Extraction of Dispersive Material Parameters using Vector Network Analyzers and Genetic Algorithms

Jianmin Zhang

Marina Koledintseva
Missouri University of Science and Technology, marinak@mst.edu

David Pommerenke
Missouri University of Science and Technology, davidjp@mst.edu

James L. Drewniak
Missouri University of Science and Technology, drewniak@mst.edu

et. al. For a complete list of authors, see https://scholarsmine.mst.edu/ele_comeng_facwork/1620

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Jianmin Zhang, Marina Y. Koledintseva, David P. Pommerenke, James L. Drewniak
Department of Electrical and Computer Engineering, University of Missouri Rolla
Emerson Electrical Co. Hall, MO 65409, Rolla, USA
Email: jzp4f@umr.edu; marina@ieee.org; davidjp@umr.edu; drewniak@ece.umr.edu

Konstantin N. Rozanov
Institute for Theoretical and Applied Electromagnetics, Russian Academy of Sciences (Russia)
Email: k_rozanov@mail.ru

Giulio Antonini, Antonio Orlandi
Department of Electrical Engineering, University of L'Aquila, Poggio di Roio, I-67040, L'Aquila, Italy
Emails: antonini@ing.univaq.it and orlandi@ing.univaq.it

Abstract – *A novel method to extract dispersive properties for dielectrics over a wide frequency range is proposed. This method is based on measuring scattering parameters for planar transmission lines and applying genetic algorithms. The scattering parameters are converted into ABCD matrix parameters. The complex propagation constant of the TEM wave inside the line is obtained from A-parameters of the ABCD matrix. For planar transmission lines, analytical or empirical formulas for dielectric loss, conductor loss, and phase constant are known. The genetic algorithm is then used to extract the Debye parameters for the dielectric substrates. FDTD modeling is used to verify the dispersive parameter extraction by comparing with the measurement.*

Keywords – *Debye dispersion law, dielectric loss, conduction loss, S-parameter measurement, genetic algorithm, planar structure transmission lines.*

I. INTRODUCTION

The results of measurements of constitutive parameters of dispersive materials in many practical cases are not of independent interest: they are needed for further numerical analysis of electromagnetic structures where these materials are employed. To simulate the wideband electromagnetic response of complex structures, it is necessary to know the frequency dispersion law of the material parameters constituting these structures. Typically, the dispersion law can be obtained from experimental data of frequency dependence of permittivity (and/or permeability) by an appropriate fitting with an analytical function. Using an analytical law instead of "raw" experimental data allows for decreasing the computational resources. As the simplest law of frequency dispersion, the Debye dependence is used, while the more complex dispersion laws may be approximated by a summation of a number of the Debye or Lorentzian terms [1]. Herein, let us consider a summation of the Debye terms with a low-frequency part associated with effective conductivity,

$$\varepsilon_r(\omega) = \varepsilon_\infty + \sum_{i=1}^{N} \frac{\varepsilon_{si} - \varepsilon_\infty}{1 + j\omega\tau_i} - \frac{j\sigma_e}{\omega\varepsilon_0}, \qquad (1)$$

where the ε_{si} and τ_i are the static dielectric constant and relaxation constant for the ith Debye component respectively, and the ε_∞ is the high-frequency ("optic") relative permittivity, and σ_e is the effective conductivity.

In many practical cases, the type of the dispersion law is known, but the parameters of this law are unknown. The problem can be formulated as an experimental determination of the parameters of dispersive curves for permittivity without getting the detailed interim information on the values of material parameters over the frequency range of measurements. To solve this problem, the known dispersion law can be fitted directly during the measurement process. This allows for simplifying the measurement procedure and increasing the accuracy, when the measuring technique itself provides a comparatively high level of error or low sensitivity. This approach is widely used for measuring dielectric constant in optics, where measurement of the phases of transmitted and reflected ways are connected with substantial experimental difficulties; then, only wave amplitudes are measured, while phases are restored using the Kramers-Kronig (KK) relations. At microwave frequencies, the Q-factors of absorption curves are significantly lower than those in the optical region, and the edges of absorption (or resonance) lines are usually beyond the frequency limits of available measurements. The alternative method for extrapolation of the edges of absorption lines is based on the KK relations in the limited frequency range [2]. Then, to restore the behavior of a material in case of insufficient sensitivity of measurements or incomplete data, an appropriate dispersion law can be applied, as done in [3].

