EVALUATION OF A STRIPLINE STRUCTURE THROUGH THE METHOD FDTD FOR ELECTROMAGNETIC CHARACTERIZATION OF FINE FILMS IN THE RANGE OF RF AND MICROWAVE

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***Abstract**. A comprehensive approach to the design of a stripline for EMC testing is given in this paper. Signal conductors between two ground planes (i.e., stripline transmission lines) serve as the means for propagation high-speed digital and analog signals on printed wiring boards (PWB's), and multichip modules (MCM's). For many important applications (e.g., wireless and other high speed technologies), the need for minituarization requires that these high-speed electrical interconnections be provided on relatively dense PWB's and MCM's. This leads to design uncertainties concerning characteristic impedance and S-parameters. The stripline device possesses a strong electromagnetic interaction with the dielectric in all his space distribution. In this sense it is a fort candidate in the characterization of films. The difficulty in characterizing fine films is the low electromagnetic sensibility of the sample port and a mathematical model need that relates the variation of the S-parameters of the structure with the variation of the permeability and complex permissiveness for materials with losses. The author's attention has been focused on the design items that are most crucial by achievement of satisfactory value of S-parameters and the impedance matching at the feeding ports in the extended frequency range from 320 MHz to 10 GHz with the relative electric permissiveness varying of the values 1, 3, 5, 7, 9. A rigorous analysis of the resonant characteristics for a broad band waveguide by using the finite difference time domain method is presented. The use of FDTD method for the description of the electromagnetic behavior of the cell discontinuities in the analysis of the S- parameters, input and output impedance and potency at the feeding ports, permits in the 320 MHz – 10 GHz frequency band the achievement of a good precision for the results on materials with low electromagnetic characteristics* $(\varepsilon_r \leq 10)$.

***Keywords**: stripline, FDTD software, electromagnetic waves.*

1. INTRODUCTION

Numerical methods, such as finite difference (FD), boundary element (BE) or finite element (FE), are efficiently used to computed near scattering fields: by comparison with other analytical methods, advantages are that scattering by any geometrical shape may be computed, and the physical nature of the scatterer may be complex (perfect electric conductor coated with an inhomogeneous dielectric for example). The modeling of stripline structures in multi-layer dielectric media has been thoroughly investigated in the past and several methods have already been investigated. The research has been based on various methods such as Finite Elements (FE), Finite Difference Time Domain (FDTD) and Method of Moments (MoM) (Itoh 1989 and Harrington 1986). The later is the most widely employed for the modeling of planar geometries, mainly due to the fact that it does not require the meshing of the whole volume of the structure.

Development of simple and robust numerical methods for wideband extraction of frequency characteristics of planar sheet materials for various electromagnetic applications is an important present day problem. In particular, characterization of dielectric substrates for printed circuit boards (PCBs) is vital to achieve the first-pass success in modern high-speed digital system designs. To build the full-wave model for a given signal path, the detailed structures are known, but the well-represented dielectric material properties of the corresponding substrates are unknown. In general, the dielectric properties (relative permittivity and loss tangent) used in the full-wave model come from PCB vendor with only one or two frequency points. However, dielectric representations with either one or two frequency points for a PCB substrate are not sufficient for accurate full-wave simulations, since complex permittivity of a PCB substrate may vary substantially over the wide frequency range. Besides, dielectric representation with only one or two points may result in causality issues in full-wave modeling, which causes the divergence problem in time domain simulations.

The dielectric constants at microwave range is important not only in scientific but also in industrial applications like in microwave heating, in the study of biological effects and in high-speed and microwave circuits where impedance, attenuation and phase velocity depend on the complex dielectric constant.

The stripline is the most convenient planar analog of the coaxial transmission line and only at the very edges of the stripline an enhanced current density is encountered. The current distributions can aid in determining the amount of coupling between adjacent traces fabricated between two common ground planes, which is important in signal-integrity analysis (Johnson 1993, Catt et al. 1979, Bogatin 2004, Hall et al. 2000 and Paul 1992).

