Examples of the EZ-FDTD modeling and measurements.

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Some examples of finite-difference time-domain simulation using EZ-FDTD tool developed at the UMR are represented along with the results of the corresponding measurements, demonstrating that the agreement between the measurements and the simulations is satisfactory within the required accuracy.

1. Debye dielectric case.

The model geometry is shown in Figure 1. This is a structure with two ports on a double-sided copper clad FR-4 dielectric plate having the dimensions 200x150x1.65 mm$^3$. Two 50-Ω SMA connectors are mounted on the plate at (33; 128) and (133; 73) [in mm]. The FR-4 dielectric material was modeled as a Debye dispersive dielectric medium with the parameters $\varepsilon_{\infty}=4.301$; $\varepsilon_s = 4.122$; $\tau = 2.91E-11$ s; $\sigma_s = 2.257E-3$ S/m. Copper planes were modeled as PEC plates. The excitation source is a voltage source with an impedance of 50 Ω. Termination is a lumped 50-Ω resistor. The measured and simulated S-parameters in the frequency range from 100 MHz to 5 GHz are presented in Figures 2 and 3, respectively.
Figure 1. Model geometry of the Debye dielectric case.

Figure 2. Frequency dependences of $|S_{11}|$ for the Debye dielectric case.
2. Lorentzian dielectric case.

The test fixture consists of two copper plates and one dielectric spacer with dimensions of 60x10x0.6 mm³. The EZ-FDTD model geometry is shown in Figure 4. Two copper plates were modeled as PEC, and the spacer was modeled as the Lorentzian dielectric with the parameters: \( \varepsilon_\infty = 2.75; \varepsilon_\infty = 2.21; \Delta f = 9.39 \cdot 10^9 \text{Hz}; f_0 = 1.08 \cdot 10^9 \text{Hz}; \sigma_e = 5.6 \cdot 10^{-3} \text{ S/m.} \) The model was excited by a voltage source with a source impedance of 50 Ohms, and was terminated with a lumped 50-Ohms resistor. Comparison of the S-parameter between measurement and simulation is represented in Figures 5 and 6.
3. Lorentzian magnetic – Debye dielectric case.

The FDTD PLRC algorithm was applied to model a microstrip fixture comprised of two copper layers with an isotropic magnetic material (Ni-Zn ferrite) spacer. An orthographic view of the structure used in the FDTD modeling is shown in Figure 7. The basic computational domain in FDTD
method consists of a rectangular mesh bounded by perfectly matched layers (PML) to simulate an infinite space. The computational domain was discretized by a uniform mesh of $0.316 \times 0.324 \times 0.316$ mm$^3$ along x, y, and z-axes, respectively, and the total number of cells was approximately 250,000. The strip and the ground plane were modeled as perfect electric conductors (PEC) 20 and 80 cells wide, respectively. The magnetic material layer between two adjacent planes was modeled with 10 cells thickness using PLRC algorithm. Two PEC plates were added vertically to model the copper tape at each side of the test fixture in experiment. A sinusoidal modulated Gaussian 50-Ohm voltage source was applied horizontally at Port 1 through a wire with 1 cell width, and Port 2 was terminated with a 50 Ohms load through a wire with 1 cell width. The wire structures were modeled using a thin wire algorithm.

Figure 7. The configuration of the microstrip structure in FDTD modeling.

To model an electrodynamic structure containing this ferrite, the frequency dependences of its permittivity and permeability must be known. A stripline resonator technique developed at the U.S. National Institute of Standardization Technology (NIST) to measure the permeability of the specimen was applied [1]. The NIST stripline cavity is schematically shown in Figure 1. It consists of a centered conductor mounted equidistantly between two ground planes and terminated by two end plates. The end plates are permanently attached to the ground planes, thus the resonator is of fixed length and non-tunable. The length of the resonator is 1 m, giving a fundamental resonance of approximately 150 MHz with harmonic resonances spaced at 150-MHz increments above the fundamental. The resonator upper frequency limit is defined by the frequency at which the first higher order TE$_{10}$ begins to propagate in the stripline cavity. For this cavity, the ground plane separation is 76.2 mm, giving a theoretical upper frequency
limit of 1970 MHz. The Q-factor of the unloaded resonator has been several thousand after removing copper oxidation from its walls and applying silver paint to the joints to improve the conductivity of the structure. A propagating TEM mode is excited within the stripline structure by the coupling loops mounted on one of the end plates. The first, or fundamental, resonance is achieved when the resonator length corresponds to a half-wavelength. There are higher-order resonances of the cavity at the harmonics of the fundamental frequency.