The extraction of the Debye parameters from a set of measured or reference data typically requires the solution of systems of non-linear equations [4], which can be

cumbersome. In addition, exact data for the real and/or imaginary parts of permittivity of materials are not available in many cases. It is beneficial to develop a simple, accurate, and reliable method for dispersive media parameter reconstruction for full-wave numerical modeling, and this paper is aimed at this objective.

II. PROPOSED APPROACH AND APPLICATION OF A GENETIC ALGORITHM

Herein, an effective technique for measuring parameters of dispersion laws of sheet materials and layered-material substrates is proposed. This method is based on a combination of (1) S-parameter measurements; (2) analytical modeling of loss and phase constant in planar transmission lines; (3) comparison of the measured and modeled total loss and phase constant according to some accepted criteria in the frequency range of interest; and (4) the correction of the dispersion curve parameters until the criteria are satisfied. The correction is fulfilled using a powerful, robust, and efficient genetic algorithm (GA) for global searching and optimization [5]. This is a straightforward method for a single-component Debye material. It can be also used for multi-components Debye or more complex frequency dependences of permittivity and permeability as well.

First, the S-parameters of the planar transmission line structures, such as parallel-plate, stripline, and microstrip test fixtures containing the dispersive dielectric medium are measured. The measured S-parameters are converted into the ABCD matrix parameters. The A-matrix parameters of the ABCD matrix can be found using the expressions given in [6]

$$A = \frac{(1+S_{11})(1-S_{22}) + S_{12}S_{21}}{2S_{21}}, \quad (2)$$

Neglecting the port effects, the propagation constant γ can be found as

$$\gamma = \frac{\cosh^{-1}(A)}{l}, \quad (3)$$

where l is the length of the transmission line with the single TEM mode propagating. Furthermore, the propagation constant γ can be expressed in terms of the attenuation constant α and the phase constant β through

$$\gamma = \alpha + j\beta = (\alpha_c + \alpha_d) + j\beta, \quad (4)$$

where the α_c is the attenuation constant due to the finite conductivity of the conductor constructing the transmission line, and α_d is the attenuation constant associated with the dielectric loss.

The total attenuation constant $(\alpha_c + \alpha_d)$ and the phase constant β obtained from the measured S-parameters are used as an optimization goal in a genetic algorithm optimization. The total attenuation and phase constants are calculated at each frequency point using interim Debye parameters during every GA iteration. The objective function for optimization is

$$\Delta = \frac{1}{N}\sqrt{\sum_{i=1}^{N}\left\{\left[\Delta_c\right]^2 + \left[\Delta_d\right]^2 + \left[\Delta_\beta\right]^2\right\}}, \quad (5)$$

where

$$\Delta_c = \frac{\left|\alpha_c^m - \alpha_c^e\right|}{\max\left|\alpha_c^m\right|}, \quad (6)$$

$$\Delta_d = \frac{\left|\alpha_d^m - \alpha_d^e\right|}{\max\left|\alpha_d^m\right|}, \quad (7)$$

and

$$\Delta_\beta = \frac{\left|\beta^m - \beta^e\right|}{\max\left|\beta^m\right|}. \quad (8)$$

The superscript m denotes the measured data, and the superscript e indicates the data from evaluation using analytical or empirical formulas. The values $\Delta_{c,d,\beta}$ are normalized to the maximum measured values of α_c, α_d, and β, respectively. N is the population number in the GA optimization. Herein, the fitness index is chosen as

$$p = \Delta^{\frac{-1}{\pi}}. \quad (9)$$

The fitness index is assigned to the individual (herein, this is the Debye parameter set including ε_s, ε_∞, τ, and σ_e) for that GA iteration. The fitness index is used to distinguish how good each individual is at competing in its environment. An individual with a higher fitness index indicates a greater chance to remain in the search pool, greater chance to produce offspring in the next generation, and more close to the final solution. Using the fitness index, the "selection" operation is applied, and new offspring with higher fitness index is produced. To maintain diversities in the GA search pool, "mutation" operation is implemented on each parameter set. Then, the "recombination" procedure generates new individuals for the next generation iteration. The global optimal solutions are achieved as soon as the chosen criteria are satisfied, or the pre-defined maximum number M of iteration steps for all N populations is reached.

III. FORMULATION

Analytical or empirical formulas of total attenuation and propagation constants for planar transmission lines including parallel-plate, stripline, and microstrip line are detailed in this section. Limitations on parameter extraction for each transmission line structure are discussed and given in this section as well.