The finite-difference time-domain (FDTD) method (Yee 1966) has been proven to be an effective means that provides accurate predictions of field behaviors for varieties of electromagnetic interaction problems.

2. THEORY

The complex relative permittivity and permeability are given bellow:

$$\varepsilon_r = \varepsilon_r^{'} - j\varepsilon_r^{''} \tag{1}$$

$$\mu_r = \mu_r^{'} - j\mu_r^{''} \tag{2}$$

The stripline has a characteristic impedance of Z_0 (free space region) which when loaded with the sample material becomes Z, where:

$$Z = Z_0\sqrt{\frac{\mu_r}{\varepsilon_r}} \tag{3}$$

In the unloaded regions, the propagation constant is $k_0 = \omega\sqrt{\mu_0\varepsilon_0}$, while in the loaded region, the propagation constant is generally complex and is designated by k, where:

$$k = k_0\sqrt{\mu_r\varepsilon_r} \tag{4}$$

At the plane boundaries there are complex reflection coefficients R and $-R$, respectively, where:

$$R = \frac{Z - Z_0}{Z + Z_0} \tag{5}$$

Solving equations (3), (4) and (5), simultaneously, it may yield the desired parameters.

A simple method to measure S-parameters using FDTD was developed and presented. The new method is based in exciting the stripline circuit with a voltage source that has an internal source resistance integrated into FDTD code.

The incident (a_i) and reflected (b_i) normalized voltages at each port of an N-port network can be expressed in terms of the total voltage (V_i) and total current (I_i) recorded at port i as:

$$a_i = \frac{V_i + R_{0i}I_i}{2\sqrt{R_{0i}}} \tag{6}$$

$$b_i = \frac{V_i - R_{0i}I_i}{2\sqrt{R_{0i}}} \tag{7}$$

where R_{0i} is the reference resistance for which S-parameters are determined. The S-parameters can be computed as:

$$S_{ij} = \left.\frac{b_i}{a_j}\right|_{a_k=0,k\neq j} \tag{8}$$

3. NUMERICAL SOLUTION

FDTD is used to solve Maxwell’s equations for arbitrary model spaces. Indeed, FDTD allows us to solve models that would be difficult or impossible with analytical methods. FDTD is a direct time-domain solution to Maxwell’s curl equations (Jackson 1999). The continuous-time expressions of Maxwell's equations for linear, isotropic, non-dispersive materials which will be discretized in XFDTD are:

$$\frac{\partial \overline{E}}{\partial t} = \frac{1}{\varepsilon}\nabla \times \overline{H} - \frac{1}{\varepsilon}\left(\overline{J_{source}} + \sigma \overline{E}\right) \tag{9}$$

$$\frac{\partial \overline{H}}{\partial t} = -\frac{1}{\mu}\nabla \times \overline{E} - \frac{1}{\mu}\left(\overline{M_{source}} + \sigma^{*} \overline{H}\right) \tag{10}$$

Where μ represents the magnetic permeability and σ represents the magnetic conductivity.

The two curl equations (9 and 10) can be discretized to obtain a total field FDTD technique. Alternately the fields can be expressed as:

$$E = E^{total} \equiv E^{incident} + E^{scattered} \tag{11}$$

$$H = H^{total} \equiv H^{incident} + H^{scattered} \tag{12}$$

The scattered fields, emanating from a scattering or interaction object, can be more readily absorbed than a total field by an outer radiation boundary condition applied at the problem space extremities or faces. This is especially important in situation where FDTD simulation and center conductor was excited for a source of radio frequency of 1 Volt operating in scale from 320 MHz to 10 GHz. The objective of this work is analyzing the electromagnetic sensibility, input in which the scattered fields are desired and are of much lower amplitude than the total fields.

The scattered wave arises on and within the interaction object in response to the incident field so as to satisfy the appropriate boundary conditions on or within the interaction object. These boundary conditions are the Maxwell equations themselves, which in the limit of a perfect conductor require $E^{scattered} = -E^{incident}$ in the scatterer. For anything other than a perfect conductor the scattered fields depend on the constitutive parameters of the material. The scattered fields are subject to the Maxwell equations for this media when in the media, while outside the media they satisfy the free space Maxwell equations. The incident field always propagates in free space (even when passing through the interaction object or scatterer material) and is defined as the field that would be present in the absence of the scatterer.

