Measurement of Planar Substrate Uniaxial Anisotropy

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***Abstract*—A new technique to measure uniaxial anisotropy in planar substrates is described. The technique uses a single dual-mode resonator. Each mode of the resonator has a different distribution of horizontal (parallel to the substrate surface) and vertical (perpendicular to the substrate surface) directed fields. Using an electromagnetic analysis of the dual-mode resonator, the resonant frequencies of the modes are space mapped to the horizontal and vertical dielectric constants. The space mapping allows the anisotropic dielectric constants to be extracted from the measured resonant frequencies. It is also suggested that this technique can be applied to magnetic uniaxial anisotropy as well as to magnetic and electric loss tangent anisotropy. The dual and quad "RA" resonators are introduced in this paper. A measurement of FR4 (a common anisotropic epoxy-glass weave composite substrate) with a detailed error analysis illustrates the technique.**

***Index Terms*—Anisotropy, dielectric constant, dispersion, electromagnetic (EM) analysis, FR-4, FR4, measurement, method of moments (MoM), printed circuit board (PCB), transmission line, uniaxial.**

I. Introduction

A COMMON low-cost printed circuit board (PCB) material is FR4. With increasing digital speeds and with need for low-cost microwave substrates, precise characterization of FR-4 is important. Published measurements in the microwave literature of FR4 are rare, but exceptions include [1]–[5]. In particular, [2] has some similarity to this technique and also provides a detailed bibliographic overview, which we supplement by noting [6], a precise means of measuring a substrate using a resonant cylindrical waveguide cavity.

FR4 is composed of a woven glass cloth embedded in epoxy. As such, one might expect the horizontal electric field (parallel to the substrate surface) to experience a different dielectric constant as compared to the vertical (normal to the substrate surface) electric field. We have found only one published measurement of FR4 anisotropy in the microwave literature [5], which uses circular patch resonators combined with rigorous electromagnetic (EM) analysis. There is a mention of anisotropy in [1]; however, that involves measuring microstrip resonators in two horizontal directions, and no anisotropy was found greater than the typical 4% variations found in other measurements. No measurement of anisotropy in the critical vertical direction is reported.

Anisotropy in PTFE/glass fiber composite substrates is reported in two nonpeer reviewed papers [7] and [8] using stripline resonators [9]. The vertical dielectric constant is determined by laminating multiple unmetallized substrates together, then fabricating new planar substrates by slicing the lamination stack in a cross section. For this new substrate, the original vertical dielectric constant becomes the new horizontal dielectric constant and the horizontal stripline measurement is repeated. Differences on the order of 5% are found. However, stripline resonators excite a mixture of horizontal and vertical electric fields, measurement of which yields a weighted average of dielectric constants (even with wide resonators), making the material appear less anisotropic than it really is. Neither [7], nor [8] discuss this situation.

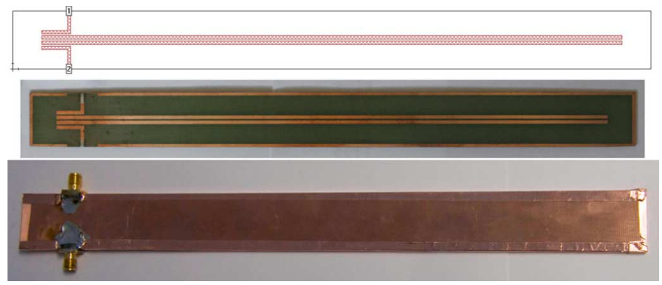

Fig. 1. RA resonator: (*top*) layout in Sonnet, (*middle*) fabricated resonator, and (*bottom*) completed resonator. The resonators are 10 in long. The input/output coupled region is 0.5 in long.

Anisotropy changes the microstrip effective dielectric constant dispersion characteristic and this can be used to determine anisotropy by EM based optimization of the EM analysis of anisotropy to match the measured dispersion [10]. This assumes that the anisotropy is frequency independent, so the frequency dependence of the anisotropy is not measured. Alumina is shown to have significant anisotropy in [10].

Bulk measurement of anisotropy for low loss substrates over broad bandwidths is possible using various waveguide resonators, with [11] and [12] as excellent examples. Anisotropic measurements excite the sample with fields predominantly in one direction at a time. However, due to the large size of the resonators, these techniques are sometimes not useful at the lower frequencies of interest for materials like FR-4. Also, they measure average bulk anisotropy, yielding misleading results for planar resonators on inhomogeneous substrates like FR-4 as discussed in Section VI.

With the ability to both measure (as described in this paper) and EM analyze anisotropic substrates; there is now no need to compromise the design of a substrate material with respect to cost or mechanical/electrical characteristics in order to achieve a higher (unneeded) order of isotropy.

