_{at}$ are the normal component of the electric field and the tangential component of the magnetic field to the aperture in region a (the feed line region), ϵ_{re}^{a} is the effective dielectric constant in region a (formulas for ϵ_{re}^{a} are very well known [16]), α_e is the electric polarizability of the aperture, α_m is the magnetic polarizability of the aperture, δ is the impulse function, and $(x_o, y_o, 0)$ are the coordinates of the center of the aperture. The electric and magnetic polarizabilities of a circular aperture of radius r [21] are given by $\alpha_e = (2r^3)/3$ and $\alpha_m = (4r^3)/3$.

The fields in the feed-line region are taken to be that of the dominant quasi-TEM mode. Assuming a perfect open at the end of the feed line, the fields are given by

$$\vec{E}_a = \hat{a}_z e^{-jk^a(x_o+l)}\cos k^a(x-x_o-l) \quad (3)$$
$$\vec{H}_a = \hat{a}_y \frac{j\omega\epsilon_{re}^{a}\epsilon_o}{k^a} e^{-jk^a(x_o+l)}\sin k^a(x-x_o-l) \quad (4)$$

where $k^a = \omega\sqrt{\epsilon_{re}^{a}\epsilon_o\mu_o}$, x_o is the x-coordinate of the center of the aperture, and l is the length of the feed line extension beyond the center of the aperture. The current dipoles $\vec{J}$ and $\vec{M}$ are then easily determined using (1)–(4).

The input impedance seen by the feed line at the aperture is the series combination of the input impedance to the resonator Z_{in1} and the input impedance at the aperture $(x_o, y_o, 0)$, Z_{in2}, of the length l of the feed line extension. The input impedance to the resonator seen by the aperture Z_{in1} can be derived using

$$Z_{in1} = -\frac{1}{2|I_o|^2}\int_v \left(\vec{E}_b \bullet \vec{J}^* + \vec{H}_b^* \bullet \vec{M}\right)dv \quad (5)$$

where $\vec{E}_b$, $\vec{H}_b$ are the electric and magnetic fields in the resonator region (region b), respectively, v is the resonator volume, and I_o is the feed line current at the aperture center. Using Apmere's circuital law, the feed-line current at the aperture is given by $I_o = H_{ay}(x_o)W$ where W is the strip width of the microstrip feed line, and H_{ay} is given in (4). In order to find $\vec{E}_b$ and $\vec{H}_b$, the resonator region is modeled as a cavity surrounded by two electric walls and two magnetic walls with effective width b_e to account for fringing fields. The effective width is obtained using microstrip line design formulas as $b_e = (120\pi h)/(Z_o^b\sqrt{\epsilon_{re}})$ where h is the thickness of the substrate under test, ϵ_{re}^{b} is the effective dielectric constant of the microstrip resonator [16], and Z_o^b is its characteristic impedance calculated using the formula in [20]. The substrate under test is assumed to be electrically thin, i.e., $h \ll \lambda$, λ being the wavelength ($h \leq 0.1\lambda$ is reasonable). As a result of this assumption, the fields in the resonator region are assumed independent of z. The resonant modes are then of the TM_{mn} type (the third subscript of the resonant modes TM_{mn} is zero and therefore it is dropped to simplify the notation). The fields in the resonator region can then be written as [17]

$$\vec{E}_b = \sum_n\sum_m B_{mn}\vec{E}_{mn} + \sum_n\sum_m C_{mn}\vec{E}_{mn} \quad (6)$$
$$\vec{H}_b = \sum_n\sum_m B_{mn}\vec{H}_{mn} + \sum_n\sum_m G_{mn}\vec{H}_{mn} \quad (7)$$

where the first double summation term in equations (6) and (7) represent the fields due to the electric current dipole $\vec{J}$ while the second double summation term represents the fields due to the magnetic current dipole $\vec{M}$, $\vec{E}_{mn}$, and $\vec{H}_{mn}$ are the source-free cavity modes, and B_{mn}, C_{mn} are coefficients to be evaluated. The coefficients B_{mn} and C_{mn} can be evaluated using [17]

