A New Algorithm for the Incorporation of Arbitrary Linear Lumped Networks into FDTD Simulators

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Abstract—The inclusion of lumped elements, both linear and nonlinear, into the finite-difference time-domain (FDTD) algorithm has been recently made possible by the introduction of the lumped-element FDTD method. Such a method, however, cannot efficiently and accurately account for two-terminal networks made of several lumped elements, arbitrarily connected together. This limitation can be removed as proposed in this paper by describing the network in terms of its impedance in the Laplace domain and by using appropriate digital signal-processing methodologies to fit the resulting description to Yee’s algorithm. The resulting difference equations allow an arbitrary two-terminal network to be inserted into one FDTD cell, preserving the full explicit nature of the conventional FDTD scheme and requiring a minimum number of additional storage variables. The new approach has been validated by comparison with the exact solution of a parallel-plate waveguide loaded with lumped networks in the transverse plane.

Index Terms—FDTD method, lumped networks, global modeling.

I. INTRODUCTION

The use of increasingly higher frequencies and of high-density microwave integrated circuits makes global electromagnetic simulators indispensable for accurate characterization of such circuits. The electromagnetic interaction among passive and active elements of the circuits must be accurately modeled to predict the overall performance. Advanced electromagnetic simulators are required to incorporate both passive and active distributed and lumped elements; additionally, they should be capable of evaluating possible radiation patterns from radiating elements, as well as common circuit parameters, such as conversion efficiencies, usually computed by circuit-based simulators. Among the currently available techniques for three-dimensional (3-D) electromagnetic analysis, the finite-difference time-domain (FDTD) method seems to be one of the best candidates to provide an efficient and powerful global electromagnetic tool.

The FDTD method has been originally proposed as a full-wave numerical technique for the electromagnetic analysis of microwave structures capable of incorporating nonlinearities. In recent years, a considerable effort has been made to incorporate lumped elements into this technique [1]–[4]. The so-called lumped-element FDTD (LE-FDTD) method can include single passive elements such as resistors, inductors, and capacitors, and also more complex nonlinear or active devices such as diodes and transistors. The LE-FDTD technique allows complex microwave circuits to be successfully analyzed; bridging the gap between electromagnetic-field and circuit-based simulators.

The main limitation of the LE-FDTD method is that it cannot easily and accurately account for two-terminal circuits consisting of the arbitrary connection of several lumped elements. To incorporate a multiple lumped circuit into the LE-FDTD scheme, one has to introduce at the appropriate node the relevant time-domain voltage–current (V–I) description of the circuit. As an alternative, the various elements of the multiple circuit can be placed at different nodes of the FDTD mesh.

The first approach requires a cumbersome derivation of the time-domain description of the multiple circuit. The simultaneous presence of different reactive elements would involve not only a difficult analytical derivation, but also a numerically heavy computation. Moreover, a new analytical preprocessing is required for any new circuit topology.

The latter approach, besides not leading to a unique topology, destroys the lumped nature of the network; thus, leading to significant inaccuracies.

This paper introduces an extension of the LE-FDTD technique [referred to as the lumped-network FDTD (LN-FDTD)] to incorporate arbitrary linear \( RLC \) one-port lumped networks. The lumped network is first described in terms of its impedance in the Laplace domain. After expressing the voltage and current in terms of the electromagnetic fields of the connected node, the resulting equation is fitted into Yee’s algorithm by using digital signal-processing techniques. The resulting difference equations allow an \( RLC \) lumped network to be incorporated into the FDTD method using only one FDTD cell, without the need to rewrite the FDTD code for each network topology. It is worth pointing out that this new approach preserves the full explicit nature of the standard FDTD method. Furthermore, it is efficiently implemented, requiring a minimum number of additional storage variables, and is based on a semi-implicit discretization scheme that
ensures good numerical stability. To validate the new formulation, a problem with an exact solution has been considered: the computation of the reflection coefficient of a finite-width parallel-plate waveguide loaded with lumped networks in the transverse plane. The results obtained by FDTD show good agreement with those calculated analytically.

