Efficient Analysis of Microwave Passive Structures Using 3-D Envelope-Finite Element (EVFE)

Hsiao-Ping Tsai, *Student Member, IEEE*, Yuanxun Wang, *Member, IEEE*, and Tatsuo Itoh, *Fellow, IEEE*

***Abstract*—A three-dimensional envelope-finite element (EVFE) technique is proposed to solve the transient responses of general microwave passive structures. EVFE simulates the signal envelope rather than the original signal waveform by de-embedding the carrier from the time-domain wave equation. The sampling rate of the time-domain waveform is only governed by the Nyquist rate of the envelope, rather than that of the carrier in traditional time-domain simulators. Compared to traditional finite-element time-domain (FETD) methods, the computational cost can be dramatically reduced when the signal envelope-to-carrier ratio is very small. It also provides much higher computational efficiency than frequency-domain finite-element methods for simulating frequency responses over certain bandwidth. This technique is applied to solve a waveguide structure with a dielectric post discontinuity and a microstrip patch antenna. The accuracy and efficiency is demonstrated and compared with traditional unconditionally stable FETD methods.**

***Index Terms*—Envelope simulation, finite-element time-domain method, full-wave approach, time-domain modeling.**

I. Introduction

As THE COMPLEXITY of microwave circuits and operating frequencies increase, development of computer-aided design (CAD) tools to efficiently and accurately predict performance of high-frequency circuits is in higher demand. Most commercial design tools are based on a circuit approach, in which the S-parameter matrix and the harmonic-balance method are applied by dividing the circuit into small elements and cascading the characteristic of each element to obtain the overall system performance. Consequently, the electromagnetic (EM) effects such as surface wave leakage, coupling between closely spaced components or subsystems, and packaging effects are ignored or approximated at best. Successful high-frequency circuit design requires full-wave analysis, which includes electromagnetic effects by solving Maxwell's equations and taking into account the above EM interaction.

Most full-wave numerical methods for characterizing EM effects require extensive computer resources. In the last few years, there has been a significant improvement in both computer resources and EM simulation techniques so that the use of full-wave EM simulators becomes economically viable. Commercially available EM simulators consist mainly of two types: frequency domain and time domain. The first includes the 2.5-D solvers based on the method of moments and the three-dimensional (3-D) solvers based on the finite-element method (FEM). However, with the increasing size of systems with complex spectral behavior, solving problems at many discrete frequency points with sufficient resolution can be very time consuming. The second type of EM simulator is the finite-difference time-domain (FDTD) technique. The FDTD method, first introduced by Yee [1], has been the most popular method for the simulation of the transient EM wave phenomena for the past few decades. The FDTD method shows great promise in its flexibility in handling a variety of circuit configurations, such as filters, microstrip transitions, bond wires, bridges, etc. [2], [3]. It was later successfully implemented to solve several nonlinear microwave circuits, including crosstalk and packaging effects [4], [5]. The most important benefit of time-domain analysis is that a broad-band pulse can be applied as the excitation so that broad-band frequency-dependent scattering parameters can be calculated from a single computation. Despite its programming simplicity, it has suffered from the staircase approximation when the method is applied to geometries with curvature and/or fine features [6].

Although less popular, the finite-element time-domain (FETD) method has some important advantages over the standard FDTD method. Since the grid is unstructured, it offers superior versatility in modeling complex geometries. Furthermore, the use of vector elements [7] provides a very natural way of enforcing tangential continuity to the electric field and normal continuity to the magnetic flux density at material interfaces, thus further enhancing modeling accuracy. Several variants of FETD have been proposed and implemented in [8]–[13]. Wong's technique [12] has been extended by Chang *et al.* [14] and provided solutions of a microwave amplifier and an injection-locked oscillator. Among those FETD techniques, the time derivatives were approximated by a difference scheme, resulting in an explicit time-domain scheme. Therefore, the methods described above are only conditionally stable with time steps which are typically equal to, or smaller than, those imposed by the FDTD technique. An implicit time-domain scheme, on the other hand, involving one-time matrix inversion and field updating of every time step, has been developed by Gedney and Navsariwala [15] for the solution of the second-order electric-field Maxwell's equations. For the field approximation, the use of the one-form Whitney element [7] gives degrees of freedom associated with its edges because it only enforces tangential continuity of vector fields. The second-order differential time-dependent formula employs a time-integration method based on the Newmark–Beta method

Manuscript received April 5, 2002; revised August 21, 2002.

H.-P. Tsai is with Mindspeed Technologies, Newport Beach, CA 92660 USA (e-mail: hsptsai@ieee.org).

Y. Wang and T. Itoh are with the Department of Electrical Engineering, University of California at Los Angeles, Los Angeles, CA 90095 USA (e-mail: ywang@ee.ucla.edu; itoh@ee.ucla.edu).

Digital Object Identifier 10.1109/TMTT.2002.805190

[16]. With appropriate values of the parameters controlling the accuracy and the stability of the scheme, the Newmark method yields an unconditionally stable scheme with second-order accuracy [15], [17]. The extended technique, called the unconditionally stable extended (USE) FETD method [18], was implemented to solve a microwave amplifier and illustrated a significant improvement in computational efficiency over the conditionally stable FETD method in terms of CPU time.

