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Design of small-size planar filters using FDTD and wavelet analysis

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Abstract – Compact planar resonators and filters were designed by using a Finite-Difference Time-Domain method. The accuracy of the FDTD method was increased by extending the Bérenger's perfectly matched layer to non-homogeneous media. By applying the wavelet analysis and signal estimation techniques, the total simulation time was reduced by 5 times. The _e..._e_ ..e.. _e.e.ope_ on a substrat_ with 10.8 dielectric constant and 0.635 mm height. Each resonator occupied down to 32% of the surface of a folded half-wavelength resonator. Two-pole and four-pole cross-coupled filters were developed for 900 MHz frequency band applications.

Keywords: FDTD, microwave devices, planar resonators, microstrip filters

I. INTRODUCTION

Compact cost-effective devices are strongly required by the modern wireless communications systems. When devices with distributed parameters are designed at rather low frequencies of mobile communications such as 900 MHz, the size reduction is a major requirement. Transmission line models are not sufficiently accurate when analyzing structures with discontinuities such as bends, junctions, open-ends, etc. In this paper, an accurate improved FDTD method is proposed for compact filters.

II. DESIGN METHOD

The 3-D Finite-Difference Time-Domain (FDTD) method [1] was employed due to its accuracy and versatility. The method selects from all four Maxwell equations the two curl equations and solves them numerically by finite differences in time-domain. The grid points where the electric and magnetic fields are calculated alternate in space forming the FDTD cell as shown in Fig. 1. Every electric field $\bar{E}$ component surrounded by four circulating magnetic field $\bar{H}$ components, and every $\bar{H}$ component is surrounded by four circulating $\bar{E}$ components.

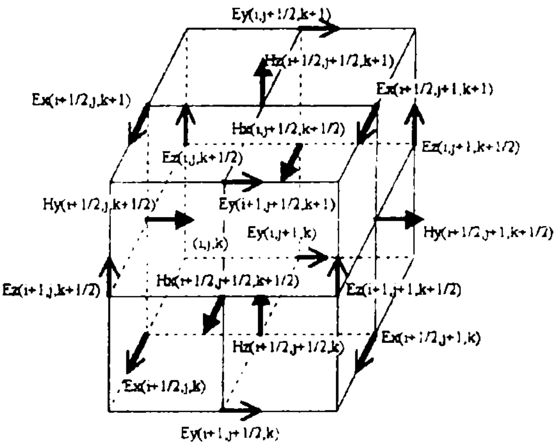

Fig. 1. The FDTD cell

Let us consider the curl equation

$$\frac{\partial E_x}{\partial t}=\frac{1}{\varepsilon}\left(\frac{\partial H_z}{\partial y}-\frac{\partial H_y}{\partial z}-\sigma E_x\right) \qquad (1)$$

A field component such as $E_x^{n+1/2}(i,j,k)$ at time step $n+1/2$ is not calculated directly in FDTD scheme, but it may be approximated by the arithmetic average between the components calculated at time steps n and $n+1$. The final reult is an explicit expression of the component at time step $n+1$ when the components at previous time steps are already known.

$$E_x^{n+1}(i,j,k)=\frac{1-\frac{\sigma(i,j,k)\Delta t}{2\varepsilon(i,j,k)}}{1+\frac{\sigma(i,j,k)\Delta t}{2\varepsilon(i,j,k)}}E_x^{n}(i,j,k)+\frac{\frac{\Delta t}{\varepsilon(i,j,k)}}{1+\frac{\sigma(i,j,k)\Delta t}{2\varepsilon(i,j,k)}}\left[\frac{H_z^{n+1/2}\left(i,j+\frac{1}{2},k\right)-H_z^{n+1/2}\left(i,j-\frac{1}{2},k\right)}{\Delta y}-\frac{H_y^{n+1/2}\left(i,j,k+\frac{1}{2}\right)-H_y^{n+1/2}\left(i,j,k-\frac{1}{2}\right)}{\Delta z}\right] \qquad (2)$$

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The devices were designed on a 0.635 mm height Rogers substrate with 10.8 dielectric constant. The FDTD grid was $\Delta x = \Delta y = 0.15$ mm and $\Delta z = 0.127$ mm. The time step was chosen of $\Delta t = 0.27$ ps, in order to satisfy the stability criterion Courant-Friedrichs-Levy. The field excitation was chosen as a Gaussian pulse

$$E_z(t) = E_{z0}(t)e^{-\frac{(t-T_0)^2}{T^2}} \quad (3)$$

where $T = 38\ \Delta t$ and $T_0 = 4\ T$.

III. THE NON-HOMOGENEOUS PERFECTLY MATCHED LAYER

The accuracy of the FDTD method depends very much on the quality of the absorbing boundary conditions. The perfectly matched layer (PML) idea is to use a lossy material to match the incident waves. However, for an isotropic lossy material, the match occurs only at normally incidence, therefore such a material has only a limited application for absorbing boundary condition [2]. A PML should match waves of arbitrary incidence, polarization, and frequency. The Bérenger's innovation consits in a derivation of a *split-field* formulation of Maxwell's equations; namely, each vector field component is split into two orthogonal components. The PML technique decomposes each field projections in two, and the wave incident to PML is attenuated via electric (σ) and magnetic (σ^*) conductivity. The extremely small reflection is satisfied by the impedance matching condition perpendicularly to the PML layer.