II. THEORY

A. The Model

Consider the time-dependent Maxwell’s curl equations in free space

\[
\nabla \times \mathbf{E} = -j\mu_0 \frac{\partial \mathbf{H}}{\partial t} \tag{1}
\]

\[
\nabla \times \mathbf{H} = \epsilon_0 \frac{\partial \mathbf{E}}{\partial t}. \tag{2}
\]

According to the standard Yee’s FDTD scheme, these equations can be expressed in difference form as

\[
H^{n+\frac{1}{2}}_\alpha(\mathbf{r}_H) = H^{n-\frac{1}{2}}_\alpha(\mathbf{r}_H) - \frac{\Delta t}{\mu_0} \left[ \nabla \times \mathbf{E} \right]^{n}_\alpha(\mathbf{r}_H) \tag{3}
\]

\[
E^{n+1}_\alpha(\mathbf{r}_E) = E^n_\alpha(\mathbf{r}_E) + \frac{\Delta t}{\epsilon_0} \left[ \nabla \times \mathbf{H} \right]^{n+\frac{1}{2}}_\alpha(\mathbf{r}_E) \tag{4}
\]

where \( \alpha = x, y, z \) and \( \mathbf{r}_E, \mathbf{r}_H \) denote the spatial position at Yee’s cell of \( E_\alpha \) and \( H_\alpha \), respectively. In (3) and (4), the curl terms are written in compact form; detailed expressions for these terms can be found in [5].

As proposed by Sui et al. [1], to account for lumped elements into the FDTD method, a current density term \( \mathbf{J}_l \) is included into Ampère’s equation

\[
\nabla \times \mathbf{H} = \epsilon_0 \frac{\partial \mathbf{E}}{\partial t} + \mathbf{J}_l. \tag{5}
\]

Assuming that the lumped network is connected at the node \( \mathbf{r}_E \), and along the \( x \)-direction, the FDTD equation to update \( E^{n+1}_x(\mathbf{r}_E) \) in (4) becomes

\[
E^{n+1}_x(\mathbf{r}_E) = E^n_x(\mathbf{r}_E) + \frac{\Delta t}{\epsilon_0} \left[ \nabla \times \mathbf{H} \right]^{n+\frac{1}{2}}_x(\mathbf{r}_E) - \frac{\Delta t}{2\epsilon_0} \left( J^l_x(\mathbf{r}_E) + J^{l+1}_x(\mathbf{r}_E) \right). \tag{6}
\]

Note that, following [2], the current density has been discretized by using time average, so as to preserve the second-order accuracy of Yee’s scheme. The remaining equation (3) needs no modifications. For the sake of brevity, in the following, we will not specify the spatial point \( \mathbf{r}_E \), at which the equations are evaluated, because this point is the same for all the remaining difference equations.

To incorporate a lumped network into the FDTD method, consider first its V–I relation in the Laplace domain

\[
V_x(s) = Z(s)I_x(s) \tag{7}
\]

\( Z(s) \) being the impedance function, which can be expressed as

\[
Z(s) = \frac{\sum_{m=0}^{M} b_m s^m}{\sum_{m=0}^{M} a_m s^m}. \tag{8}
\]

In the above expression, \( M \) represents the order of the impedance function, i.e., the number of elements with memory: capacitors and inductors, in the lumped network. To incorporate (7) into the FDTD algorithm, the circuit quantities first have to be related to the electromagnetic fields; the resulting equation then have to be consistently expressed in the discrete time domain with Yee’s scheme.

B. Discretization of the V–I Relation

Following [2], the voltage is related to the electric field as \( V_x = \Delta_y \Delta_z E_x \) and the current to the current density as \( I_x = \Delta_y \Delta_z J_x \). Therefore, substituting (8) into (7) we obtain

\[
J_x(s) = \sigma_x(s)E_x(s) \tag{9}
\]

where

\[
\sigma_x(s) = \frac{\Delta_x}{\Delta_y \Delta_z} \sum_{n=0}^{M} a_n s^n \tag{10}
\]

represents an equivalent conductivity in the Laplace domain.

To include equations of the same type as (9) into the FDTD method, two different approaches could be adopted: the recursive convolution method [6] and auxiliary differential equation technique [7]. Both approaches have proven their validity in the context of the extension of the FDTD method to the inclusion of dispersive media. They can also be used to incorporate lumped networks into the FDTD method. Both approaches, however, are based on the discretization of a continuous time-domain model: a convolution integral and a high-order differential equation, respectively. In practice, the discretization and implementation of an arbitrary-order continuous time model is troublesome. In this paper, we employ a simpler and more systematic technique based on the application of the bilinear transformation into the \( Z \)-domain [8]

\[
s = \frac{2(1 - z^{-1})}{\Delta t(1 + z^{-1})} \tag{11}
\]

to (9). This approach can be interpreted as a numerical integration, by using the trapezoidal rule, of the continuous time version of (9) expressed as a high-order differential equation. However, it does not require one to explicitly express (9) in the continuous time domain.