The envelope finite-element (EVFE) technique recently proposed by Wang *et al.* [19] is an even more efficient full-wave time-domain modeling scheme. When the unconditionally stable FETD scheme is applied to solve a problem which has a much narrower signal bandwidth than the carrier frequency, the time step is still constrained by the maximum operating frequency, and much computation time is unnecessarily wasted. By adapting the concept of envelope simulation, this computation expense can be saved. The circuit envelope technique has been recently introduced in [20] and exploited in HP EEsof's ADS and MDS circuit design software. By discretizing and simulating the signal envelopes on the defined carrier, the envelope waveform has to be sampled faster than only the Nyquist rate of the signal envelope. This method has been proven to be much more efficient than transient simulators like SPICE for narrow-band cases. A similar concept is introduced to solve the time-domain wave-propagation equations in [19], which results in a new type of transient electromagnetic solver, the so-called EVFE technique. The excitation and unknown fields are represented by a modulated format. The carrier signal is then de-embedded from the time-domain wave equation so that only the time-varying complex envelopes of the electromagnetic waves are simulated. Since only the signal envelope needs to be sampled, a much sparser time step can be used than in the FDTD or FETD techniques, which results in much higher computation efficiency when the envelope-to-carrier ratio is small.

This paper extends the EVFE technique as a 3-D full-wave solver for microwave passive structures. The detail EVFE formulation is presented in Section II. To confirm the validity and efficiency, the frequency-dependent scattering parameters have been calculated for a guided wave structure and a line-fed rectangular patch antenna. Calculated results are compared with other numerical techniques, such as FEM, FDTD, and unconditionally stable FETD in Section III. Finally, conclusions are drawn in Section IV.

II. Formulations Using Absorbing Boundary Conditions (ABCs)

This section derives the formulation of the EVFE technique in three dimensions. The computational domain is terminated by first-order ABCs.

A. EVFE Formulation

The general time-harmonic form of the second-order electric-field Maxwell's equations in a lossless region is

$$\nabla \times \nabla \times \left(\frac{\mathbf{E}}{\mu}\right) + \varepsilon \frac{\partial^2 \mathbf{E}}{\partial t^2} + \frac{\partial \mathbf{J}}{\partial t} = 0 \qquad (1)$$

where $\varepsilon = \varepsilon_r \varepsilon_0$ and $\mu = \mu_r \mu_0$ are the permittivity and permeability, respectively and J denotes an applied current density inside the solution domain. If the operational frequency bandwidth is narrow, then the first-order ABCs based on the traveling-wave assumption can be used to terminate the boundary. The Silver-Muller condition (first-order ABCs) used in FDTD or FETD can be expressed as

$$\hat{n} \times \nabla \times \mathbf{E} = -\hat{n} \times \hat{n} \times \frac{1}{v_g} \frac{\partial \mathbf{E}}{\partial t} \qquad (2)$$

where the unit vector $\hat{n}$ represents the outward surface normal to the boundary and v_g is the wave velocity traveling toward the boundary. By defining the carrier frequency, ω_c, the field component and the current density can be written in a modulated signal format

$$\mathbf{E}(t) = \mathbf{V}(t)\, e^{j\omega_c t}, \mathbf{J}(t) = \mathbf{U}(t)\, e^{j\omega_c t} \qquad (3)$$

where $\mathbf{V}(t)$ and $\mathbf{U}(t)$ are the time-varying complex envelopes of the electric field and the excitation current at the carrier frequency, respectively. Substituting (3) into (1)–(2) and dividing both sides by $e^{j\omega_c t}$ yields the envelope equation

$$\nabla \times \nabla \times \left(\frac{\mathbf{V}}{\mu}\right) + \varepsilon_0 \varepsilon_r \left(-\omega_c^2 \mathbf{V} + 2j\omega_c \frac{\partial \mathbf{V}}{\partial t} + \frac{\partial^2 \mathbf{V}}{\partial t^2}\right) = -\left(j\omega_c \mathbf{U} + \frac{\partial \mathbf{U}}{\partial t}\right). \qquad (4)$$

The ABC equation in the EVFE simulation becomes

$$\hat{n} \times \nabla \times \mathbf{V} = -\hat{n} \times \hat{n} \times \frac{1}{v_g} \left[\left(\frac{\partial \mathbf{V}}{\partial t} + j\omega \mathbf{V}\right)\right]. \qquad (5)$$

The inner product of (4) with a testing function $\mathbf{T}$ leads to a weak integral form given by

$$\begin{aligned} &\int_v \nabla \times \left(\frac{\mathbf{V}}{\mu}\right) \cdot \nabla \times \mathbf{T} dv + \int_v \left(-\varepsilon\, \omega_c^2\right) \mathbf{T} \cdot \mathbf{V} dv \\ &\quad + \int_\Omega \left(\frac{1}{\mu v_g}\right) (\hat{n} \times \mathbf{T}) \cdot \left[\hat{n} \times \left(\frac{\partial \mathbf{V}}{\partial t} + j\omega_c \mathbf{V}\right)\right] ds \\ &\quad + \int_v (2j\omega_c \varepsilon) \mathbf{T} \cdot \frac{\partial \mathbf{V}}{\partial t} dv + \int_v \varepsilon \mathbf{T} \cdot \frac{\partial^2 \mathbf{V}}{\partial t^2} dv \\ &= \int_v -\mathbf{T} \cdot \left(j\omega_c \mathbf{U} + \frac{\partial \mathbf{U}}{\partial t}\right) dv \end{aligned} \qquad (6)$$

where the volume v represents the truncated solution domain containing an arbitrarily shaped 3-D discontinuity. The surface Ω denotes all open surfaces. To solve (6) numerically, we discretize the space domain with tetrahedral elements and express the envelope field in terms of basis functions associated with the edges of the elements as follows:

$$\mathbf{V} = \sum_{j=1}^{N} W_j^{(1)} v_j \quad \mathbf{U} = \sum_{j=1}^{N} W_j^{(1)} j_j \qquad (7)$$

where N is the total number of the edges. $W_j^{(1)}$, a one-form Whitney element [7], is the vector basis function associated with edge j, and v_j and j_j are the circulations of the electric field and the current along the edge j, respectively. The application of Galerkin's process results in a system of ordinary differential equations