$$B_{mn} = D_{mn}\int_v \vec{J} \bullet \vec{E}_{mn}^* \, dv \quad (8)$$

$$C_{mn} = D_{mn} \int_v \vec{M} \bullet \vec{H}^*_{mn}\, dv \tag{9}$$

$$D_{mn} = \frac{jk^b \sqrt{\epsilon^b_{re}\epsilon_o\mu_o}}{(k^b)^2 - (k_{mn})^2} \tag{10}$$

where $k_{mn} = \sqrt{\left(\frac{m\pi}{a}\right)^2 + \left(\frac{m\pi}{b_e}\right)^2}$ and $k^b = \omega\sqrt{\epsilon^b_{re}\epsilon_o(1-j\delta_e)}$; a is the length of the resonator and b_e is the effective width of the resonator. To take into account losses in the conductor walls of the resonator, the effective loss tangent δ_e is given by $\delta_e = \delta_d + (1/Q_c)$ where δ_d is the loss tangent of the resonator dielectric material and Q_c is the Q-factor due to conductor loss in the resonator walls. The conductor loss is mostly in the top and bottom walls of the resonator since the resonator is assumed to be electrically thin, hence $Q_c = h\sqrt{\frac{\omega\mu_o\sigma_c}{2}}$. The TM_{mn} cavity modes are given by

$$\vec{E}_{mn} = \hat{a}_z A_{mn} \sin\frac{m\pi}{a}x \cos\frac{n\pi}{b_e}y \tag{11}$$

$$\vec{H}_{mn} = -\frac{j\omega\epsilon^b_{re}\epsilon_o}{(k_{mn})^2} A_{mn}\Big[\hat{a}_x \frac{n\pi}{b_e} \sin\frac{m\pi}{a}x \cdot \sin\frac{n\pi}{b_e}y + \hat{a}_y \frac{m\pi}{a} \cos\frac{m\pi}{a}x \cos\frac{n\pi}{b_e}y\Big] \tag{12}$$

where A_{mn} is a normalizing constant given by $A_{mn} = \left(g_{mn}/\sqrt{ab_e\epsilon^b_{re}\epsilon_o h}\right)$; $g_{mn} = \sqrt{2}$ if either m or n equals zero and $g_{mn} = 2$ if neither m nor n equals zero. Using (1)–(4), (8), (9), (11), and (12), the following expressions for B_{mn} and C_{mn} are derived:

$$B_{mn} = jD_{mn}\omega\epsilon^a_{re}\epsilon_o\alpha_e e^{-jk^a(x_o+l)} \cos k^a l A_{mn} \cdot \sin\frac{m\pi}{a}x_o \cos\frac{n\pi}{b_e}y_o \tag{13}$$

$$C_{mn} = -jD_{mn}k^a\alpha_m e^{-jk^a(x_o+l)} \sin k^a l \frac{\omega\epsilon^b_{re}\epsilon_o}{(k_{mn})^2}\frac{m\pi}{a} \cdot A_{mn} \cos\frac{m\pi}{a}x_o \cos\frac{n\pi}{b_e}y_o \tag{14}$$

where D_{mn} is given in (10). Using (5)–(7), the following expression for the resonator input impedance seen at the aperture can be derived:

$$Z_{in1} = -\frac{1}{2|I_o|^2}\sum_n\sum_m\left[\frac{|B_{mn}|^2}{D^*_{mn}} + \frac{|C_{mn}|^2}{D_{mn}} + \frac{2C_{mn}B^*_{mn}}{|D_{mn}|^2}\mathrm{Re}\{D_{mn}\}\right] \tag{15}$$

where I_o was given earlier.

The input impedance, at the aperture $(x_o, y_o, 0)$, of the length l of the feed-line extension Z_{in2} is easily found using transmission line theory. Radiation and fringing from the open end of the feed line extension are modeled by a load impedance located at the end of the feed line extension. The real part of the load impedance is found by calculating the radiation resistance of the slot at the open end and the imaginary part is accounted for using an effective line extension calculated using microstrip design formulas [18]. The overall input impedance Z_{in} seen by the feed line at the aperture is then given by the series combination of Z_{in1} and Z_{in2}. Having obtained the input impedance, the evaluation of the reflection coefficient and return loss is then straight forward.