As a result of the bilinear transformation (11), and after some algebraic manipulations, the current density term can be expressed in the \( Z \)-domain as

\[
\hat{J}_x(z) = \sigma_x(z)\hat{E}_x(z) \tag{12}
\]

where \( \sigma_x(z) \) represents an equivalent conductivity in the \( Z \)-domain. It is given by

\[
\sigma_x(z) = \frac{1 + \sum_{k=1}^{M} c_k z^{-k}}{\sum_{k=0}^{M} d_k z^{-k}} \tag{13}
\]

where the coefficients \( c_k \) and \( d_k \) are related to the coefficients \( a_n \) and \( b_m \) in (9), and to the temporal and spatial discretization steps. General expressions are given in the Appendix.
To obtain the difference form of (12), we simply have to consider the relationship \( z^{-k} \tilde{F}(z) \leftrightarrow F^{n-k} \), which leads to

\[
E_x^{n+1} + \sum_{k=1}^{M} c_k E_x^{n-k+1} = \sum_{k=0}^{M} d_k J_{lx}^{n-k+1}. \tag{14}
\]

This high-order difference equation constitutes the discrete time version of (7) written in terms of the electric field and current density. Note that (14) is coupled to (6) since both equations involve \( E_x^{n+1} \) and \( J_{lx}^{n+1} \). This coupling shows the semi-implicit nature of the formulation. To recover the full-explicit nature of the standard FDTD scheme, (14) and (6) must be solved for \( E_x^{n+1} \) and \( J_{lx}^{n+1} \) leading to the following difference equations:

\[
E_x^{n+1} = \frac{1}{d_0} \left( \frac{d_0 \Delta t}{\Delta t} \right) \left( \frac{d_0 \Delta t}{\Delta t} + \frac{1}{2} \right) \left( E_x^n + d_0 \left[ \nabla \times \vec{H} \right]_x \right)
- \frac{1}{2} A_x \left( \frac{d_1 - d_0}{2} \right) J_{lx}^{n+1}, \tag{15}
\]

\[
J_{lx}^{n+1} = \frac{1}{d_0} (E_x^{n+1} + A_x) \tag{16}
\]

where

\[
A_x = \sum_{k=1}^{M} \left( c_k E_x^{n-k+1} - d_k J_{lx}^{n-k+1} \right). \tag{17}
\]

The above equations allow lumped networks to be incorporated into FDTD simulators preserving the explicit nature of Yee’s scheme. With respect to the standard FDTD technique [(3) and (4)], the implementation of (15) and (16) require \( 2M-1 \) additional variables (\( M-1 \) for the electric field and \( M \) for the current density), and \( 2M+4 \) additional real products.

### C. Efficient Implementation of Lumped Networks into FDTD

To reduce memory requirements, more efficient implementations than (14) can be devised. Such an equation, in fact, can be interpreted as an infinite impulse response (IIR) digital filter. Well-known techniques to implement digital filters can thus be applied to our problem. In this paper, we adopt the transposed direct form II to implement (14) [8]. Following this approach, (14) is expressed in the following equivalent form:

\[
E_x^{n+1} = \frac{1}{d_0} \left( \frac{d_0 \Delta t}{\Delta t} + \frac{1}{2d_0} \right) \left( \frac{c_0 \Delta t}{\Delta t} - \frac{c_1}{2d_0} \right) E_x^n + \left[ \nabla \times \vec{H} \right]_x^{n+\frac{1}{2}}
+ \frac{d_1 - d_0}{2d_0} J_{lx}^{n+1}, \tag{19}
\]

\[
J_{lx}^{n+1} = \frac{1}{d_0} \left( c_2 E_x^n + E_x^{n+1} - d_1 J_{lx}^{n+1} \right) \tag{20}
\]

where, for this case, coefficients \( c_k \) and \( d_k \) are given by

\[
c_1 = \frac{c_0 \Delta t - 2a_1}{a_0 \Delta t + 2a_1}, \tag{21}
\]

\[
d_0 = \frac{b_0 \Delta t + 2b_1 \Delta y \Delta z}{a_0 \Delta t + 2a_1 \Delta x}, \tag{22}
\]

\[
d_1 = \frac{b_0 \Delta t - 2b_1 \Delta y \Delta z}{a_0 \Delta t + 2a_1 \Delta x}. \tag{23}
\]