$$[G_A]\, \underline{v} + [G_B]\, \frac{\partial \underline{v}}{\partial t} + [G_C]\, \frac{\partial^2 \underline{v}}{\partial t^2} = [I] \left(j\omega_c \underline{j} + \frac{\partial \underline{j}}{\partial t}\right) \qquad (8)$$

where $\underline{v}$ and $\underline{j}$ are the vectors and $[G_A]$, $[G_B]$, $[G_C]$, and $[I]$ are time-independent matrices. Those terms are given by

$$[G_C]_{ij} = \int_v \varepsilon W_i^{(1)} \cdot W_j^{(1)} dv$$
$$[G_D]_{ij} = \int_\Omega \left(\frac{1}{\mu v_g}\right) \left(\hat{n} \times W_i^{(1)}\right) \cdot \left(\hat{n} \times W_j^{(1)}\right) ds$$
$$[G_A]_{ij} = \int_v \frac{1}{\mu} \nabla \times W_i^{(1)} \cdot \nabla \times W_j^{(1)} dv - \omega_c^2 [G_C]_{ij} + j\omega_c [G_D]_{ij}$$
$$[G_B]_{ij} = 2j\omega_c [G_C]_{ij} + j\omega_c [G_D]_{ij}$$
$$[I]_i = \int_v -W_i^{(1)} \cdot dv.$$

The discretization in the time domain is based on the Newmark–Beta formulation [16]

$$\begin{cases} \dfrac{d^2 v}{dt^2} = \frac{1}{\Delta t^2}\left[v(n+1) - 2v(n) + v(n-1)\right] \\ \dfrac{dv}{dt} = \dfrac{1}{2\Delta t}\left[v(n+1) - v(n-1)\right] \\ v = \beta v(n+1) + (1-2\beta) v(n) + \beta v(n-1) \end{cases} \tag{9}$$

where $v(n) = v(n\Delta t)$ is the discrete-time representation of $v(t)$. Gedney [15] has proven that unconditional stability for the FETD technique is achievable by choosing the interpolation parameter $\beta \geq 1/4$. The unconditional stability condition for EVFE techniques, derived by Wang [19] in a similar way, is satisfied only when $\beta = 1/4$. By choosing $\beta = 1/4$, an unconditionally stable EVFE update scheme allows the time step to be chosen in order to give a specific accuracy without being constrained by the mesh size. Therefore, the resulting update scheme is

$$\left\{\frac{[G_C]}{\Delta t^2} + \frac{[G_B]}{2\Delta t} + \frac{[G_A]}{4}\right\} v(n+1) = \left\{\frac{2[G_C]}{\Delta t^2} - \frac{[G_A]}{2}\right\} v(n) + \left\{-\frac{[G_C]}{\Delta t^2} + \frac{[G_B]}{2\Delta t} - \frac{[G_A]}{4}\right\} v(n-1) + [I]\left(j\omega_c \underline{j} + \frac{\partial \underline{j}}{\partial t}\right). \tag{10}$$

The resulting equations are solved using multifrontal sparse Gaussian elimination.[1] For each time step, the updating of the electric field requires solving a matrix equation of the type $[A]X = [B]$, where $[A]$ is a well-conditioned positive-definite sparse matrix. Since this matrix is not time dependent, it can be factorized and inverted only once before time marching.

B. A Simplified Source Excitation for the Time-Domain Method

Choosing a proper source excitation scheme is one of the critical factors in successful EM simulators. Rigorously, one needs to have prior knowledge of the transverse-field distribution to launch a proper propagation mode, either by using analytical approaches, or by running a preliminary simulation to obtain the correct transverse-field profile. However, if we allow a certain distance between the source and the observation points, it is more effective to use pulsed signal source and still realize successful excitation of the desired propagation mode. We adapt a simplified source-excitation technique proposed by Zhao [21] for the FETD and EVFE methods, and excite only several electric probes uniformly at several cells away from the near-end boundary inside the EM grid. All the other field components at the remaining part in the source plane are left "free-running" and updated by the FETD or EVFE equations. This simplified excitation scheme is different from both Sheen's [2] and Zhang's [22] approaches. The assumption of artificial PEC or PMC wall which induces dc source distortions is not a necessity, and ABCs can be applied directly without any special treatment.

[1] Provided by the Harwell Subroutine Library, AEA Technology, Oxfordshire, U.K.

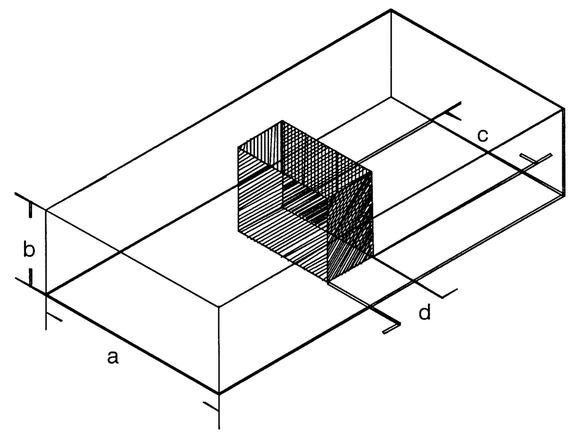


Fig. 1. Dielectric post discontinuity in a rectangular waveguide. $a = 22.86$ mm, $b = 10.16$ mm, $c = 12$ mm, $d = 6$ mm, and the dielectric constant of the slab $\varepsilon_r = 8.2$. [24].

III. SIMULATION RESULT

A FORTRAN code based on the 3-D EVFE formulation given in Section II has been implemented. To confirm the validity and efficiency of the proposed technique, in this section we apply the FETD and EVFE schemes to analyze waveguide and microstrip discontinuities. Because of the unconditional stability, the time steps in FETD and EVFE are all only constrained by the Nyquist rules.