In this work we explore using a single planar dual mode resonator, Fig. 1, to extract precise and complete uniaxial dielectric measurements. Lamination of substrates for a second resonator, as in [7] and [8], is unneeded. Each mode of this RA

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resonator (the name is derived from the authors' initials) has a different mix of vertical and horizontal electric fields. The uniaxial dielectric constants are determined by space-mapping the substrate dielectric constants to the resulting resonant frequencies. This mapping is dependent on and determined by precise numerical EM analysis, as initially proposed in [13].

Several substrate measurement techniques depend on mapping of measured data to some kind of model, either empirical or numerical EM. For example, [2] relies on an unshielded method of moments (MoM) tool. The dielectric constant in the EM tool is adjusted until it matches the measured data. This is undesirable, especially for an anisotropic dielectric, as a large number of EM analyses are required to determine both dielectric constants at multiple frequencies. In addition, the error in the EM tool translates directly into measurement error. Reported measurements typically neglect error analysis. For example, [2] relies only on a high level of "confidence" in the EM tool, a subjectively assigned quality that should properly be quantified, for example, by convergence analysis. In contrast to relying on "confidence," we quantify all pertinent error sources. In fact, the majority of effort expended has been in the quantitative evaluation of error.

The degree of anisotropy found for the particular sample measured in this paper is small; however, FR-4 from different manufacturers could have widely different degrees of anisotropy as anisotropy is not considered, nor even measured by different manufacturers. The technique described in this paper allows the designer to judge the importance of anisotropy and allows the manufacturer to control for anisotropy based on precise easily performed measurements.

II. Space-Mapped Approach

We illustrate our approach by describing the usual microstrip resonator approach for isotropic substrate dielectric constant measurement and place it in terms of space mapping. This approach is then generalized to uniaxial anisotropy.

To measure an isotropic dielectric constant, we construct, for example, a microstrip resonator, a length of microstrip line that is an integer multiple of half-wavelengths long at the desired frequencies of measurement. We then add very light coupling structures (short lengths of line placed at some modest distance from the resonator) and measure the resonant frequencies. This information is still insufficient to determine the dielectric constant as the effective dielectric constant of the microstrip resonator is a weighted average of the substrate dielectric constant and that of air. In addition, there is fringing capacitance off both ends that lowers the resonant frequencies.

These problems are resolved by proper use of EM analysis. The resonator is analyzed using EM analysis combined with a best guess at the value of the substrate dielectric constant. The resonant frequencies extracted from the EM analysis (with a known dielectric constant) and the measured resonant frequencies determine the dielectric constant of the substrate.

Specifically, we assume that the known dielectric constant (ε_a, we use relative dielectric constants) in the EM analysis is proportional to the inverse of the EM analysis resonant frequency (f_a) squared as follows:

$$A\varepsilon_a = f_a^{-2}. \tag{1}$$

Solving for the inverse of the constant of proportionality,

$$A^{-1} = \varepsilon_a \left(f_a^{-2}\right)^{-1}. \tag{2}$$

Given the measured resonant frequency (f_m) of the microstrip resonator, we solve for the dielectric constant (ε_m) that corresponds to the measured resonant frequency

$$\varepsilon_m = A^{-1} f_m^{-2}. \tag{3}$$

This assumes that the EM analysis field configuration is nearly the same as the field configuration for the slightly different dielectric constant of the measurement. If desired, a match between EM and measurement can be realized by repeating the procedure using the newly determined value of ε_m. In addition, a slightly improved value for ε_m is then realized. This approach is directly affected by error in the EM analysis resonant frequency. For a complete measurement, this and other errors must be quantified.

The above approach is easily cast in the framework of space mapping [14]. The EM analysis is the "fine" model (i.e., the very accurate model that requires a relatively long time to evaluate) and the constant A^{-1} in (2) is the space mapping. The "coarse" model, the model that is quickly evaluated and corresponds to the fine model over a range of input parameters (the measured resonant frequency), is represented by (3).

III. Generalization to Uniaxial Anisotropy

For uniaxial anisotropy, we use a dual-mode stripline resonator (Fig. 1). A stripline resonator is used because the resonant frequency is more strongly dependent on the substrate material than a microstrip resonator. As described in Section V, error considerations suggest it might be possible to get better results using a microstrip resonator.

While this resonator is dual mode, the resonant frequencies of the two modes are typically not apparent from a two-port analysis or measurement because they are close to each other and strongly coupled to each other. In fact, any attempt to precisely measure even- and odd-mode resonances directly from the two-port data will fail. Even if separate resonances are seen, they "pull" each other's frequencies, corrupting the measurement. Rather, we take the two-port measurement and the analysis and convert both to two one-ports (Fig. 2). In this case, connection for the even mode (Fig. 2, top) completely suppresses the odd mode. The resulting data (to the degree that the resonator is symmetric) is as though we had built a single line resonator with a perfect magnetic conducting (PMC) wall close to the resonator. When we connect the two-port data for the odd mode (Fig. 2, bottom), it is as though we built a second single mode resonator with a perfect electric conducting (PEC) wall close to the resonator. Thus, neither even, nor odd modes affect each other in any way.