A. Parallel-plate transmission line

The parallel-plate transmission line structure is the simplest structure, when only the TEM mode is taken into account. Formulas derived for the calculation of total loss and propagation constant are based on the assumption that the higher modes and fringing fields associated with the structure are ignored. This is true for a parallel-plate structure only over a limited frequency range, depending on its dimensions. Therefore, it is necessary to design a set of parallel-plate structures for studying the frequency behavior of various dielectrics in different frequency ranges. The assumptions given herein really mean two things. First, the PMC (perfect magnetic boundary condition) is applicable. Second, the cutoff frequency associated with the first order mode limits the thickness, d, of the dielectric medium between the two plates, referring to the cross-section of the structure shown in Figure 1.
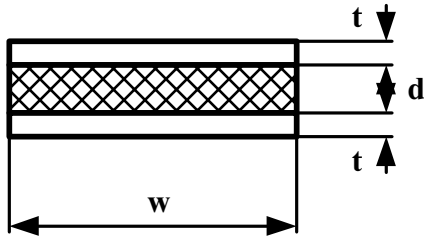

Fig. 1. Cross-section of a parallel-plate structure.

Using the PMC boundary condition, the width, w, of the transmission line is limited by the TEM assumption as well. Since the width is much larger than the thickness, the highest frequency before first higher order mode propagating along the transmission line is limited by the plate width w. In order to make the fringing fields really small, the ratio between the width of the structure to the thickness of the dielectric, w/d, must be large enough.

Based on the assumption above, the propagation constant for wave travelling in a parallel-plate transmission line is defined as

$$\beta = \omega\sqrt{\mu\varepsilon} = \omega\sqrt{\mu_r \varepsilon_r' \mu_0 \varepsilon_0}, \tag{10}$$

where ω is the angular frequency; μ_0 and ε_0 are the permeability and permittivity of free space respectively; μ_r is the relative permeability of the substrate material, and it is considered as 1 in all the studied cases through the paper. The ε_r' is the real part of ε_r (1). The value ε_r' is interim during the GA parameter extraction. The attenuation constant due to the conductors in a parallel-plate transmission line with the only TEM mode can be calculated as [6]

$$\alpha_c = \frac{R_s}{\eta d}, \tag{11}$$

where $\eta = 120\pi\sqrt{\dfrac{\mu_r}{\varepsilon_r'}}$ is the TEM wave impedance, and R_s is the conductor surface resistance. The dielectric loss is found from (1) as

$$\alpha_d = \frac{\beta \tan\delta}{2} = \frac{\beta}{2}\left(\frac{\varepsilon_r''}{\varepsilon_r'}\right). \tag{12}$$

The total loss of a parallel-plate transmission line is the summation of the conductor loss (11) and the dielectric loss (12).

B. Microstrip transmission line

The calculation of the phase and attenuation constants for a microstrip line shown in Figure 2 is analogous to that for the parallel-plate geometry, provided only the quasi-TEM wave propagating along the structure. The only difference is that the relative dielectric constant used in the formula is an effective one instead that obtained directly from the Debye model. Thus, the phase constant for microstrip line is

$$\beta = \omega\sqrt{\mu_0 \varepsilon_0 \varepsilon_e'}, \tag{13}$$

where the effective dielectric constant is calculated as [6]

$$\varepsilon_e' = \frac{\varepsilon_r' + 1}{2} + \frac{\varepsilon_r' - 1}{2\sqrt{1 + \dfrac{12h}{w}}}. \tag{14}$$

In (14), ε_r' is the real part of the relative permittivity of the substrate dielectric material under study, and h and w are the cross-section dimensions of the microstrip line structure shown in Figure 2.
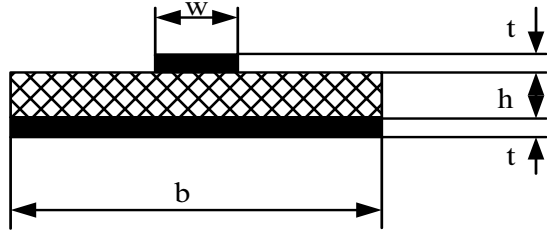
Fig. 2. Cross-section of a microstrip structure.