The characteristic impedance of the stripline test section should be smoothly matched with the feed and terminations points in order to minimize the standing waves (Paul 1992). Thereby, the most critical parameters that directly determine the physical design of the stripline are impedance matching at the feed port (S11 parameter) and transmission between two ports (S21 parameter). All simulations presented in this paper are made using FDTD.

An analysis can be performed for the stripline configuration as depicted above. The width of the strip is denoted by w, the length of the stripline by l and the distance between the metal ground planes by h. For simulations was chosen the aspect ratio $w/h = 1.3$.

The critical dimensions for the stripline sample holder are as follows: the housing is perfect electrical metal with inside dimensions of $(12.00 \times 10.00 \times 0.64)$ mm, the width of the conductor perfect center is 2.6 mm. For each simulation the relative permittivity varied from $(1.0, 3.0, 5.0, 7.0, 9.0)$. The sample holder showed in Figure 1 was meshed with cells, each of which are $(0.1 \times 0.1 \times 0.1)$ mm in size for the impedance, determining the largest work area.

Measuring the permittivity of lossy materials, circuit boards, thin films, and substrates nondestructively is frequently of interest (Baker-Jarvis and Kabos 2004). These materials may be thin and may be clad in copper, making measurements with closed transmission lines difficult. The figure 1 shows the cross section of a covered stripline, the line loss for a stripline consists of components from both the strip and its ground planes.