The measured resonators include the connectors so the connector models of Fig. 2 are not included when converting measured data. For this work, we use AWR Microwave Office, but any circuit theory tool may be used. It is important to float the ground terminal of the resonator and connector models for the odd mode as shown. Note that with these connections, the even

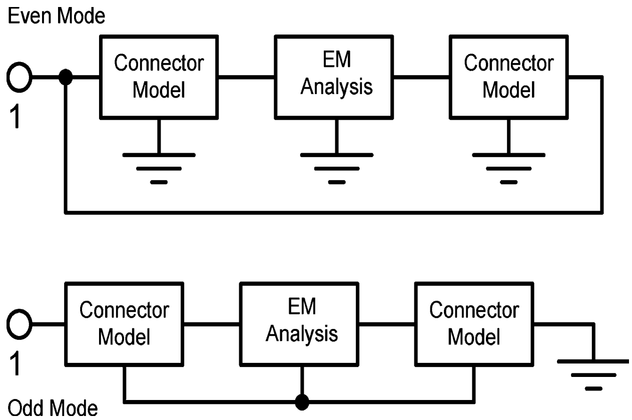

Fig. 2. Schematics for converting a symmetric two-port (EM analysis) into two one-ports yielding the even and odd modes. It is critical that the ground terminals are explicit and tied together and not connected to global ground for the odd mode.

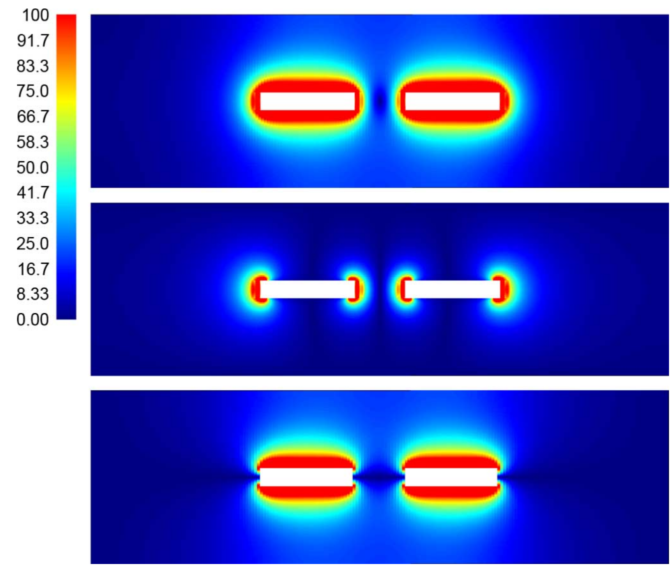

Fig. 3. Even-mode electric field of coupled stripline. Fields are taken in the transverse plane at an open end with 1 V applied between both lines and ground. The scale on the left is in volts per meter. The top image is the total electric field, the middle image is the magnitude of the horizontal component. The bottom image is the magnitude of the vertical component (from [10]).

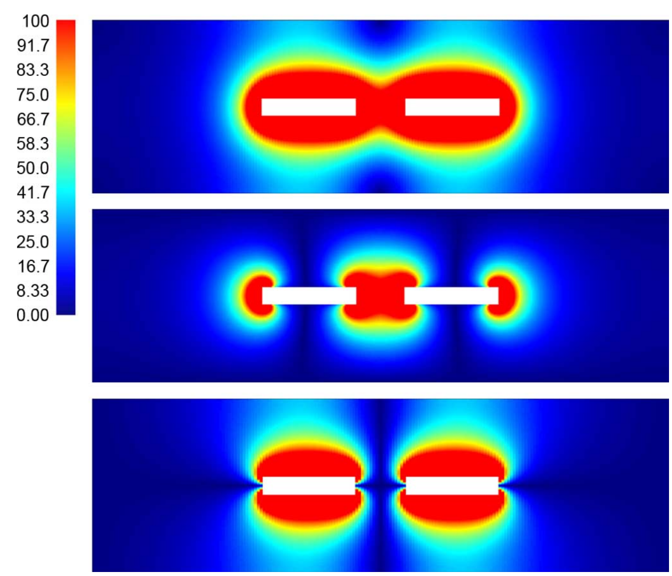

Fig. 4. Odd-mode electric field of coupled stripline. One volt is applied be- tween the two lines. All other information is the same as Fig. 2. Fields are stronger for the odd mode because the odd-mode impedance is lower, resulting in higher current, from [10].

mode result is normalized to 25 Ω and the odd mode is normalized to 100 Ω. Thus, this single measurement yields one-port data for both the even and odd modes from which the modal resonant frequencies are determined.