In order to analyze the new microstrip devices, a Non-Homogeneous Perfectly Matched Layer (NH-PML) was developed. For such a non-homogeneous boundary, the values of electric (σ) and magnetic (σ^*) conductivities should satisfy the impedance matching condition.

$$\sqrt{\left(\mu - j\frac{\sigma^*}{2\pi f}\right)\Big/\left(\varepsilon - j\frac{\sigma}{2\pi f}\right)} = const. \quad (4)$$

where μ is the magnetic permeability, ε is the electric permittivity and f is the frequency. The conductivities increase with the depth into the PML. The efficiency of the NH-PML increases with its thickness. For a given thickness best profile for the conductivity is the geometric series profile such as

$$\sigma_l^* = \frac{\sigma_0 \varepsilon_r (g-1)}{\eta^2 \ln g} g^l \quad (5)$$

where $l = 0, 1, 2\ldots$ represents the grid point index inside the NH-PML, σ_0 is the electrical conductivity at the PML interface and g is the geometric progression ratio. The absorption profile of the non-homogeneous PML was established after empirical investigations. Satisfactory results were found for $\sigma_0 = 1$ mS, g = 2.3 for a layer thickness of 13 cell.

IV. FDTD SIGNAL PROCESSING

The long computation time is a major drawback of the FDTD method. The FDTD signal was processed in order to reduce the number of iterations. The wavelet analysis is based on the possibility of developing a signal $f(t)$ in wavelet packets as in relation

$$f(t) = \sum_k c_{j0}(k)\varphi_{j0,k}(t) + \sum_k \sum_{j=j0}^{\infty} d_j(k)\psi_{j,k}(t) \quad (6)$$

where the functions $\varphi_{j0,k}(t) = 2^{j_0/2}\varphi(2^{j_0}t - k)$ and $\psi_{j,k}(t) = 2^{j/2}2^{j_0/2}\psi(2^j t - k)$, $\psi(t)$ is the mother wavelet and $\varphi(t)$ is the scaling function.

The wavelet analysis rejects the noise caused by the finite precision; it also clears the effects of the FDTD signals of too high frequency, which cannot be accurately computed. An accurate FDTD simulation of the propagation of signals of too high frequency, greater than ~30 GHz in our case, is unnecessary and it would involve a very fine mesh and extensive computer resources.

The time step required by the stability criterion is too small and a conventional Fast Fourier Transform would provide a low-resolution signal in frequency domain. Therefore, the FDTD signal is not only de-noised but also de-sampled. The first n_{min} points of the FDTD signal, are used as a training set to the signal estimation technique. In Fig. 2, $n_{min} = 100$ and the estimated signal $y_4(t)$ is a close approximation of the FDTD signal $y_3(t)$. The total FDTD number of iteration time was reduced to a fifth of the initial number.

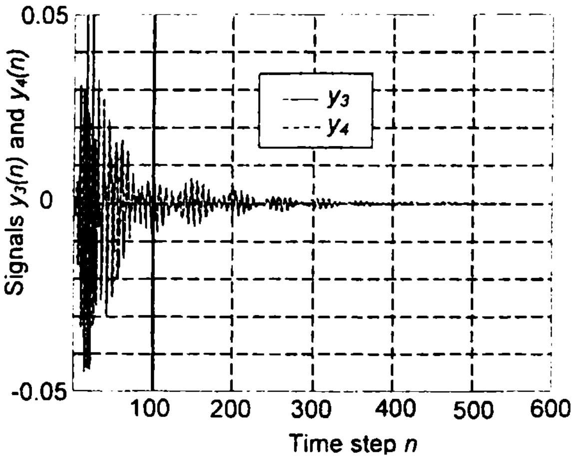

Fig. 2 The comparison between the the FDTD signal $y_3(t)$ and the estimated signal $y_4(t)$

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V. FILTER DESIGN

Resonators and filters were developed on the substrate with the characteristics mentioned above. In the first approximation, the novel compact resonators were described by using the transmission line theory as modified stepped-impedance resonators (SIR) [5]. However, due to the discontinuities effects, the final geometry for a given resonance frequency was established by FDTD analysis.

For an accurate filter design, the variation of the coupling coefficients with the relative position of the .es...t..s ..as ....es.iga..d by ..si..g ....he square s'ape o' t'e propose' resonators a''ows a variety of electric, magnetic and mixed couplings between resonators. As an example, the dependence of the magnetic coupling coefficient with the coupling gap *s* is e is presented in Fig. 3.