For first-order networks, coefficients \( a_k \) and \( b_k \) are quoted in Table I. Note in (19) that, for the particular case \( d_1 = d_0 \) (i.e., \( b_1 = 0 \)), the current density is decoupled to the electric field

\[
E_x^{n+1} = \frac{1}{d_0} \left( \frac{c_0 \Delta t}{\Delta t} + \frac{1}{2d_0} \right) \left( \nabla \times \vec{H} \right)_x^{n+\frac{1}{2}} + \left( \frac{c_0}{\Delta t} - \frac{c_1}{2d_0} \right) E_x^n. \tag{24}
\]

It can be observed in Table I that this simplification applies to three cases: a resistor, a capacitor, and a resistor in parallel with a capacitor. For all three cases, the above equation leads to identical expressions as those given in [2].

#### 2) M-Order RLC Networks (\( M > 1 \)):

For \( RLC \) networks involving more than one element with memory, (6) and (18) apply. Decoupling this set of difference equations and

\[
E_x^{n+1} = \frac{1}{d_0} \left( \frac{c_0 \Delta t}{\Delta t} + \frac{1}{2d_0} \right) \left( \nabla \times \vec{H} \right)_x^{n+\frac{1}{2}} + \left( \frac{c_0}{\Delta t} - \frac{c_1}{2d_0} \right) E_x^n. \tag{24}
\]

1 Other approaches, e.g., an implementation in terms of state variables, would also be convenient as it has been suggested in the context of the extension of the FDTD method to include dispersive media [9].

2 Strictly speaking, a resistor is a zeroth-order lumped network; however, for the sake of brevity, it is included in the first-order lumped network case.
eliminating the auxiliary variable $w_M$, we obtain

$$ E^{n+1}_x = \frac{1}{\frac{\varepsilon_0}{\Delta t} + \frac{1}{2\varepsilon_0}} \left( \frac{\omega}{\Delta t} E^n_x + \nabla \times \vec{H} \right)_{x}^{n+\frac{1}{2}} $$

$$ + \frac{1}{2\Delta x} w^n_x - \frac{1}{2} J^{n+1}_x $$

$$ J^{n+1}_x = \frac{1}{\varepsilon_0} (E^{n+1}_x - w^n_x) $$

$$ w^{n+1}_k = w^{n+1}_k - \alpha_k E^{n+1}_x + d_k J^{n+1}_x $$

$$ w^{n+1}_{M-1} = -c_M E^n_x - c_M E^{n+1}_x + d_M J^n_x + d_{M-1} J^{n+1}_x $$

(25)

where

$$ k = 1, 2, \ldots, M-2. $$

With respect to the standard FDTD method, the implementation of these equations requires $2M+4$ additional real products and $M$ additional storage variables (one for the current density and $M-1$ for the auxiliary variables $w_k$). Therefore, compared with (15) and (16), this implementation requires the same number of real products, but approximately half of the storage variables; just one additional storage variable for each lumped element with memory.

### D. Equivalent-Medium Parameters

From (9), or alternatively from (12), it can be seen that the incorporation of lumped networks into the FDTD scheme is formally equivalent to the inclusion of anisotropic (uniaxial) frequency-dependent materials. More specifically, a lumped network connected to an electrical node of the FDTD mesh is equivalent to the presence of a medium with complex conductivity localized at such a node. The bilinear transformation (11) maps the frequency domain in the $S$-plane ($j\omega$-axis) onto the unit circle of the $Z$-plane. Therefore, the equivalent numerical conductivity of a lumped network can be obtained by simply evaluating (13) on the unit circle of the $Z$-plane, i.e., by letting $z = e^{j\omega \Delta t}$. This leads to the following numerical conductivity:

$$ \hat{\sigma}_{L}(e^{j\omega \Delta t}) = \frac{1 + \sum_{k=1}^{M} c_k \exp(-j\omega \Delta t)}{\sum_{k=0}^{M} d_k \exp(-j\omega \Delta t)} $$

(26)

which, in terms of the coefficients $c_k$ and $d_k$, can be expressed as

$$ \hat{\sigma}_{L}(e^{j\omega \Delta t}) = \frac{\Delta_x}{\Delta_x \Delta_z} \frac{2}{\Delta_z} \tan\left(\frac{\omega \Delta t}{2}\right)^k $$

(27)

It can be observed that the numerical conductivity $\hat{\sigma}_{L}(z = e^{j\omega \Delta t})$ tends to the exact conductivity $\sigma_{L}(s = j\omega)$ as $\omega \Delta t$ tends to zero.