![Diagram](image)

**Figure 8. NIST stripline cavity for permittivity and permeability measurements.**

Measurements of the complex magnetic permeability ($\mu^* = \mu' - \mu''$) are conducted by placing the specimen under test at an axial E-field node, located at the structure’s end plates, where the electric field intensity is zero and does not influence on the measurement of permeability.

From Waldron’s small perturbation theory [2], for a thin rectangular magnetic slab with $2y$ in the $x$-direction, $s$ in the $y$-direction, and $l_1$ in the $z$-direction, the real and imaginary parts of permeability are found as

$$
\mu'_r = 1 + \frac{\Delta f}{f_0} \frac{b(b-t)l_0}{Bysl_1}; \quad \mu''_r = \left(1 - \frac{1}{Q_0} \right) \frac{b(b-t)l_0}{2Bysl_1} \tag{1}
$$

In (1), $2b$ is the distance between the cavity ground planes; $t$ is the thickness of center conductor; $l_0$ is the length of the resonator; and $B$ is a geometry factor determined by the dimensions of the resonator [1,2]. The real and imaginary parts of permeability are measured by observing, respectively, the shift of the cavity resonance frequency $\Delta f$ to the lower frequencies, and the decrease of the cavity $Q$-factor at the cavity loading with the sample under investigation.
Measured complex permeability is shown in Figure 2. The results were checked to satisfy the causality principle, applying the Kramers-Kroenig relations [3].

Ni-Zn ferrite is a medium that has both dispersive permeability and permittivity in the frequency range of interest, up to 2 GHz. Hence, the FDTD modeling should include not only updating equations for magnetic field with recursive convolution term taking into account the dispersive law for magnetic susceptibility, but also for electric field taking into account the law for electric susceptibility.

The real part of Ni-Zn ferrite permittivity was measured over the frequency range from 150 MHz to 2 GHz using the NIST cavity method [1]. The rectangular specimen under test was placed in the region of maximum and uniaxial E-field, specifically, in the midpoint of the center strip, and with the smallest dimension oriented along the x-axis to minimize the E-field nonuniformity across the specimen. The H-field in the point where the ferrite specimen is placed is zero. Thus, the slab is located at the midpoint of the center strip in experiment. This measurement demonstrates that the permittivity of this sample can be approximated by the Debye dispersive curve. The imaginary part of the permittivity was reconstructed based on the Debye parameters extracted from the measurements of the real part of the permittivity.

The measured and fitted permeability and permittivity of the Ni-Zn ferrite slab are shown in Figures 9 and 10. The parameters of a single-component \( (v_1 = 1.0) \) Ni-Zn ferrite as the Lorentzian magnetic material are \( A = 8.167497 \times 10^6 \text{ rad/s}; \delta = 1.088 \times 10^{10} \text{ rad/s}; \omega_0 = 1.109 \times 10^{10} \text{ rad/sec} \). Parameters of the corresponding Debye curve approximating Ni-Zn ferrite permittivity are \( \varepsilon_\infty = 20.0; \quad \varepsilon_\infty = 7.8; \quad \tau = 6.233 \times 10^{-10} \text{ s}; \quad \sigma = 10^{-4} \text{ S/m} \). The frequency dependences for \( |S_{11}|, |S_{21}| \), and phases of \( S_{11} \) and \( S_{21} \) were modeled using the PLRC algorithm, as well as the standard Yee’s algorithm for the FDTD technique [4,5]. The results are shown in Figure 11.

The standard Yee’s algorithm incorporates the frequency-dependent loss of the media by a constant real value of effective conductivity and magnetic loss sigma in a number of individual frequency points was also used here. Three uniformly distributed frequency “points” were chosen within the frequency region from 100 MHz to 2GHz. The normal “non-dispersive” FDTD simulation was run independently 10 times for every chosen frequency “point”, and the corresponding S-parameters were obtained as a result of the simulations.
Figure 9. Real and imaginary parts of Ni-Zn ferrite permeability.

Figure 10. Real and imaginary parts of Ni-Zn ferrite permittivity.

Figure 11. Measured and modeled S-parameters of the microstrip fixture with Ni-Zn substrate: (a) amplitude of $S_{11}$; (b) phase of $S_{11}$; (c) amplitude of $S_{21}$; and (d) phase of $S_{21}$. 
4. Double-sided substrate with wideband Lorentzian and Debye dielectrics.

Computations using a Debye and a wideband Lorentzian model were run for the copper-clad double-sided printed circuit board containing an FR-4 dielectric layer, as shown in Figure 12.