For the structure shown in Figure 2, the attenuation constant due to conductivity is given by

$$\alpha_c = \frac{R_s}{w Z_0}, \tag{15}$$

and the dielectric attenuation constant is calculated as

$$\alpha_d = \frac{\omega\sqrt{\mu_0 \varepsilon_0} \cdot \varepsilon_r' \left(\varepsilon_e' - 1\right)}{2\sqrt{\varepsilon_e'} \cdot \left(\varepsilon_r' - 1\right)} \tan\delta$$

$$\tag{16}$$

$$= \frac{\omega\sqrt{\mu_0 \varepsilon_0} \cdot \varepsilon_r' \left(\varepsilon_e' - 1\right)}{2\sqrt{\varepsilon_e'} \cdot \left(\varepsilon_r' - 1\right)} \left(\frac{\varepsilon_r''}{\varepsilon_r'}\right)$$

The characteristic imedance of the microstrip line Z_0 can be found from [6]

$$Z_0 = \frac{60}{\sqrt{\varepsilon_e'}} \ln\left(\frac{8h}{w} + \frac{w}{4h}\right) \qquad \text{with } \frac{w}{h} \le 1, \tag{17}$$

or, when $\frac{w}{h} \ge 1$

$$Z_0 = \frac{120\pi}{\sqrt{\varepsilon_e'} \cdot \left[\frac{w}{h} + 1.393 + 0.667\ln\left(\frac{w}{h} + 1.444\right)\right]} \tag{18}$$

The total loss $(\alpha_c + \alpha_d)$ and the wave propagation constant then can be calcualted. It is worth mentioning that the formulas given above for the loss and propagation constant calculations for micrstrip line are based on the quasi-TEM assumption. The higher-order modes, surface wave propagating in the metal-dielectric-air structure, and radiation effects in the open structure are not taken into account. This limits the upper frequency of the dispersive material parameter extraction.

C. Stripline

The calculations of propagation constant and dielectric loss for a stripline with the cross-section given in Figure 3 can be implemented by using (10) and (12), respectively.
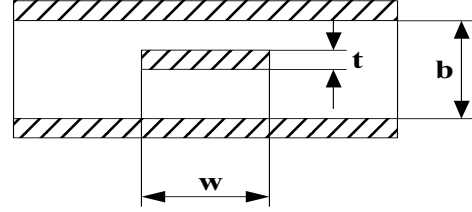

Fig. 3. Cross-section of a stripline structure.

The conduction loss can be estimated by the perturbation method or Wheeler's incremental inductance rule as [6]

$$\alpha_c = \frac{2.7 \cdot 10^{-3} R_s \varepsilon_r' Z_0}{30\pi \left(b - t\right)} \xi \qquad \text{for } Z_0\sqrt{\varepsilon_r'} < 120, \tag{19}$$

or

$$\alpha_c = \frac{0.16 R_s}{Z_0 b} \zeta \qquad \text{for } Z_0\sqrt{\varepsilon_r'} \ge 120. \tag{20}$$

The coefficients A and B in (19) and (20) are

$$\xi = 1 + \frac{2w}{b - t} + \frac{b + t}{\pi \left(b - t\right)} \ln\left(\frac{2b - t}{t}\right), \tag{21}$$

and

$$\zeta = 1 + \frac{b}{0.5w + 0.7t}\left(0.5 + \frac{0.414t}{w} + \frac{1}{2\pi}\ln\left(\frac{4\pi w}{t}\right)\right). \tag{22}$$

The characteristic impedance of the stripline can be found in [7, p. 130].

The formulas (19) and (20) are valid only for a single TEM mode propagating in a stripline, when fringing fields are neglected, and there is no edge coupling. In reality, the higher-order modes can be easily suppressed by limiting the reference spacing to the quarter wavelength ($\lambda / 4$). These assumptions are always true, for example, in multilayer printed circuit boards (PCBs), since typically $t \ll b$, and $b \ll \lambda / 4$.

IV. MEASUREMENTS AND CASE STUDIES

Three test structures were fabricated. Two structures, parallel-plate and microstrip, were made from the same

sample of the material, which was a large double-sided copper clad FR-4 sheet.

Parallel-plate and microstrip lines. The dimensions of the parallel-plate structure were 71.36 mm x 19.80 mm x 1.25 mm, the dielectric spacing of the FR-4 was 1.05 mm, and the thickness of the copper sheet was 0.1 mm. Two SMA connectors were symmetrically mounted at the ends of the structure. The distance between the centers of the SMA connectors was 63.4 mm.

The dimensions of the microstrip line were 69.00 mm x 19.80 mm x 1.25 mm, and the distance between the centers of two SMA connectors was 61 mm.