In the cross-sectional fields for the even and odd modes (Figs. 3 and 4), one can qualitatively see that vertical and horizontal electric fields are present in different ratios between the two modes. For the case of uniaxial anisotropy, we assume that a weighted sum of the horizontal (subscript "h") and vertical (subscript "v") dielectric constants is proportional to the inverse of the even (subscript "e") and odd (subscript "o") resonant frequencies squared

$$A_{he}\varepsilon_h + A_{ve}\varepsilon_v = f_e^{-2} \tag{4}$$
$$A_{ho}\varepsilon_h + A_{vo}\varepsilon_v = f_o^{-2}. \tag{5}$$

If we evaluate two EM cases (subscripts "a" and "b") of dual mode stripline resonators, each with different uniaxial anisotropic dielectric constants (selected to be linearly independent), we have

$$\begin{bmatrix} A_{he} & A_{ve} \\ A_{ho} & A_{vo} \end{bmatrix} \begin{bmatrix} \varepsilon_{ha} & \varepsilon_{hb} \\ \varepsilon_{va} & \varepsilon_{vb} \end{bmatrix} = \begin{bmatrix} f_{ea}^{-2} & f_{eb}^{-2} \\ f_{oa}^{-2} & f_{ob}^{-2} \end{bmatrix}. \tag{6}$$

This is the uniaxial anisotropic generalization of (1). Next, the matrix $\mathbf{A}^{-1}$ is evaluated in analogy with (2). Finally, measured dielectric constants (case "m") are

$$\begin{bmatrix} \varepsilon_{hm} \\ \varepsilon_{vm} \end{bmatrix} = [\mathbf{A}]^{-1} \begin{bmatrix} f_{em}^{-2} \\ f_{om}^{-2} \end{bmatrix} \tag{7}$$

in analogy with (3) with the same limitations and error considerations as described above. The $\mathbf{A}$ matrix is evaluated independently for each pair of even- and odd-mode resonances,

IV. Measurements

Two resonators were built on FR4, both nominally identical. The resonator lengths are 254 mm (10 in), all linewidths are 1.524 mm (60 mil, 1 mil = 0.001 in), all gaps are 0.762 mm (30 mil), and the input/output coupling sections are 1.27 mm (500 mil) long.

Subminature A (SMA) connectors are used. To characterize the connectors, we measured the connectors both open and short circuited. A model of the connector was then created and parameters adjusted to fit both open- and short-circuit data at all frequencies simultaneously. The resulting model is a 52.2-Ω Z_0 lossless line, 10.022 mm (420.5 mil) long (free space length), in parallel with a 5300-Ω resistor. A 0.25-pF capacitor to ground on the circuit side of the connector completes the model. The

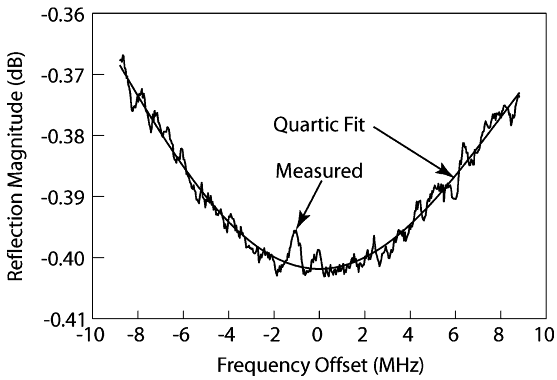

Fig. 5. Performing a least squares quartic polynomial fit on data around a resonance allows accurate extraction of the resonance even in the presence of considerable noise. This is the seventh even-mode resonance centered at 2046.172 MHz.

model matches both the open- and short-circuit measured data over the entire frequency range to within measurement noise. This model is used to embed all EM analysis results, as in Fig. 2, prior to comparison with measurements. Alternatively, the network analyzer could be calibrated using on-board calibration standards and the EM results used directly.

A spreadsheet (Microsoft Excel)[1] was written to process all data, determine $\mathbf{A}^{-1}$, and evaluate (7). In addition, the spreadsheet automatically finds resonances in the data and performs a quartic least squares fit to estimate each resonant frequency. An example result, Fig. 5 shows the seventh even-mode resonance (higher order resonances have more noise). This figure illustrates how a least squares fit extracts accurate resonant frequencies even in the presence of noise.

For evaluation of dielectric constants, measured resonant frequencies are compared with results from an EM analysis of the same resonator. Substrate thickness is 1.481 mm (58.3 mil) with a 0.01778-mm (0.7 mil) air gap (equal to the metal thickness) between the top and bottom substrates. Both thicknesses are measured. EM analysis uses a two-sheet model for metal thickness. Additional sheets are not needed because the metal thickness is much less than the coupled line gapwidth [15]. The manufactured resonators include a metal ring around the edge to act as a spacer to maintain the air gap. The air gap is in both the measured resonator and in the EM analysis. A loss tangent of 0.01 for the substrate and metal conductivity of 5.8×10^7 S/m for the metal is used. The two EM analysis cases of (6) use dielectric constants $\varepsilon_{ha} = \varepsilon_{va} = 3.9$ and $\varepsilon_{hb} = 3.7$ and $\varepsilon_{vb} = 4.1$.