A. Waveguide Discontinuities

The simulation structure is a rectangular waveguide with a dielectric post discontinuity shown in Fig. 1. The same geometry was analyzed by Wang *et al.* using FEM [23]. The waveguide has a width $a = 22.86$ mm and height $b = 10.16$ mm. The dielectric post has a height equal to that of the guide, a width $c = 12$ mm, and a length $d = 6$ mm. The dielectric constant of the post is 8.2. The excitation of an electric probe consisting of a modulated Gaussian pulse is applied in between the ABC boundary and the discontinuity to excite a TE_{10} mode inside the waveguide. The carrier frequency is $f_c = 10$ GHz, and the half bandwidth is $\Delta f = 2.5$ GHz. The EVFE excitation in (3) is represented as

$$\mathbf{J}(t) = \mathbf{U}(t)\, e^{j\omega_c t} = \exp\left[-4\frac{(t-t_0)^2}{T^2}\right] \bullet e^{j\omega_c t} \tag{11}$$

where $T = 4/(\pi\Delta f)$, and $t_0 = 1.4T$. In FETD, a modulated Gaussian pulse defined by

$$\mathbf{J}(t) = \exp\left[-4\left(\frac{t-t_0}{T}\right)^2\right] \sin\omega_c(t-t_0) \tag{12}$$

is applied.

TABLE I
COMPUTATION TIMES OF THE STRUCTURE IN FIG. 1 ON AN AMD 1.2-GHz PC MACHINE

	EVFE	FETD
Mesh element Preprocessing	5 min.	3.5 min.
Matrix Inversion	43 sec.	30 sec.
Time stepping	85 sec. (NT = 240) (Δt = 20 ps)	7 min. (NT = 1200) (Δt = 5 ps)

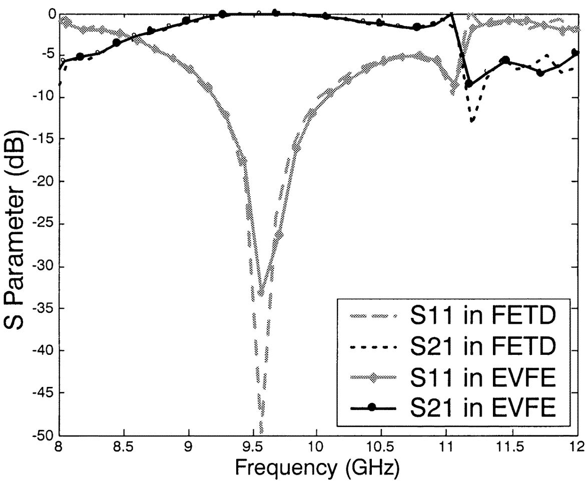


Fig. 2. Reflection coefficient of the dielectric post discontinuity in a rectangular waveguide.

The waveguide is terminated with simple first-order ABCs, located far away from the dielectric post discontinuity. The mesh sizes are $\Delta x_{\min} = \Delta y_{\min} = \Delta z_{\min} = 1$ mm, and $\Delta t_{\max} = 1.9$ ps, if imposed by the stability criteria of the FDTD technique. The number of tetrahedrons in the finite element region is 72 373. A comparison of computational time between the unconditionally stable FETD method and the EVFE method is listed in Table I when the simulation is executed with a PC with a 1.2-GHz AMD Athlon Processor. The unknowns in FETD are real numbers, but are complex numbers in EVFE. Thus, the matrix inversion in the latter method requires more time. The time steps are 5 ps in FETD and 25 ps in EVFE. Since the two simulations have the same time duration, the total number of time steps is reduced from 1200 in unconditionally stable FETD to 240 in EVFE.

Fig. 2 shows the magnitude of S_{11} and S_{21} in the spectral domain for the structure shown in Fig. 1 using the unconditionally stable FETD method and the EVFE method. Fig. 3 shows the transient response of incident, reflected, and transmitted waves. The results show that in this case the time step in the EVFE method can be chosen to be five times larger than that in the unconditionally stable FETD method to obtain the same quantitative accuracy.

B. *Microstrip Discontinuities*

For microstrip circuits, the line-fed rectangular microstrip antenna shown in Fig. 4 is analyzed. The antenna is printed on an

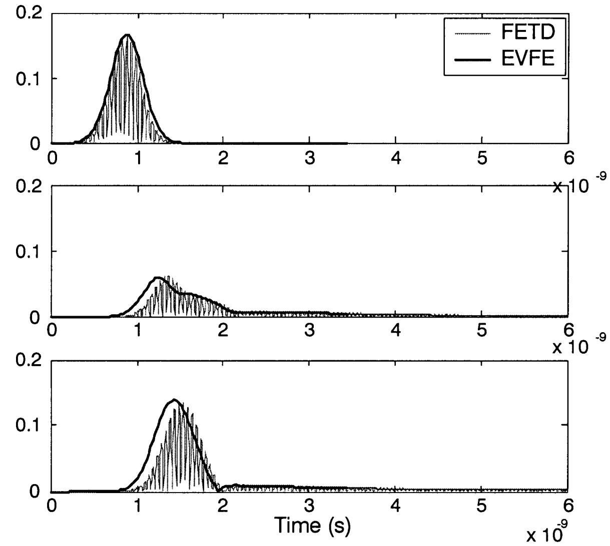


Fig. 3. Transient signal and envelope response of incident, reflected, and transmitted waves of the dielectric post discontinuity in a rectangular waveguide. The electric field is obtained by the FETD method and the field envelope is obtained by the EVFE method.