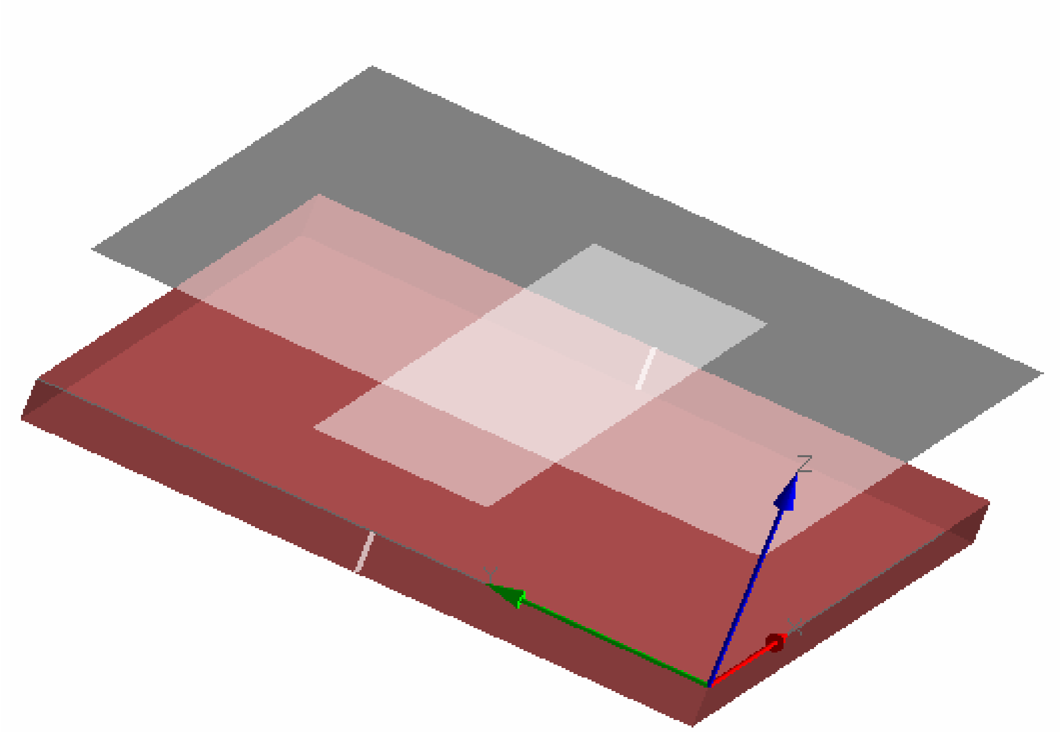

Figure 1. Holder sample simulation

4. RESULTS

To accomplish a simulation three criteria are important. First, the sample port was designed for a characteristic of impedance of 50 Ohms for the sample port emptiness. The second criterion is to accomplish one analysis meticulous in the modules of the electromagnetic field to determine as much the maximum intensity of the electromagnetic field as the positioning where will have the largest uniformity, the result of those it analysis will be the determination of the positioning of the film and the guarantee of an increase of electromagnetic sensibility. We have been studying the influence of the driver's width in function of the thickness of the dielectric in function of that sensibility Lutif et al. (2010). The third criterion is to simulate a fine film and to analyze it, holding of the electromagnetic sensibility (Kim 2000). We considered the work area in the simulation as being the largest transmission parameter, above 99% of the transmitted wave and parameter of smaller reflection or same to -30dB in function of the width of the frequency band.

Relative complex permittivity of the dielectric material affects the bandwidth (necessary for high speed signal transmission) and circuit density. High speed signal transmission and low signal attenuation are desired in many applications. The relationship of relative complex permittivity to signal transmission speed is shown below:

$$V = c/\sqrt{\varepsilon_r} \tag{13}$$

This equation shows that low permittivity and low loss are required to achieve high operating frequencies. For this reason, PC boards and substrate material have lower permittivity (typically 2 to 10) compared with other dielectric materials.

Figure 2 presents the behavior of the magnitude S11 as a function of applied frequency and Fig. (3) shows the behavior of magnitude S21 as a function of applied frequency. The figure 4 describes the behavior of the effective complex impedance as a function of frequency, with a no magnetic sample and changes in the relative permissiveness.

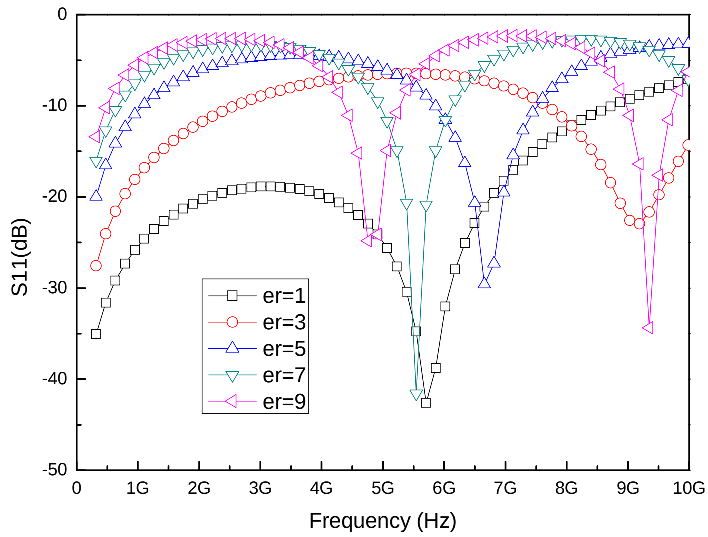

Figure 2. Magnitude S11 as a function of frequency.

We can see in the Figure 2 that the work area changes a lot with different magnetic materials. The electromagnetic sensibility is related with low reflection smaller than -20 dB and transmission near than 0dB. $\varepsilon_r = 1$ showed the largest work area of 320 MHz until 6.5 GHz and better electromagnetic sensibility and with the ratio of the conductor width with thickness of the dielectric has $w/h = 1.3$, according Fig. 1 and Fig. 2. Figure 3 shows the analysis of the input impedance as a function of frequency. On the other hand $\varepsilon_r = 7$ has the work area between 5.3 GHz and 5.5 GHz, and $\varepsilon_r = 9$ has three picks of transmission from 4.9 GHz until 5.2 GHz, 5.5 GHz until 7.2 GHz and 9 GHz until 9.5 GHz.