![Figure 12. Schematic of the test board with dispersive dielectric in between: (a) top view and (b) profile at Port 1. All dimensions are in mm.](image)

The board size was 150 mm × 200 mm, and the dielectric layer was 1.65 mm thick. Port 1 (30 mm, 125 mm) was located close to the board corner to excite as many modes of the parallel-plane waveguide as possible. Port 2 was selected arbitrarily and had coordinates (130 mm, 70 mm). Both test ports in the experimental board were built using semi-rigid coaxial cables having outer shields soldered to the ground plane with a 360-degree connection. The center conductors extended through the thickness of the board and were soldered to the opposite plane. SMA connectors were mounted on the other ends of the semi-rigid cables for the connection of test instruments. The computational domain used in the FDTD simulation was discretized by a uniform mesh with the steps $\Delta x = \Delta z = 1$ mm, and $\Delta y = 0.165$ mm. A sinusoidally modulated 50-Ω Gaussian voltage source applied vertically above the ground plane was used in the simulation. The power and ground planes were modeled as perfect electric conductors (PECs) of zero thickness. Eight perfectly matched layers (PML) were placed at each boundary plane of the computational domain, and seven free space layers were placed between the PML and the test board. The FDTD codes used
double-precision numbers. Experimental data was obtained using a 37275A Wiltron network analyzer.

The extracted from the manufacturer’s data parameters of the wideband Lorentzian dielectric were $\varepsilon_s=4.301$, $\varepsilon_\infty=4.096$, $\sigma_e=2.295\cdot10^{-3}$ S/m, $f_0=39.5$ GHz, and $\Delta f=200$ GHz, where $\Delta f=\delta/\pi$ is half of the resonance line width, and $f_0=\omega_0/(2\pi)$ is the resonance frequency. A Debye material with a frequency dependence of permittivity close to the above wideband Lorentzian material has parameters of $\varepsilon_s=4.301$, $\varepsilon_\infty=4.096$, $\sigma_e=2.295\cdot10^{-3}$ S/m, and $\tau_r=2.320\cdot10^{-11}$ s. The experimental and the modeled scattering-matrix parameter $|S_{21}|$ of the described board with the wideband Lorentzian dielectric, the Debye dielectric, and corresponding experimental curves are shown in Figure 13. For both the Debye and the wideband Lorentzian models, the FDTD-simulated and experimental data are very close (within 1-3 dB of amplitude, and the shift of resonance peaks is less than 5 % in most of the frequency range). The similarity of the FDTD results with the Debye and wideband Lorentzian model is because the FR-4 is a low-loss dielectric used at comparatively low frequencies where the material resonance effects are not very pronounced. In this example, the ratio $\delta/\omega_0=2.53$, or $Q=0.4$, which is considerably less than 1. However, at higher frequencies the wideband Lorentzian algorithm may be preferable even for FR-4, because of a greater number of parameters involved and more flexibility when approximating the parameters of the dielectric substrate.

![Figure 13. Measured and FDTD modeled $|S_{21}|$ for the test substrate as a wideband Lorentzian dielectric and as a Debye dielectric.](image)
An example of a material, for which the Debye model is insufficient, is presented in Figure 14. This is a composite dielectric with the polymer matrix of $\varepsilon_r \approx 2.15$ filled with an aluminum powder. The concentration of aluminum powder is 10%, and the diameter of particles is 10-15 $\mu$m. A 0.8 mm thick sheet of this material was placed between two copper plates with dimensions of 100 mm x 75 mm. Ports 1 and 2 for the $|S_{21}|$ parameter measurement and modeling had the coordinates in millimeters of (15, 63) and (65, 35), respectively. The composite is modeled as a wideband Lorentzian dielectric with the parameters $\varepsilon_r = 2.5, \varepsilon_\infty = 2.15, \Delta f = 190$ GHz, $f_0 = 85$ GHz, and $\sigma_\varepsilon = 10^{-4}$ S/m. In this case, the ratio $\delta / \omega_0 = 1.12$, or $Q = 0.9$, which is close to 1. The mesh for the FDTD modeling was chosen as 0.5 mm x 0.16 mm x 0.5 mm. The frequency range where measurements and modeling were made was from 40 MHz to 14 GHz.

![Figure 14. Measured and FDTD modeled $|S_{21}|$ frequency dependence for a composite material as the wideband Lorentzian dielectric and as the Debye dielectric.](image)

The modeled curve with the wideband Lorentzian dielectric fits the experimental results well up to 6 GHz (within 3 dB of the amplitude, and the frequency shift of resonance peaks is less than 10 %). At higher frequencies the discrepancy increases, but the model remains satisfactory (within 5 dB of the amplitude, and less than 15 % shift of resonance peaks). The Debye model with the parameters $\varepsilon_r = 2.5, \varepsilon_\infty = 2.15, \tau_r = 2.5 \cdot 10^{-12}$ s, and $\sigma_\varepsilon = 10^{-4}$ S/m that approximates the behavior of the same composite material, does not fit the experimental $|S_{21}|$ curve at higher frequencies. The amplitude
discrepancy is approximately 20 dB, and the shift of resonance peaks is up to 50 % in the frequency range of 6-15 GHz. This can be explained by an insufficient loss in the Debye model at higher frequencies. Therefore, the Debye model in this case is not suitable.