V. Results and Error Evaluation

Table I shows the resonance frequencies for seven modes for the EM analysis cases a and b, and for the R1 resonator measurement, case m. Note that even though case a is isotropic, the even- and odd-mode resonances are different. This is due to the end capacitance being different between the two modes [the even and odd modes have different field configurations (Figs. 3 and 4)].

Table II shows the relative dielectric constants extracted from the resonant frequencies for both resonators. The smoothness as a function of frequency to three digits suggests high accuracy. However, there is a small "roughness" in the fourth digit. Regardless, a detailed error evaluation is required in order to establish confidence in any measurement. All conclusions are *italicized* below.

TABLE I
Even- and Odd-Mode Resonant Frequencies (in Megahertz)

#	f_{ea}	f_{oa}	f_{eb}	f_{ob}	f_{em}	f_{om}
1	297.495	298.682	292.399	296.628	290.547	293.055
2	595.458	597.848	585.236	593.737	583.007	587.729
3	893.578	897.168	878.218	890.996	876.102	883.166
4	1191.830	1196.812	1171.323	1188.554	1169.260	1178.374
5	1490.227	1496.953	1464.562	1486.528	1459.354	1469.836
6	1788.782	1797.019	1757.945	1784.431	1752.633	1764.573
7	2087.504	2096.871	2051.480	2082.162	2046.172	2058.951

TABLE II
Extracted Uniaxial Dielectric Constants for Resonators R1 and R2

#	MHz	R1 ε_h	R1 ε_v	R2 ε_h	R2 ε_v
1	290	3.937	4.117	3.878	4.106
2	590	3.934	4.094	3.889	4.077
3	880	3.926	4.081	3.876	4.075
4	1170	3.935	4.074	3.906	4.078
5	1470	3.979	4.083	3.923	4.075
6	1760	3.990	4.076	3.936	4.072
7	2060	4.001	4.070	3.926	4.063

Eight measurements were performed for each resonator. Each measurement is itself an average of 16 measurements automatically performed and averaged by the network analyzer. All measurements use the same network analyzer calibration. The average extracted dielectric constant is reported in Table II. The sample standard deviation (not listed) gives an indication of the sensitivity to measurement noise. The maximum standard deviation is 0.0013. Nearly all the standard deviations are about 0.0001. *Thus, measurement noise is of small importance for the data of* Table II.

Error in the underlying EM analysis, Sonnet,[2] and [16], directly translates into error in the extracted dielectric constant. The principle error in the EM analysis is error due to cell (i.e., mesh) size [17]. The cell size is 0.127×0.00635 mm (5.0×2.5 mil) (length $\times$ width). Fig. 6 shows a detailed view of the actual mesh near one end of the resonator. This technique is critically dependent on accurate calculation of resonant frequencies. The critical mesh dimension is cell length. "Cell size" refers to the smallest possible subsection size. To quantify the effect of EM analysis error, we double the cell length (thus increasing error) for the EM data (cases a and b) and repeat the extraction. When doubling the cell size, we also halve the meshing frequency (10 GHz for the original cell length, 5 GHz for new cell length) so that the longest subsections (Fig. 6) double in length. Extracted dielectric constants increase by 0.0006 (horizontal) and 0.0008 (vertical) at the first resonance, and by 0.004 (horizontal) and 0.002 (vertical) at the seventh resonance. Error increases smoothly from low to high frequency.

To check the rate of convergence, we double the cell length again, to 0.508 mm (20 mil), while keeping the cell width con-

[1][Online]. Available: http://www.sonnetsoftware.com/support/downloads/publications

[2][Online]. Available: http://www.sonnetsoftware.com

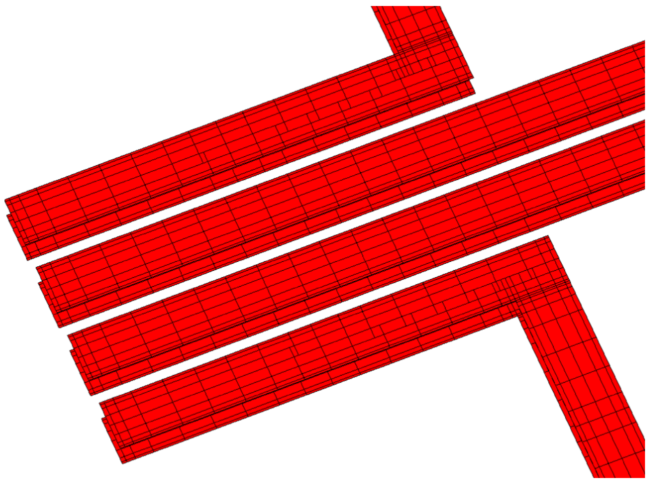

Fig. 6. Fine meshing is required in the EM analysis. The minimum subsection size (i.e., the cell size) used here is 5 × 2.5 mil. The input coupling sections are 0.5 in long. The complete resonators (see Fig. 1) are 10 in long. Two sheets are used; the vertical vias connecting all edges not shown. Vertical dimension magnified 44 times.