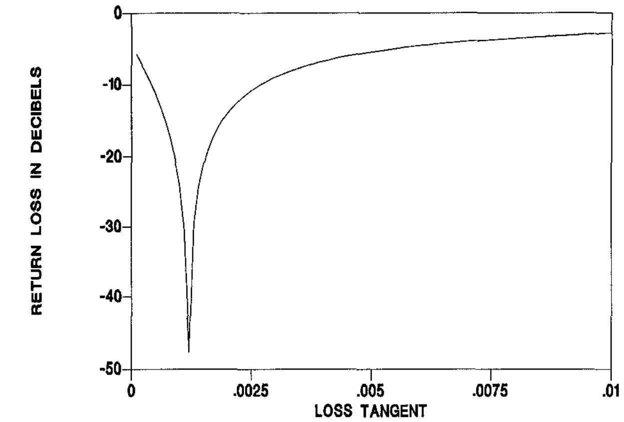

Fig. 3. Sensitivity of return loss to loss tangent of resonator substrate.

Two computer programs were written implementing the aforementioned equations to study the feasibility of this approach. One program is an analysis program which calculates the input impedance and return loss given material properties, resonator dimensions, aperture location, aperture diameter, and feed-line extension. The other program is an optimization program which enables the calculation of the aperture location, aperture diameter, feed-line extension, and resonator dimensions given the material properties, desired resonance frequency, and desired input impedance. The optimization procedure minimizes the following error function:

$$F(x_o, y_o) = |Z_{in}(x_o, l, r) - Z_f|^2 \tag{16}$$

where Z_{in} is the calculated input impedance for a given aperture location x_o, aperture radius r, and feed-line extension l. Z_f is the characteristic impedance of the feed line (the desired value of the input impedance). The aperture coordinate y_o is always taken to be in the middle of the resonator width. In the iterative process, first initial values for x_o, l, and r are assumed. Then, the input impedance is calculated. The error function in (16) is evaluated and, if a specified tolerance is not met, new estimates for x_o, l, and r are obtained using Powell's hybrid method [19].

To study the sensitivity of the return loss to variations in the loss tangent of the resonator substrate material, the optimization program was used to design an aperture-coupled resonator using RT/Duroid 5870 as the resonator substrate which has a dielectric constant of 2.33 and loss tangent of 0.0012 at 10 GHz (according to the manufacturer). The obtained results were then used in the analysis program to calculate the return loss at the resonance frequency as a function of the loss tangent of the resonator substrate. The result is shown in Fig. 3 which suggests that the return loss at resonance is extremely sensitive to the loss tangent as expected in a resonator. The correct value of the return loss (0.0012) corresponds to the minimum of the return loss in Fig. 3. This is very important in order to be able to precisely determine the loss tangent of low-loss materials using this technique. Changes in the dielectric constant will mainly affect the resonance frequency.

III. Determination of Complex Permittivity

To obtain the complex permittivity given the resonance frequency and return loss or input impedance, an optimization process, similar to the one described in the previous section, was first attempted. However, this optimization process needed initial estimates for the complex permittivity that are very close to the correct value for the iterative process to converge to the minimum of error function. The reason for this convergence difficulty is the very sharp resonance exhibited by the high-Q resonator which results in a very deep valley in the error function. Therefore, an alternative approach was used where the analysis computer program described earlier was modified to output two look-up tables: one for the dielectric constant and the other for the loss tangent. First, initial estimates for the dielectric constant ϵ_{ro}^{b} and loss tangent δ_{do} are calculated using the measured values of the resonance frequency f_r, loaded Q-factor Q_L, and microstrip design formulas as follows:

$$\epsilon_{ro}^{b} = \frac{2\epsilon_{reo}^{b} + T - 1}{1 + T} \tag{17}$$

$$\delta_{do} = \frac{1}{Q_L} - \frac{1}{Q_c} \tag{18}$$

where $\epsilon_{reo}^{b} = (\lambda_o^2/4a^2)$, $T = 1/\sqrt{1 + 12h/b}$, λ_o is the wavelength in free space, and Q_c is the Q-factor due to conductor loss given earlier. Then the program calculates the input impedance and return loss at the resonance frequency for a desired range of dielectric constant around ϵ_{ro}^{b} using the analysis results of the previous section, thus creating a look-up table for the dielectric constant. The loss tangent used to create this table is the rough estimate δ_o, obtained using the loaded Q-factor. Since the substrate materials under consideration are low-loss, using precise value for the loss tangent to create the dielectric constant table is not necessary because the shift in resonance is insignificant for small variations in the loss tangent. The correct value of the dielectric constant is the value that yields minimum return loss. The dielectric constant range can then be narrowed (zoomed), if desired, to improve accuracy. Having obtained the correct value of the dielectric constant, a similar look-up table, at the resonance frequency, for the loss tangent versus return loss is then created. The correct value of the loss tangent is then simply the value that yields a calculated return loss very close to the measured value. Input impedance can be used instead of return loss, if desired.

To validate this technique, two experiments were carried out. In the first experiment, the optimization program was used to design an aperture-coupled resonator using RT/Duriod 5870 substrates which has a dielectric constant of 2.35 at 1 MHz and 2.33 at 10 GHz and loss tangent of 0.0005 at 1 MHz and 0.0012 at 10 GHz (according to the manufacturer). The thickness of the resonator substrate was 0.158 cm while that of the microstrip feed line was 0.0787 cm. The optimization program described in the previous section was used to design an aperture-coupled resonator having its first resonance at 5 GHz using these substrates assuming a resonator width of 1 cm and feed-line characteristic impedance of 50 Ω. The results of the optimization program are: $x_o = 0.25614$ cm, $l = 1.06642$ cm, $r = 0.123$ cm. The resonator length $A =$ 2.0508 cm. The aperture-coupled resonator was then built using standard photolithographic etching and the return loss was measured using a Marconi 6500 Scalar Network Analyzer. The measured resonance frequency was 4.99 GHz, the measured return loss there was −36 DB, and the measured loaded Q-factor was 490. Using the loaded Q-factor value the initial estimate of the loss tangent δ_{do}, was calculated to be 0.00163. Part of the calculated lookup table for the dielectric constant at a frequency of 4.99 GHz is given in Table I. From Table I, it is clear that the dielectric constant is 2.336 which yields minimum value of return loss at the resonance frequency. Using this value of the dielectric constant, a lookup table for the loss tangent was created and part of it is given in Table II. From Table II, the correct value of the loss tangent is 0.001235 which resulted in a return loss of −36.05 dB. These results agree very well with the manufacturer's specifications. A flow-chart illustrating the above procedure is given in Fig. 4.

TABLE I
Dielectric Constant Lookup Table for RT/Duroid 5870.

dielectric constant	return loss
2.33000	−5.16148
2.33100	−6.34527
2.33200	−7.90952
2.33300	−9.99084
2.33400	−12.74824
2.33500	−16.14561
2.33600	−18.37579
2.33700	−16.29086
2.33800	−12.89631
2.34900	−10.11931
2.34000	−8.01977

TABLE II
Loss Tangent Lookup Table for RT/Duroid 5870.

loss tangent	return loss
0.001200	−45.636231
0.001205	−43.454028
0.001210	−41.699201
0.001215	−40.236461
0.001220	−38.984567
0.001225	−37.891537
0.001230	−36.922296
0.001235	−36.052133
0.001240	−35,263016
0.001245	−34.541396
0.001250	−33.876845

In the second experiment, RT/Duriod 5880 substrates having a thickness of 0.2286 cm were used. These substrates have a dielectric constant of 2.2 and a loss tangent of 0.0009 at 10 GHz, according to the manufacturer. A resonator was built having a length of 1.02 cm, a width of 2 cm, aperture radius $r = 0.325$ cm, aperture location $x_o = 0.325$ cm, feed-line extension $l = 0.76$ cm, and feed-line width of 0.7 cm. The measured resonance frequency was 9.978 GHz, the measured return loss was −9.5 DB, and the measured Q-factor was 351. Two lookup tables were created following the same procedure as explained in the flow-chart in Fig. 4. Parts of these look-up tables are given in Table III and Table IV. From these tables