5. Comparison of wideband and narrowband Lorentzian dielectric behavior

The narrowband Lorentzian (NBL) model can be useful for full-wave analysis of structures containing high-loss composite media, for example, polymer composites with conducting fibers that find application as antistatic materials, electromagnetic shields, radar absorbers, etc. Such materials have pronounced dispersion at microwaves, and their effective permittivity behaves according to a single-pole or a multi-pole NBL law depending on the contents of the composite. The same material as shown in Figure 7 (a, b) of Lagarkov et al. [6], was used in the present study for NBL FDTD modeling. This is a two-phase composite, filled with aluminum-coated glass fibers (the fiber length is 8 mm) with concentration of 0.02 %, and 10% of aluminum powder. The aluminum powder contains particles with sizes in the range of 10-15 µm. The aluminum-coated glass fibers have an overall thickness of approximately 25 µm, where the aluminum coating is approximately one-third of the glass fiber cross-section. The resistivity of the coating is assumed to be equal to that of bulk aluminum, 2.65 µΩ⋅cm. The fibers are embedded in a polymer matrix of Teflon type having a dielectric constant \( \varepsilon_r = 1.8 \). The dielectric loss in the matrix is negligible. The thickness of the composite material is of on the order of sub-millimeter to millimeters. In the FDTD modeling, any effects of possible anisotropy are not taken into account.

The experimental characteristics of this NBL material are shown in Figures 15 (a, b). The fitted curve used in the FDTD models was determined by the least mean square method [7,8]. The characteristics of a fictitious wideband Lorentzian (WBL) material are also shown in Figures 15 (a, b). The WBL material has the same parameters as the NBL, except the width of the resonance line. For the WBL material the ratio \( \delta / \omega_0 = 1.03 \), so that \( Q = 0.96 \), which is close to 1, and for the NBL material the ratio \( \delta / \omega_0 = 0.16 \), and \( Q = 6.1 \). These materials – NB with parameters as in [6], and fictitious WBL – were located between two metal perfect electric conducting planes for comparison in FDTD modeling. The size of the simulated board was 100 mm x 75 mm, and ports 1 and 2 had coordinates in millimeters of (15, 63)
and (65, 35), respectively. The frequency dependences of $|S_{21}|$ for the NBL and WBL materials over the band of 100 MHz -15 GHz are compared in Figure 16. The FDTD model of a test board with a NBL material demonstrates significantly greater number of resonances than the WBL, as expected given the nature of the materials. Multiple resonances due to the geometry of the two-sided copper-clad substrate are present only at lower frequencies where they are not damped significantly by the loss in the materials. At higher frequencies the NBL material exhibits higher loss than the WBL material, as follows from the difference in amplitude of $\varepsilon''$ for NBL and WBL materials shown in Figure 15 (b). The frequency characteristic of the NBL material behaves as that of a band-reject filter, and the filtering effect of the WBL material is less pronounced.

Because of the material resonance effects, the narrowband algorithm requires at least 60,000 time steps, while the wideband requires approximately 30,000. This is because of the longer high-Q-resonance “ringing” with the narrowband Lorentzian model. Both narrowband and wideband Lorentzian recursive convolution FDTD algorithms could be useful for other dispersive and lossy dielectric, magnetic, and magneto-dielectric materials. An algorithm for the narrowband Lorentzian model application to magnetic (ferro-, ferri-, and antiferromagnetic) materials that exhibit a resonance behavior at microwave frequencies due to spin magnetic moment precession, for example, a hexagonal ferrite composite material, has been reported [9].

![Figure 15. Frequency dependence of (a) the real part of the permittivity, and (b) the imaginary part of permittivity for the Lorentzian dielectrics. WBL material: $\varepsilon_s=0.1, \varepsilon_s=6.8, f_0=8.6$ GHz, $\Delta f =17.8$ GHz. NBL: $\varepsilon_s=10.1, \varepsilon_s=6.8, f_0=8.6$ GHz, $\Delta f =2.8$ GHz. Conductivity is $\sigma_e = 0$ in both cases.](image)
Figure 16. Modeled frequency dependence of $|S_{21}|$ for the wideband and the narrowband Lorentzian dielectrics.

6. Conclusions.

The represented herein and many other EZ-FDTD modeled cases demonstrate that the agreement between the measurements and the simulations is satisfactory within the required accuracy. The EZ-FDTD is a robust and powerful full-wave electromagnetic simulation tool, which successfully has been used for numerous EMC/EMI and other electromagnetic problems containing complex geometries and materials.

References


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