Rogers Duroid substrate with $\epsilon_r = 2.2$ and height, 31 mil. This structure is first analyzed by Sheen *et al.* [2] using FDTD, then compared by Polycarpou *et al.* [24] using FEM. The computation effort reported is as follows. For wide-band simulation from dc to 20 GHz, the FEM is more computationally intensive than the FDTD method. Polycarpou *et al.* [24] has reported that the computational time for FEM was approximately 30 min per frequency point in the lower frequency range and 15 min per frequency point in the upper frequency range on a 370 IBM RISC/6000 UNIX workstation. Sheen's FDTD code [2] was run on a Silicon Graphics Power Indigo2 workstation with an R8000 processor and the computation time required is around 45 min with $\Delta t = 0.441$ ps and 8192 time steps.

1) Wide-Band Circuit Simulation: To verify the algorithm and compare the computation effort, the geometry was run using both the unconditionally stable FETD and EVFE codes for a wide frequency range from dc to 20 GHz. The mesh sizes, bigger than those suggested in [2] and [24], are $\Delta x_{\min} = 31$ mil, $\Delta y_{\min} = 32.33$ mil, $\Delta z_{\min} = 39.37$ mil, and $\Delta t_{\max} = 1.65$ ps if the stability criteria of the FDTD technique is imposed. The rectangular patch antenna is $13\Delta y \times 16\Delta z$. The resulting finite-element mesh consists of $6 \times 25 \times 36 \times 5$ tetrahedral elements and a total of 33 141 unknowns. The source plane is located at $z = 240$ mil, and three vertical electric probes are excited between the microstrip fed line and ground plane. The observation is located at $z = 360$ mil.

The excitation applied in EVFE is the same form as (12), but to compared with FETD, the parameters are changed as $f_c = 10$ GHz, $\Delta f = 10$ GHz, $T(\text{ps}) = 500/\Delta f(\text{GHz})$, and $t_0 = 3T$. A nonmodulated Gaussian pulse applied in FETD is given as

$$\mathbf{J}(t) = \exp\left[-\left(\frac{t-t_0}{T}\right)^2\right] \tag{13}$$

where $\Delta f = 20$ GHz, $T(\text{ps}) = 500/\Delta f(\text{GHz})$, and $t_0 = 3T$.

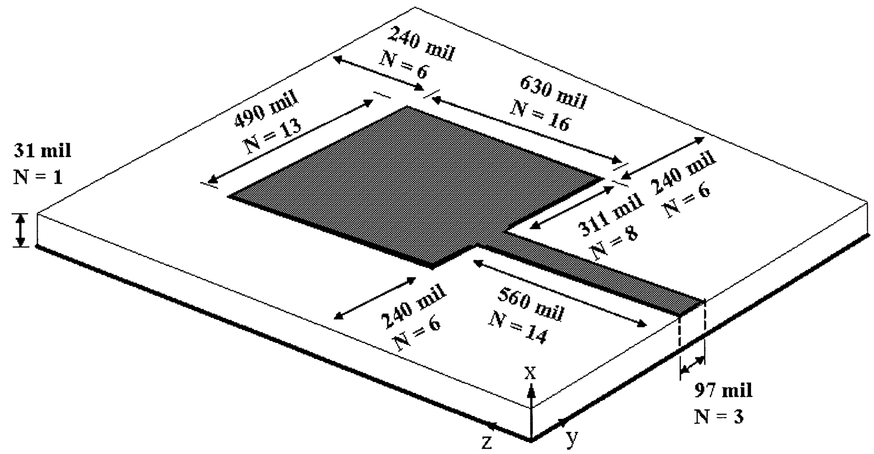
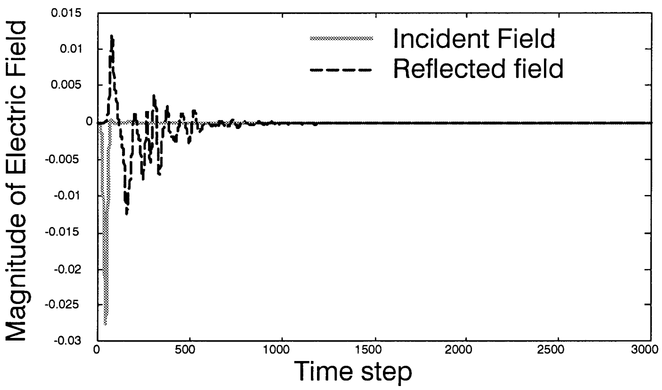
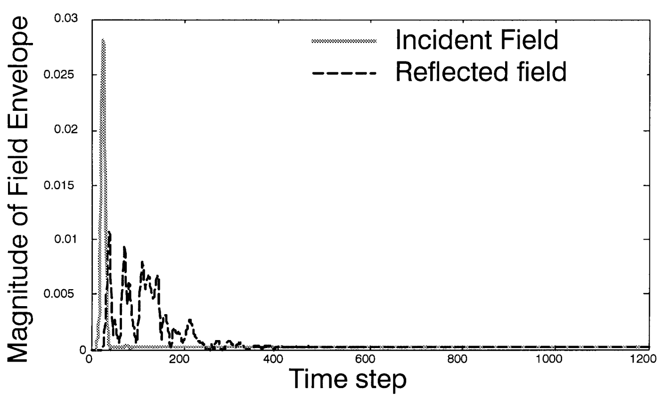


Fig. 4. Simulated dimension of line-fed rectangular microstrip antenna. (N is the mesh number.).



(a)



(b)

Fig. 5. Transient response at the input port of line-fed rectangular microstrip antenna. (a) Electric field simulated by FETD, with $\Delta t = 2$ ps. (b) Electric-field envelope simulated by EVFE, with $\Delta t = 5$ ps.