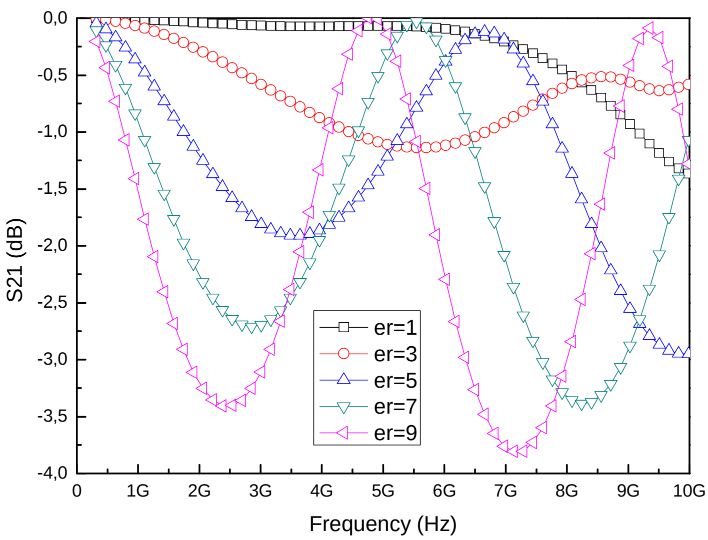

Figure 3. Magnitude S21 as a function of frequency.

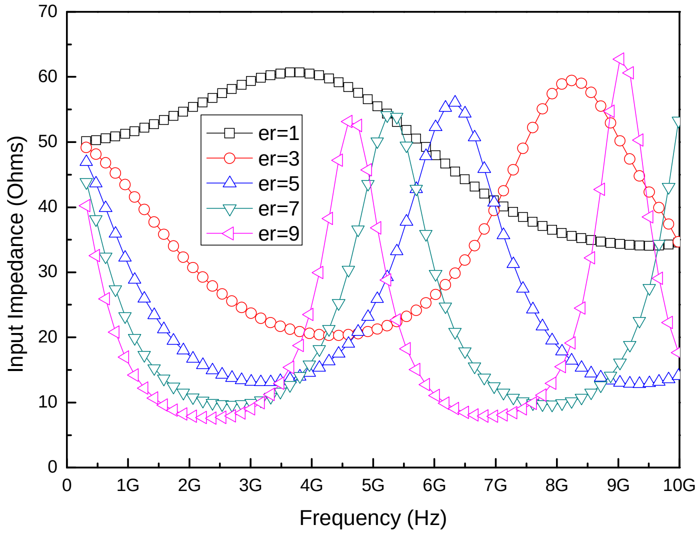

Figure 4. Input Impedance as a function of frequency.

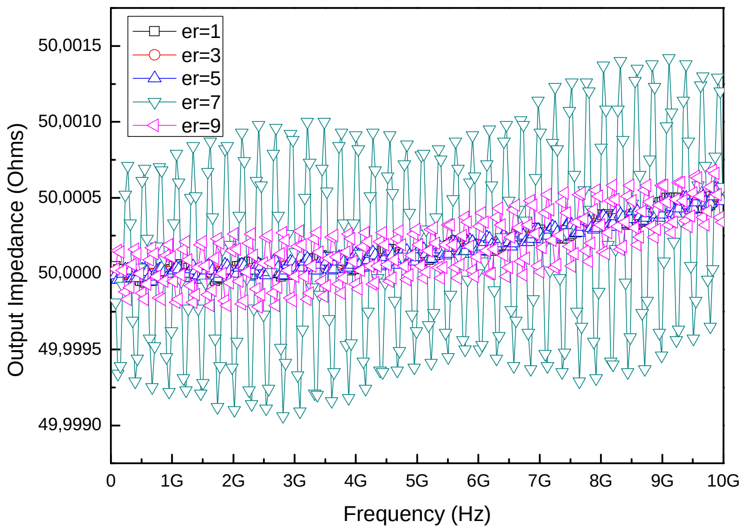

Figure 5. Output Impedance as a function of frequency.