stant and halving the meshing frequency to 2.5 GHz. The increased cell length results in the extracted dielectric constant change precisely doubling at the first resonance, as expected for the typical linear convergence we often see in this EM analysis. However, at higher frequencies, super-linear convergence is seen. At the seventh resonance, the error change for this final doubling of cell length is 4–5 times that of the initial doubling. Thus, for higher order resonances, we cannot simply subtract the estimated error. *We estimate the cell size error is bounded by one-quarter the largest change in the dielectric constant seen in the initial cell length doubling, or 0.001 with much smaller error at lower frequencies.*

The meshing used in the EM analysis is particularly important if this technique is to be successful. We have attempted the common approach of meshing the line one subsection wide, and the slightly more advanced approach of including simple narrow edge mesh subsections. Such meshing yields basically unusable results due to large (with respect to our requirements) EM analysis error. *As shown in Fig. 6, meshing must be fine in corners and along edges and it can gradually grade to larger interior subsections.*

In directly viewing the original two-port measurements, we could see up to 2° of phase and several hundredths of a decibel of magnitude difference between S_{11} and S_{22}. If the measurements were perfectly symmetrical, then both S-parameters would be identical. The differences are likely due to network analyzer noise, small mechanical asymmetries in the resonators, and imperfect network analyzer calibration.

The phase asymmetries were different between the two resonators. Specifically, the R1 phase asymmetry could be removed by adding a lossless transmission line -0.25° long per gigahertz to port 2. The R2 resonator phase asymmetry was constant with frequency. A $+0.53^\circ$ phase shift was added to port 2, bringing the asymmetry down almost into measurement noise.

The fact that the phase asymmetries are fundamentally different between the two resonators suggests that the measured asymmetry is due to resonator asymmetry, as the same calibration was used for both resonators. We numerically checked the sensitivity of the technique to phase asymmetry by extracting the dielectric constants for two cases. The first case is with the measured data corrected for phase asymmetry and the second case is with the measured data uncorrected. *There is less than 0.0001 change in the extracted dielectric constants after removing correction for the measured phase asymmetry.*

To test magnitude asymmetry, a 0.1-Ω resistor is connected in series with port 2 of the measured data, roughly doubling asymmetry. The resulting extracted dielectric constants change by less than 0.0001. *We conclude that the asymmetries in the measured S-parameters are insignificant.*

The range of data used to determine the measured resonant frequencies by means of a quartic fit does affect the result. To evaluate the sensitivity, we decrease the range of data by 10% and note the consequent change in dielectric constant. For example, 450 data points might be used for a quartic fit. If we fit using 10% fewer points, 405 data points, then the resulting resonant frequency is different, leading to a different extracted dielectric constant. We find the largest change in dielectric constant is 0.0037 (horizontal, for the seventh resonance), but that most changes are under 0.001. *Thus, the data range used for fitting might affect the last digit of the results.*

Stripline is well known to be sensitive to the air gap between the two substrates. In the EM analysis, we assume an air gap equal to the measured metal thickness (0.7 mil) so the nominal air gap introduces no error. We used EM data ($\varepsilon_{hm} = \varepsilon_{vm} = 3.9$) in place of measured data and compared results for a 0.01778-mm (0.70 mil) and 0.018034-mm (0.71 mil) air gap leaving a 0.000254-mm (0.01 mil) air gap above the top of the resonator conductor. The dielectric constant extracted from the larger air gap is lower by about 0.008 (horizontal) and 0.001 (vertical) at all frequencies. *An air gap could be critical and the horizontal dielectric constant has about five times more sensitivity than the vertical.*

We also checked uncertainty in metal thickness by increasing the metal thickness to 0.018034 mm (0.71 mil) with no air gap above the metal. Change in the vertical dielectric constant was less than 0.0001. Change in the extracted horizontal dielectric constant was 0.0017. *Thus, metal thickness is also critical for horizontal dielectric constant.*

We note, in Table II, the resonator R2 extracted dielectric constants are about 0.05 (horizontal) and 0.01 (vertical) lower than resonator R1. This is consistent with an almost 0.1 mil wider air gap in resonator R2. However, it is difficult on a physical basis to justify an air gap of that size in the structure actually measured because the two substrates were compressed together with a uniform weight placed on top. This variability could be due to the well-known variability in the manufacture of FR-4. Position with respect to the strands of the fiber glass weave might even be important. A strand of glass fiber close to the gap in the R1 resonator can easily explain the increased horizontal dielectric constant.