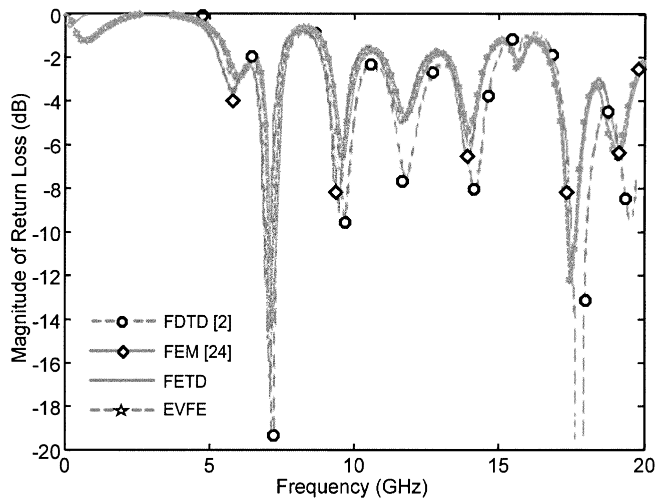


Fig. 6. Return loss of the rectangular antenna from dc to 20 GHz.

Fig. 5 shows the corresponding transient response at the input port of the microstrip antenna. A comparison of the input return loss obtained using the FDTD, FEM, FETD, and EVFE methods is shown in Fig. 6. We choose $\Delta t = 2$ ps and 3000 steps in FETD. For EVFE, $\Delta t = 5$ ps and 1200 steps are used. The simulation is done on an AMD 1.4-GHz PC Machine. The results show a fairly good agreement with the FEM and the FDTD method. The total computation time obtained by either the FETD or EVFE method is only approximately 10 min for the over-all simulation. If the mesh element preprocessing time and matrix inverting time is excluded, the average simulation time for each time step is 0.4 s in EVFE and 0.175 s in FETD.

If the frequency band of interest is much smaller than the carrier frequency, then we can use a bigger time-step size and reduce computation time significantly.

2) Narrow-Band Circuit Simulation: To illustrate the efficiency of EVFE, a narrow-band analysis of the fundamental radiation frequency is demonstrated and compared with FETD. The excitation parameters in (11) and (12) are $f_c = 7$ GHz, $\Delta f = 1$ GHz, $T = 1/(2\Delta f)$, and $t_0 = 1.78T$. To have a good frequency resolution, every transient simulation is executed in the same time duration, 10 ns. Fig. 7 shows the input return loss from FETD and EVFE at different time step sizes. From this analysis, FETD with 5-ps step size and EVFE with 100-ps step size give an accurate prediction of the radiation frequency. Based on the Nyquist theory, the sampling rates of FETD and EVFE are inversely proportional to the maximum operational frequency and the half bandwidth, respectively. Thus, the maximum time-step size in EVFE should be around 5.6 times larger than in FETD in this analysis. However, the result of this example shows the maximum time step in EVFE is bigger than the step allowed by the Nyquist theory. This is because the EVFE technique samples the slowly varying signal envelope

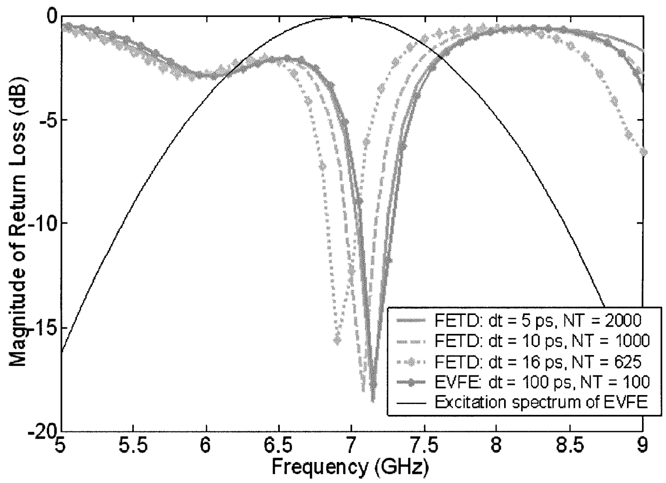


Fig. 7. Return loss of the rectangular antenna in narrow-band simulation. (dt is the time step size and NT is the time-step number.).

TABLE II
WIDE-BAND COMPUTATION TIMES OF THE MICROSTRIP ANTENNA BY AN AMD 1.4-GHZ PC MACHINE

	EVFE	FETD
Mesh element Preprocessing	63 sec.	55 sec.
Matrix Inversion	65 sec.	20 sec.
Time stepping (Wideband)	8 min. (NT = 1200) (Δt = 5 ps)	8 min. 45 sec. (NT = 3000) (Δt = 2 ps)
Time stepping (Narrowband)	40 sec. (NT = 100) (Δt = 100 ps)	5 min. 50 sec. (NT = 2000) (Δt = 5 ps)

rather than the fast oscillating signal waveform, so that the dispersion error in EVFE is less serious than other time-domain approaches. Wang *et al.* [19] has shown that EVFE has the minimum numerical dispersion error at the carrier frequency by calculating the resonant frequency of a 2-D rectangular cavity. Assuming a perfect spatial discretization scheme, the relationship between the dispersion error and time domain discretization for the 3-D FETD and EVFE methods are predicted as follows:

$$\text{Error} \propto (f - f_c)^3 \Delta t^2, \qquad \text{in EVFE} \tag{14}$$

$$\text{Error} \propto f^3 \Delta t^2, \qquad \text{in FETD} \tag{15}$$

where f is the operation frequency and f_c is the carrier frequency. The FETD method can be considered as one of the special cases of the EVFE with carrier frequency at dc and the dispersion error increases as frequency increases. Therefore, the EVFE method gives better performance overall in dispersion error than the FETD method assuming the carrier frequency is appropriately defined at the center of the frequency of interest. In the above analysis, the solution of EVFE around the carrier frequency 7 GHz suffers less dispersion error than the FETD analysis so that choosing a larger time-step size still gives us a fairly accurate result. When we increase the step size to 120 ps, the solution is also stable, but the corresponding radiation frequency shifts due to increased dispersion error. The overall computation times for the same accurate results are around 8 min in 2000 time steps using FETD and 2 min 48 s in 100 steps using EVFE. Table II lists the computation effort of the patch antenna in details.