In the Figure 5, it presents the behavior of the impedance and executes the complex of the port 2 in function of the frequency. The sample port with those dimensions assures a perfect matching with 50 Ohms of the frequency of 300 MHz up to 4 GHz. That frequency it is the great area to do the electromagnetic characterization of fine films.

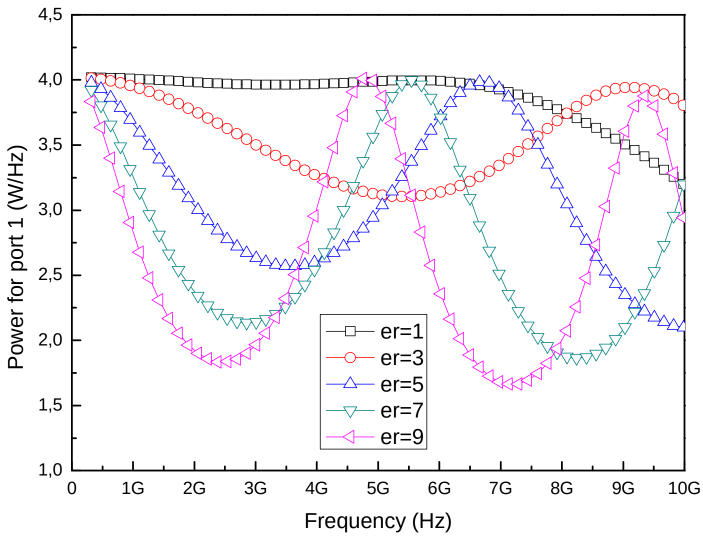

Figure 6. Power for port 1 in function of the frequency.

The Figures 6 and 7 describes power for ports 1 and 2 as a function of frequency, in these figures we can see the energy dissipation due to dielectric loss for each permittivity.

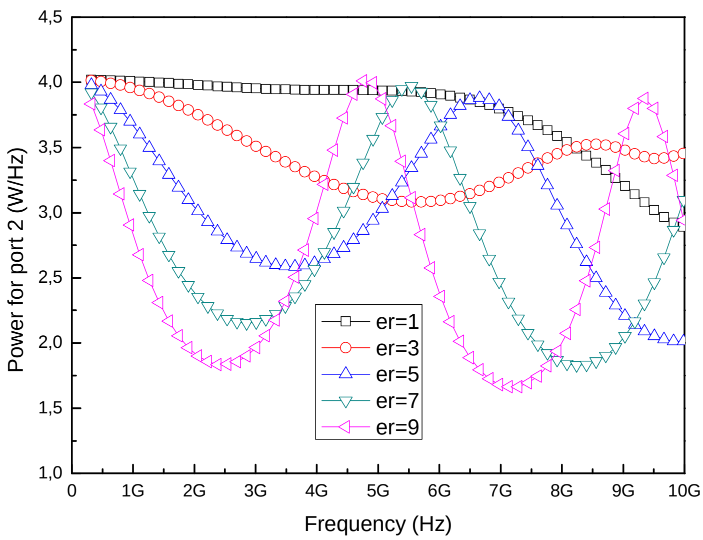

Figure 7. Power for port 2 in function of the frequency.

5. TABLES

The values of S-parameters are in decibels (dB) as a function of frequency (GHz). The conversions of decibels in percentages are showed in the table 1 according with the Eq. (11).

$$A(\%) = \left(1 - 10^{-dB/10}\right) * 100 \tag{11}$$

Table 1. Values of S parameters in dB converted in percentages (Lee 1991).

S-parameters (dB)	S-parameters (%)
0	0
-3	50
-10	90
-20	99
-30	99,9
-40	99,99

6. CONCLUSIONS

That computational analysis went decisive to describe some effects involved in the electromagnetic characterization of fine films. The positioning of the film is extremely relevant for the sensibility and it is related with the length of the sample.

The sample holder with $\varepsilon_r = 1$ and ratio $w/h = 1.3$ it showed to be a good option to do characterization of nanoscale films due to excellent electromagnetic sensibility in the scale from 320 MHz to 6.5 GHz. The simulations also show sensibility with changes in the magnetic properties of the conductor.

7. ACKNOWLEDGEMENT

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