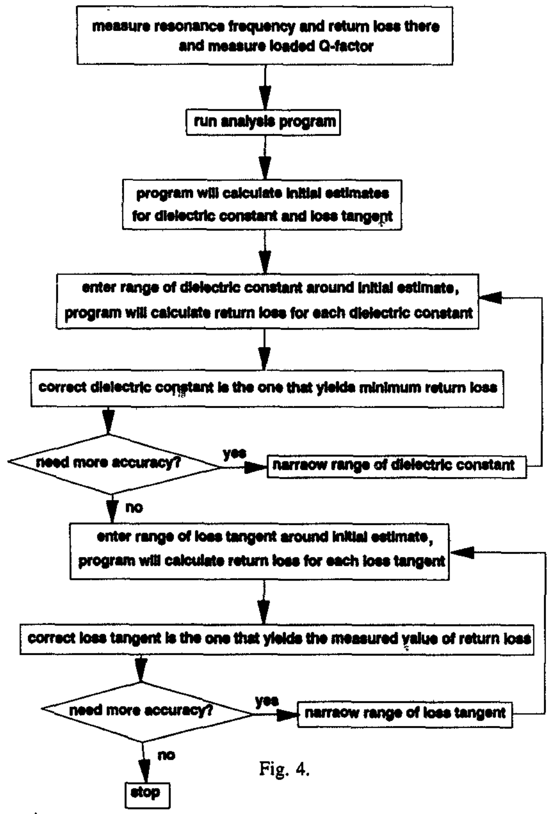

Fig. 4. Flowchart for the calculation of the complex permittivity.

TABLE III
DIELECTRIC CONSTANT LOOKUP TABLE FOR RT/DUROID 5880.

dielectric constant	return loss
2.21000	−14.83351
2.21100	−15.60294
2.21200	−16.28394
2.21300	−16.81536
2.21400	−17.13670
2.21500	−17.20955
2.21600	−17.03545
2.21700	−16.65499
2.21800	−16.12914
2.21900	−15.51810
2.22000	−14.86939

it is clear that the dielectric constant is 2.215 (yields minimum return loss) and the loss tangent is 0.00977 (yields return loss close to the measured value), which are very close to the manufacturer's specifications.

To characterize an unknown substrate material, the optimization program can be used first to obtain x_o, l, and r using a rough estimate of the dielectric constant of the substrate material assuming it to be lossless. A rough estimate of the dielectric constant is very easy to obtain using simple techniques. This procedure is used to assure good coupling to the resonator and, thus, avoid unnecessary fabrication iterations. If needed, the coupling can be further improved easily by just sliding slightly the resonator layer back and forth along the feed line until satisfactory coupling is obtained. In effect, this process is equivalent to moving the aperture location underneath the resonator to find the proper location

TABLE IV
LOSS TANGENT LOOKUP TABLE FOR RT/DUROID 5880.

loss tangent	return loss
0.000970	−9.453587
0.000971	−9.461266
0.000972	−9.468946
0.000973	−9.476629
0.000974	−9.848313
0.000975	−9.491998
0.000976	−9.499686
0.000977	−9.507375
0..000978	−9,515066
0.000979	−9.522759
0.000980	−9.530453

for reasonable coupling. Actually, a fixture using precise micrometers can be designed to facilitate precise testing of substrates using this technique.

As mentioned in the introduction, this method can also be used for characterizing other materials such as thin film and thick film materials. In this case, the feed line with the ground plane and aperture are built on a substrate with known properties; the material to be characterized along with the shorted resonator is then printed or deposited on the back of the substrate. It would be interesting to experimentally test this technique for such materials; unfortunately, the author does not have access to a fabrication facility to build resonators using such materials.

Possible sources of errors in this technique include possibility of the presence of an air gap between the feed line substrate and the substrate under test, error in determining the coordinates of the coupling hole (alignment errors), as well as measurement errors. Also, one has to keep in mind that the analysis is valid for electrically thin substrates and small coupling apertures.

IV. CONCLUSIONS

A new technique for the measurement of the dielectric constant and loss tangent of low-loss dielectric substrates using aperture-coupled microstrip resonators with shorted ends was presented. Derivation as well as analysis and optimization programs were described. The validity of the technique was confirmed experimentally using substrates with known dielectric properties. This technique can be used for testing thick film and thin film materials used at microwave frequencies. Also, a precise fixture using micrometers can be designed to facilitate precise testing of substrate materials using this technique, as explained in Section III.

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