IV. CONCLUSION

Based on the unconditionally stable FETD method, the 3-D EVFE has been implemented by sampling signal-field envelopes. This technique has been applied to solve a guided-wave structure and a microstrip patch antenna. The scattering parameters have been verified by the comparison with results from FDTD, FETD, and FEM. It has a number of advantages over other numerical techniques, including geometric flexibility in modeling curved surfaces and inhomogeneous materials, better efficiency in broad-band simulation than frequency-domain techniques, better efficiency in narrow-band simulation than other time-domain techniques by using a larger time-step size because of unconditional stability, the envelope sampling concept, and less dispersion error. This paper has shown that the EVFE method can improve the computation efficiency of the unconditionally stable FETD method by more than a factor of the carrier-to-envelop frequency ratio. When the need for microwave circuit simulators and CAD tools is increasing with the development of a wide range of commercial products operating at a high frequency, this method becomes the best candidate for cost-effective full-wave simulation. Finally, this technique can be easily extended to incorporate lumped microwave circuits by using the method introduced in [18].

REFERENCES

[1] K. S. Yee, "Numerical solution of initial boundary value problems involving Maxwell's equations in isotropic media," *IEEE Trans. Antennas Propagat.*, vol. AP-14, pp. 302–307, May 1966.

[2] D. M. Sheen, S. M. Ali, M. D. Abouzahra, and J. A. Kong, "Application of the three-dimensional finite-difference time-domain method to the analysis of planar microstrip circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 38, pp. 849–857, July 1990.

[3] A. Christ and H. L. Hatnagel, "Three-dimensional finite-difference method for the analysis of microwave-device embedding," *IEEE Trans. Microwave Theory Tech.*, vol. 35, pp. 688–696, Aug. 1987.

[4] C.-N. Kuo, B. Houshmand, and T. Itoh, "Full-wave analysis of packaged microwave circuits with active and nonlinear devices: An FDTD approach," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 819–826, May 1997.

[5] M. Chen, W. R. Deal, B. Houshmand, and T. Itoh, "Global time-domain full-wave analysis of microwave FET oscillators and self-oscillating mixers," *IEEE MTT-S Int. Microwave Symp. Dig.*, pp. 1135–1138, June 1998.

[6] A. C. Cangellaris and D. B. Wright, "Analysis of the numerical error caused by the stair-stepped approximation of a conducting boundary in FDTD simulations of electromagnetic phenomena," *IEEE Trans. Antennas Propagat.*, vol. 39, pp. 1518–1525, Oct. 1991.

[7] A. Bossavit, "Whitney forms: A class of finite elements for three-dimensional computations in electromagnetism," *Proc. Inst. Elect. Eng.*, pt. A, vol. 135, no. 8, pp. 495–500, Nov. 1988.

[8] K. Mahadevan, R. Mittra, D. Rowse, and J. Murphy, "Edge-based finite-element frequency and time domain algorithms for RCS computation," *Proc. IEEE Antennas and Propagation Symp. Dig.*, vol. 3, pp. 1680–1683, June 1993.

[9] J. T. Elson, H. Sangani, and C. H. Chan, "An explicit time-domain method using three-dimensional Whitney elements," *IEEE Microwave Guided Wave Lett.*, vol. 7, pp. 607–610, Sept. 1994.

[10] J. F. Lee, "WETD: A finite-element time-domain approach for solving Maxwell's equations," *IEEE Microwave Guided Wave Lett.*, vol. 4, pp. 11–13, Jan. 1994.

[11] Z. S. Sacks and J. F. Lee, "A finite-element time-domain method using prism elements for microwave cavities," *IEEE Trans. Magn.*, vol. 37, pp. 519–527, Nov. 1995.

[12] M. F. Wong, O. Picon, and V. F. Hanna, "A finite-element method based on Whitney forms to solve Maxwell equations in the time-domain," *IEEE Trans. Magn.*, vol. 31, pp. 1618–1621, May 1995.

[13] M. Feliziani and F. Marradei, "Hybrid finite-element solutions of time dependent Maxwell's curl equations," *IEEE Trans. Magn.*, vol. 31, pp. 1330–1335, May 1995.

[14] S.-H. Chang, R. Coccioli, Y. Qian, and T. Itoh, "A global finite-element time domain analysis of active nonlinear microwave," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 2410–2416, Dec. 1999.

[15] A. D. Gedney and U. Navsariwala, "An unconditionally stable finite element time-domain solution of the vector wave equation," *IEEE Microwave Guided Wave Lett.*, vol. 5, pp. 332–334, Oct. 1995.

[16] N. M. Newmark, "A method of computation for structural dynamics," *J. Eng. Mechan. Divis., ASCE*, vol. 85, pp. 67–94, July 1959.

[17] C. Zienkiewicz, "A new look at the Newmark, Houboult and other time stepping formulas. A weighted residual approach," *Earthquake Eng. Struct. Dynam.*, vol. 5, pp. 413–418, 1977.

[18] H.-P. Tsai, Y. Wang, and T. Itoh, "An *u*nconditionally *s*table *e*xtended (USE) finite element time domain solution of active nonlinear microwave circuits using perfectly matched layers," *IEEE Trans. Microwave Theory Tech.*, vol. 50, pp. 2226–2232, Oct. 2002.

[19] Y. Wang and T. Itoh, "Envelope-finite element (EVFE) technique – A more efficient time domain scheme," *IEEE Trans. Microwave Theory Tech.*, vol. 49, pp. 2241–2247, Dec. 2001.