A vector network analyzer (VNA) *HP 8753 D* was used to measure S-parameters for both test structures. The frequency range of measurements was from 100 MHz to 5 GHz. Prior to measurements, the SOLT ("Short-Open-Load-Thru") calibration was implemented. Port extension procedure, done after calibration, allowed for removing the influence of the electrical lengths of the SMA connectors. The measured S-parameters were converted into the ABCD matrix parameters, and the total attenuation and propagation constants were found. The Debye parameters were extracted for both the parallel-plate transmission line and the microstrip line (see Table I) applying the GA procedures.

Table I. Extracted Debye parameters for parallel-plate and microstrip transmission lines.

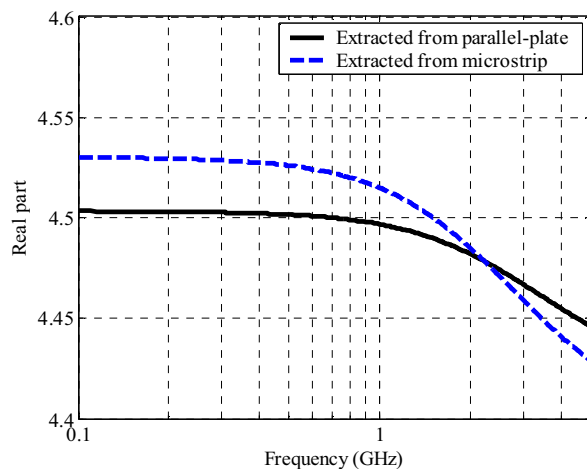| Structure	ε_s	ε_∞	τ (ps)	σ_e (mS/m)
Parallel-plate	4.504	4.420	46.37	2.531
Microstrip	4.530	4.398	57.22	2.351

Fig. 4. Real part of the extracted permittivity for different structures with identical material.

The real and imaginary parts of the permittivity, as well as its phase, for the two extracted Debye curves are shown in Figures 4, 5, and 6, respectively. The maximum difference between the corresponding pairs of curves is 0.025 for the real

part, 0.02 for the imaginary part (with effective conductivity term), and 0.172º for the phase of permittivity. The above comparison results demonstrate that the proposed method works well for the extraction of dispersive material parameters for substrate materials.
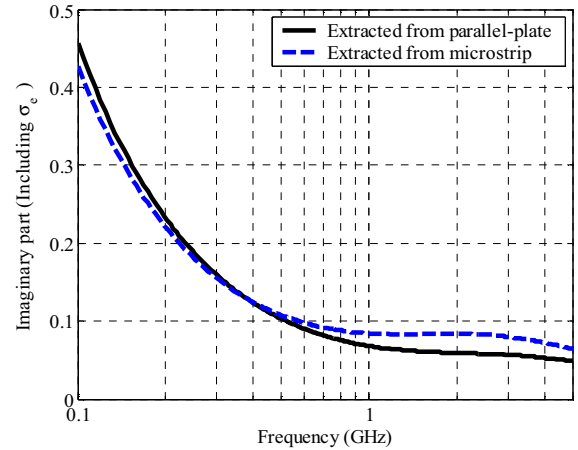

Fig. 5. Imaginary part (including effective conductivity) of the extracted permittivity for different structures with identical material.
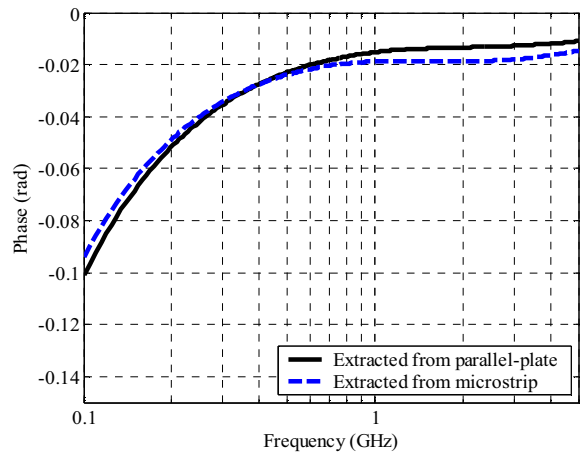

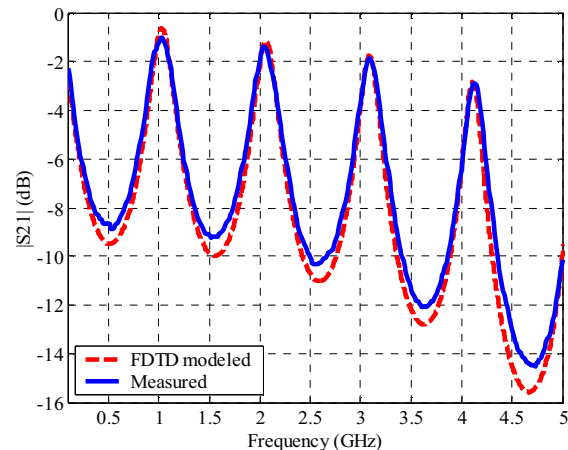Fig. 5. Phase of the extracted permittivity for different structures with identical material.