To test for sensitivity to the resonator gapwidth, we again use EM calculated data for the measured data ($\varepsilon_{hm} = \varepsilon_{vm} = 3.9$). We then compare extracted dielectric constants for the nominal gap width of 0.762 mm (30 mil) to a gap width of

0.889 mm (35 mil). The new extracted horizontal dielectric constant is larger by 0.016 and the vertical is smaller by 0.004. Once more, the error is independent of frequency and horizontal results are more sensitive than vertical results. *Since the gap was photo-etched to within ±0.00254 mm (±0.1 mil), gap width should not be a significant error source.*

Substrate thickness is tested in a similar manner by comparing dielectric constants extracted from EM data calculated for a 1.524-mm-thick (60 mil) and then for a 1.481-mm-thick (58.3 mil) substrate (same thickness as the measured substrate). All dielectric constants at all frequencies changed by 0.002 to 0.003 with horizontal decreasing and vertical increasing for the 1.524-mm (60 mil) substrate. *The board was measured to about ±0.00254 mm (±0.1 mil) so substrate thickness is not expected to influence the results.*

It was stated above that this technique is valid provided that the field configuration is not significantly different between the two EM analysis cases (a and b), and the actual measured resonator (case m). This is equivalent to stating that the exactly correct matrix $\mathbf{A}$ in (6) and in (7) are identical. Of course, this is not true. To quantify the approximation, we change the EM calculated data, cases a and b to $\varepsilon_{ha} = \varepsilon_{va} = 4.2$ and $\varepsilon_{hb} = 4.3$ and $\varepsilon_{vb} = 4.1$. These are poor choices, as neither case a or b are close to the actual dielectric constants being measured, case m.

We find that for the above selection, the extracted dielectric constants change by up to 0.0017 (horizontal, decrease at low frequency, increase at high frequency) and by up to 0.0006 (vertical, most change at low frequency). We refer to this error as first iteration error. *For especially poor selection of cases a and b, first iteration error can be significant*. The error fades to insignificance when dielectric constants for case a or b (but not both, they must be linearly independent) are chosen equal to the results of the first iteration.

To investigate the importance of including connector models in the EM data, or equivalently removing the affect of the connectors from the measurement, we extract dielectric constants without accounting for the connectors at all. Thus, the EM data has no connector model and the measurement includes connectors. The maximum change in extracted dielectric constants is 0.01. *If not removed, connectors would insert a moderate error*. However, we also see that high accuracy for connector modeling or network analyzer calibration is not critical.

In comparing resonators R1 and R2 in Table II, we note some roughness in the data. For example, the R1 vertical dielectric constant is slightly higher than R2. However, for resonance number four, R2 is slightly higher. All significant error sources described above insert error that varies smoothly with frequency. We hypothesize that the roughness described here might be due to resonances in the narrow ring around the edges of the resonator boards (Fig. 1) that we used as a spacer to maintain a precise air gap. To test this hypothesis, the experiment could be repeated with the spacer ring broken up into small nonresonant patches, or with the ring firmly attached to the stripline grounds.

Maximum anisotropy is seen in Table II at lower frequencies with anisotropy nearly gone at high frequency. Note that dielectric "constants" must vary with frequency when loss is present [3]. A lossy substrate with a dielectric constant that is independent of frequency (a common EM analysis assumption) results in a noncausal (and thus, nonphysical) system. In our measurements, we see the horizontal dielectric constant increasing with frequency and the vertical dielectric constant decreasing with frequency. Perhaps the models of [3] can be modified to include this behavior.

TABLE III
UPPER LIMITS FOR EXTRACTED DIELECTRIC CONSTANT ERROR

Error Source	Error Magnitude	Importance
Measurement noise	< 0.001	Small
EM cell size error	< 0.001	Small
Measurement asymmetries	< 0.0001	None
Data range for fitting	< 0.004	Moderate
Air gap uncertainty	< 0.008/0.01 mil	Moderate
Metal thickness uncertainty	<0.0017/0.01 mil	Moderate
Coupled line gap uncertainty	< 0.0003/0.1 mil	None
Substrate thickness uncertainty	< 0.0002/0.1 mil	None
First iteration error	< 0.002	Removed
Connector model error	< 0.01	Removed

All error sources (Table III) are either inconsequential or reduced to insignificance, except for fitting error and air gap error. These two sources limit the result accuracy. Note also that if the EM analysis cell (mesh) size error is not carefully controlled, it can easily overwhelm the entire error budget.