[20] H. S. Yap, "Designing to digital wireless specifications using circuit envelope simulation," in *Proc. 1997 Asia Pacific Microwave Conf.*, pp. 173–176.

[21] A. P. Zhao and V. Raisanen, "Application of a simple and efficient source excitation technique to the FDTD analysis of waveguide and microstrip circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 44, pp. 1535–1539, Dec. 2001.

[22] X. L. Zhang and K. K. Mei, "Time-domain finite-difference approach to the calculation of the frequency-dependent characteristics of microstrip discontinuities," *IEEE Trans. Microwave Theory Tech.*, vol. 36, pp. 1775–1787, Dec. 1988.

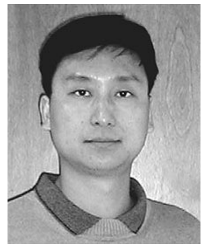
[23] J.-S. Wang and R. Mittra, "Finite element analysis of MMIC structures and electronic packages using absorbing boundary conditions," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 441–449, Mar. 1994.

[24] A. C. Polycarpou, P. A. Tirkas, and C. A. Balanis, "The finite-element method for modeling circuits and interconnects for electronic packaging," *IEEE Trans. Microwave Theory Tech.*, vol. 45, pp. 1868–1874, Oct. 1997.

Hsiao-Ping Tsai (S'00) received the M.S. Degrees in Electrical Engineering from National Taiwan Univlsersity, Taiwan, R.O.C., in 1997, and the Ph.D. degree in electrical engineering from University California at Los Angeles (UCLA) in March 2002.

From 1997 to 1998, she was a Teaching Assistant with National Taiwan University. During May–July 2002, she was a Postdoctoral Researcher at UCLA, where she conducted research on simulation of semiconductor devices for IC design. Since August 2002, she has been with Mindspeed Technologies, Newport Beach, CA, as a Package Application Engineer. Her research interests include the time-domain full-wave modeling of nonlinear semiconductor devices and microwave circuits.

Dr. Tsai received the Honorable Mention Award in the student paper competitions of the IEEE International Microwave Symposium in 2000 and 2002.



Yuanxun Wang (S'96–M'99) received the B.S. degree in electrical engineering from the University of Science and Technology of China, Hefei, in 1993, and the M.S. and Ph.D. degrees in electrical engineering from University of Texas at Austin in 1996 and 1999, respectively.

From 1995 to 1999, he was a Research Assistant in the Department of Electrical and Computer Engineering, University of Texas at Austin. In 1999, he joined the Department of Electrical Engineering, University of California at Los Angeles, where he is currently an Assistant Professor. His research interests include enabling technology for RF and microwave front-ends in wireless communication and radar systems, as well as numerical modeling, simulation, and feature-extraction techniques for microwave circuits, antennas, and electromagnetic scattering. He has authored or co-authored more than 60 refereed journal and conference papers.

Tatsuo Itoh (S'69–M'69–SM'74–F'82) received the Ph.D. degree in electrical engineering from the University of Illinois at Urbana-Champaign, in 1969.

From September 1966 to April 1976, he was with the Electrical Engineering Department, University of Illinois at Urbana-Champaign. From April 1976 to August 1977, he was a Senior Research Engineer in the Radio Physics Laboratory, SRI International, Menlo Park, CA. From August 1977 to June 1978, he was an Associate Professor at the University of Kentucky, Lexington. In July 1978, he joined the faculty at The University of Texas at Austin, where he became a Professor of electrical engineering in 1981 and Director of the Electrical Engineering Research Laboratory in 1984. During the summer of 1979, he was a Guest Researcher at AEG-Telefunken, Ulm, Germany. In September 1983, he was selected to hold the Hayden Head Centennial Professorship of Engineering at The University of Texas at Austin. In September 1984, he was appointed Associate Chairman for Research and Planning of the Electrical and Computer Engineering Department, The University of Texas. In January 1991, he joined the University of California at Los Angeles, as Professor of Electrical Engineering and Holder of the TRW Endowed Chair in Microwave and Millimeter Wave Electronics. He was an Honorary Visiting Professor at the Nanjing Institute of Technology, Nanjing, China, and at the Japan Defense Academy. In April 1994, he was appointed as Adjunct Research Officer for the Communications Research Laboratory, Ministry of Post and Telecommunication, Japan. He currently holds a Visiting Professorship at The University of Leeds, Leeds, U.K., and is an External Examiner of the Graduate Program of the City University of Hong Kong. He has authored or co–authored 280 journal publications, 585 refereed conference presentations, and has written 30 books/book chapters in the area of microwaves, millimeter-waves, antennas and numerical electromagnetics. He has overseen 49 Ph.D. students.

Dr. Itoh is a member of the Institute of Electronics and Communication Engineers of Japan and Commissions B and D of USNC/URSI. He has been the recipient of numerous awards including the Shida Award presented by the Japanese Ministry of Post and Telecommunications in 1998, the Japan Microwave Prize in 1998, the IEEE Third Millennium Medal in 2000, and the IEEE Microwave Theory and Techniques Society (IEEE MTT-S) Distinguished Educator Award in 2000. He served as the editor of the IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES from 1983 to 1985. He serves on the Administrative Committee of the IEEE MTT-S. He was vice president of the IEEE MTT-S in 1989 and president in 1990. He was the editor-in-chief of IEEE MICROWAVE AND GUIDED WAVE LETTERS from 1991 through 1994. He was elected as an Honorary Life Member of the MTT-S in 1994. He was the chairman of the USNC/URSI Commission D from 1988 to 1990, and chairman of Commission D of the International URSI from 1993 to 1996. He is chair of the Long Range Planning Committee of the URSI. He also serves on advisory boards and committees of a number of organizations.