Fig. 6. |S₂₁| comparison between measurement and full wave modeling for the parallel-plate transmission line.
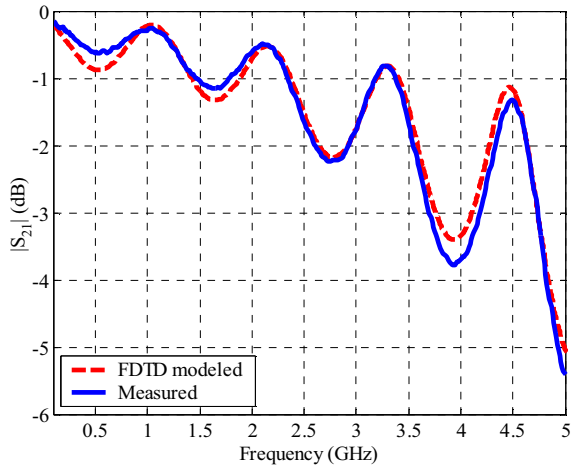
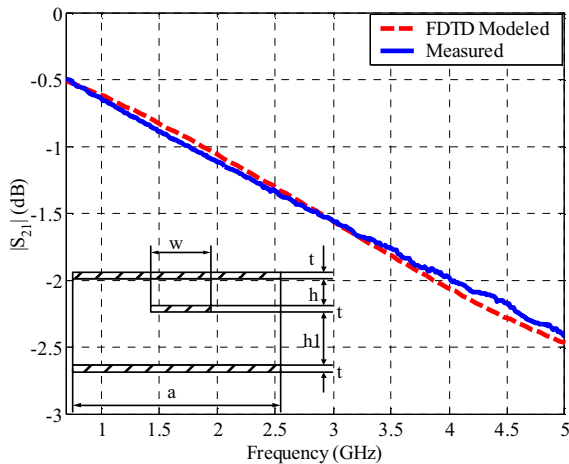Fig. 7. |S$_{21}$| comparison between measurement and full wave modeling for the microstrip transmission line.

Fig. 8. |S$_{21}$| comparison between measurement and full wave modeling for the stripline.

Stripline. The stripline structure was built on a PCB with the substrate of NELCO 4000-13SI dielectric. The total length of the stripline was 203.20 mm. Other dimensions were h_1=0.51 mm; h=0.25 mm; w=0.36 mm; a=4.09 mm; and t=0.033 mm, referring to Figure 8. The VNA *HP 8720 ES* with the ATN-4112A S-parameter test set was used for measurements in the frequency range from 800 MHz to 5 GHz. The TRL calibration method was implemented before the measurement, and the influence of ports (SMA connectors) on the measured data was eliminated. The Debye parameters of the substrate in this structure, extracted using the GA, are represented in Table II.

Table II. Extracted Debye parameters for the stripline structure

Structure	ε_s	ε_∞	τ (ps)	σ_e (mS/m)
Stripline	3.46	3.41	29.3	1.08

It should be mentioned that despite the port extension procedure eliminated phase effects of the ports on the results of measurements, it could not compensate losses associated with the ports (Table I). In contrast to the parallel-plate and microstrip cases, the extracted Debye parameters in the stripline (Table II), due to the employed TRL calibration technique, are not affected either by loss or by phase shifts associated with ports.

The extracted Debye parameters for the stripline were verified using the full-wave FDTD modeling. Comparison between the results of measurements and simulations is shown in Figure 8.

V. CONCLUSIONS

A method for the extraction of the Debye parameters for the dispersive dielectric media is presented. It is based on the S-parameter measurements in planar transmission line structures and application of the genetic algorithm. The parallel-plate, microstrip, and stripline test structures used in this method are simple for implementation both in measurements and analytical modeling, provided that the only TEM wave propagates in any of these structures. For verification purposes, the extracted Debye parameters of the dispersive dielectric (FR-4) substrates in these transmission lines are used in the full-wave FDTD simulations. The good agreement achieved between the measured and the FDTD-modeled scattering parameters justifies the validity, high accuracy, and simplicity of the proposed method.

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