VI. Future Efforts

This measurement technique has moderate sensitivity to the number of frequencies used for fitting a quartic polynomial to find resonant frequencies. This sensitivity can be reduced if a better fitting function (perhaps a Padé rational polynomial combined with the technique of [18]) is used to find resonances. The error due to the thickness of the air gap above the top side of the resonator metal can be eliminated by using microstrip. In addition, if a stripline resonator is used, and data higher than 2 GHz is desired, the two ground planes should be more solidly shorted together and the shorting walls should be placed as close to the resonators as is practical. This minimizes parallel plate modes and cavity resonator modes.

In our measurement, the horizontal dielectric constant is less than the vertical dielectric constant. This is the reverse of that reported by others (for other materials), e.g., [5], [7], and [8]. This could be due to the presence of more epoxy (and less glass fiber) on the board surface in the gap region of the resonators. Since coupled lines have most of their tangential field near the surface of the board (Fig. 4, middle), our measurement gives greater weight to the surface epoxy, thus measuring a lower horizontal dielectric constant. Others ([5], [7]–[9]) measure the entire volume of the substrate. For boards that have a thin epoxy-only layer at their surface, a three-layer model of the substrate might be useful. Alternatively, the anisotropic one-layer model automatically includes the fact that horizontal electric fields concentrate at the substrate surface for planar circuits. This suggests that bulk measurement of the horizontal dielectric constant should probably not be used for planar circuits on FR-4 and other inhomogeneous substrates.

In additional work to be reported [19], we have measured Rogers 3010 material, which is perfectly homogeneous and is also anisotropic. Measurements up to 10 GHz resolve nearly

100 even- and odd-mode resonances. The results correspond well to independent measurements. Note that many such anisotropic substrates exist (including all composite substrates). In fact, measurably anisotropic substrates might be more common than isotropic substrates. In present day practice, most substrates are approximated as isotropic because of lack of good dielectric measurements and because anisotropy was once difficult to include in planar EM analysis.

The presence of anisotropy can be inferred when an assumed isotropic dielectric constant measurement is found to depend on resonator line width and substrate thickness. This is because, for different line dimensions, the measured effective isotropic dielectric constant is formed from differently weighted averages of the true anisotropic dielectric constants.

Extension of this technique to measure magnetic anisotropy is straightforward, being exactly analogous to dielectric anisotropy measurement. It should be possible to extend this technique to extract loss information. The three quantities to be extracted are horizontal loss tangent, vertical loss tangent, and metal conductivity.

To extract three quantities, we need three equations, in analogy with (4) and (5). We will assume that a weighted sum of the three loss quantities is equal to the measured circuit loss (i.e., magnitude of the reflection coefficient). We assume a different weighting for each selection of a specific mode and frequency. For example, the reflection coefficient inside a resonance dip is more affected by conductor loss than outside a resonance dip.

First choose three mode/frequency combinations. For example, one could choose the reflection coefficient at: 1) the even-mode resonance; 2) the odd-mode resonance; and 3) 10% above (or below) the even (or odd mode)-mode resonance. These three quantities are analogous to the right-hand sides of (4) and (5), only now there are three equations instead of two. Finally, EM analyze three linearly independent loss cases with specific values of the loss tangents and conductivity. As long as frequencies are selected relative to the resonance frequencies, the exact values of the dielectric constants are not critical. The resulting three sets of equations are formed into a matrix equation [analogous to (6)] and solved for the analog to matrix $\mathbf{A}$ of (6). The inverse of this matrix is then used to extract the two loss tangents and one metal conductivity from a set of measured data, analogous to (7). Of course, as with all stripline/microstrip resonator measurements, measurement of low dielectric loss substrates is limited by conductor loss.

The dual symmetric RA resonator is evaluated for even and odd modes by using circuit theory to place PEC and PMC walls midway between the two coupled lines. Since the wall is placed there by circuit theory, the conductivity is perfect to the degree that the resonator structure is symmetric. This results in a lower loss and higher Q resonator than if a physical wall were actually placed there (not to mention that a PMC wall cannot be fabricated).

Carrying this idea further, we can invoke quad symmetry. Two identical RA resonators are placed one above the other (with no ground plane between them). With this configuration and the proper excitation of the four ports, an additional PEC or PMC wall is placed horizontally between the two dual RA resonators.

VII. Conclusion

A resonator formed from a length of a coupled line has two modes, even and odd. The velocity of propagation for each mode is dependent on the substrate dielectric constants. In the case of uniaxial anisotropy, the dielectric constant for the vertical (i.e., normal to the substrate surface) electric field is different from that of the horizontal (i.e., tangential to the substrate surface). The effective dielectric constant of each mode is assumed to be a weighted sum of the two anisotropic dielectric constants. Since the weighted sum is different for even and odd modes, we can extract the anisotropic dielectric constants by comparing measured resonant frequencies with two EM analysis results. The technique is demonstrated with measurements of FR-4, a common glass-fiber-weave/epoxy anisotropic substrate. Numerous error sources are evaluated quantitatively. We also describe extension of this technique to anisotropic loss and anisotropic magnetic parameters as well.

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