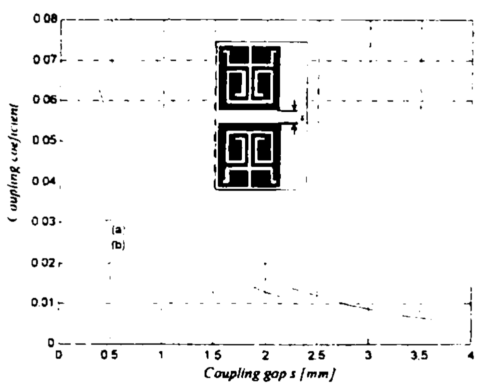

Fig. 3 The dependence of the magnetic coupling coefficient versus the coupling gap *s*

The proposed designs allow direct couplings with the external circuit. The variation of the external quality factor Q_{ext} with the coupling line position was also analyzed by employing the FDTD method.

Several two-pole filters were developed by employing different couplings between resonators. For the two-pole filters in Fig. 4 the coupling between resonators is the mixed. Each resonator in the insert of Fig. 3 has only a 9.52 mm size, thus the proposed resonator upies only 32% of 'he sur'ace area of an open loop resonator designed for the same frequency on the same substrate.

Despite the fact that filter shown in the insert (b) of Fig. 3 exhibits two transmission nulls at each side of the pass-band, it is very hard to control the positions of these nulls. For an improved filter response, it was already shown that a full control of these nulls can be provided by filters with cross-coupled resonators [3]. The extra negative couplings result in a sharper filter roll-off for an increased filter selectivity. The newly developed square planar resonator can be effectively employed for cross-coupled filter design [4, 5].

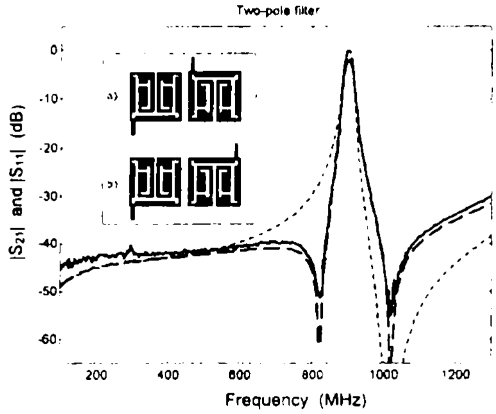

Fig. 4. Two-pole filter response using compact resonators. The solid line and the dashed line represent the measured and simulated response, respectively, for the structure shown in the a) insert. The dotted line represents the simulated response of the filter shown in the insert b).

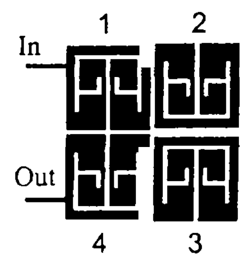

Fig. 5. Four-pole cross-coupled resonator using compact resonators of different type.

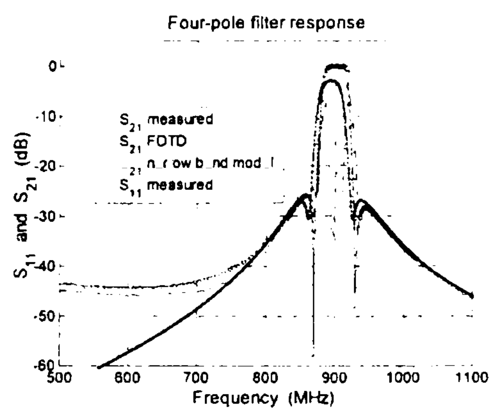

Fig. 6. Measured versus simulated responses of the filters in Fig. 5.

An example of a cross-coupled filter developed with new compact planar resonators is shown in Fig. 5. Each resonator is 10.64 mm in size. This filter is characterized by a coupling matrix

$$\mathbf{M} = \begin{bmatrix} 0 & 0.0360 & 0 & -0.0081 \\ 0.0360 & 0 & 0.0259 & 0 \\ 0 & 0.0259 & 0 & 0.0360 \\ -0.0081 & 0 & 0.0360 & 0 \end{bmatrix}, \quad (7)$$

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and an external quality factor Q_{ext}= 17.86. As it is shown in Fig. 6, the measured response of the four-pole cross-coupled filter follows closely the simulated response.

VI. CONCLUSIONS

A 3-D FDTD method was developed in order to accurately design compact resonators and filters. A non-homogeneous perfectly matched layer (NH-PML) with a geometric series profile for the electric and magnetic absorption was developed as absorbing boundary conditions. The FDTD signal processing by using wavelet packets and signal estimation techniques resulted in a reduction by up to five times of the computation time.

The FDTD method was successfully applied to small-size filter design. The proposed resonators occupy down to 32% of the surface area of a folded half-wavelength resonator designed on the same substrate for the same frequency.

Two-pole and four-pole cross-coupled filters were developed for 900 MHz wireless systems such as GSM and GPRS for the 900 MHz. However, the same design technique can be applied for devices of the 3G communications standards. The devices are cost-effective; they do not require via-holes and any additional lumped elements.

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