For a general lumped network, the numerical conductivity is a complex function; however, for purely resistive networks, it is real and, for purely reactive networks, it is imaginary.

### III. Validation

To validate this new method, we considered a finite-width parallel-plate waveguide loaded with lumped networks in the transverse plane. The interest in this structure relies on the fact that it allows an analytically exact solution; therefore, it can be considered as a benchmark for validation purposes. We considered the following three different examples of lumped networks:

1) capacitor in series with a resistor (a first-order network);
2) series RLC network (a second-order network);
3) linear equivalent circuit of a Schottky diode HMS-8101 (a fourth-order network).

The waveguide has a width $a = 15$ mm and a height $b = 2$ mm, and has been discretized with an uniform mesh with $\Delta_x = \Delta_y = \Delta_z = 1$ mm. The waveguide is loaded with a uniform plane of $15 \times 2$ identical lumped networks; 15 branches in parallel with two networks in series in each branch. The impedance of the whole plane is, therefore, $2 \Delta Z/15$, where $Z$ is the impedance of each individual lumped network. The results are plotted up to the frequency of 30 GHz, which corresponds to a discretization of ten cells per wavelength. All simulations have been carried out close to the stability limit of the standard FDTD method ($\Delta t \approx 1.926$ ps).

Figs. 1 and 2 show the modulus and phase of the reflection coefficient, respectively, of the parallel-plate waveguide loaded with series $RC$ circuits with $R = 25 \Omega$ and $C = 0.02$ pF. An
As a second example, the same waveguide has been loaded with series \( RLC \) networks with \( R = 75 \Omega \), \( L = 3 \text{ nH} \), and \( C = 0.08 \text{ pF} \). The computed results are plotted in Figs. 3 and 4, again showing an excellent agreement with those computed analytically.

Finally, as an example of more practical interest, we have considered the linear equivalent circuit of a Schottky diode, shown in Fig. 5. An impedance \( Z = 15Z_d/2 \), where \( Z_d \) is the input impedance of the diode, has been connected to each node of the transverse plane in such a way that the overall impedance is equal to \( Z_d \). The results calculated for this case are plotted in Figs. 6 and 7. When the results computed by FDTD are compared with those obtained analytically, we observe an excellent agreement for both the modulus and phase of the reflection coefficient up to the frequency of 20 GHz. In the range from 20 to 30 GHz, a small discrepancy is observed,
which can be ascribed to the numerical dispersion error that increases with the order of the lumped network.

IV. CONCLUSION

The LE-FDTD method has been extended to include arbitrary linear RLC lumped networks. To this purpose, a lumped network is characterized by its V–I relation in the Laplace domain. This relation is incorporated into the FDTD method in a simple and systematic way by applying to it the bilinear transformation. The resulting relation in the Z-domain is expressed as a high-order difference equation and efficiently implemented by using techniques commonly employed to implement IIR digital filters. The whole procedure to include lumped networks into the FDTD method constitutes a general technique that can be applied to incorporate other frequency-dependent quantities like dispersive dielectric constants or surface impedance boundary conditions into the FDTD method.

By using this formulation, high-order RLC lumped networks can be incorporated into the FDTD method using only one cell, therefore, preserving the lumped nature of the network. On the other hand, if desired, homogeneously distributed models of lumped networks can also be built. To this purpose, the same lumped network of interest is connected to all electric nodes over an extended region of the FDTD mesh.

Eventually, this approach allows an interesting interpretation of the extension of the FDTD method to lumped elements. This is formally equivalent to the incorporation of anisotropic (uniaxial) frequency-dependent media. This equivalence suggests that FDTD codes capable to incorporate frequency dependent media can be easily adapted to include lumped networks and vice versa.

APPENDIX

The coefficients $c_k$ and $d_k$ of the equivalent conductivity (13) can be computed by using the following expressions:

$$c_k = \frac{1}{\alpha_0} \sum_{m=0}^{M} d_m \sum_{\ell=\ell_{\text{min}}}^{\ell_{\text{max}}} p_{m, \ell} q_{M-m, \ell} \quad k = 1, \ldots, M$$

(29)

$$d_k = \frac{\Delta y \Delta z}{\alpha_0 \Delta x} \sum_{m=0}^{M} h'_m \sum_{\ell=\ell_{\text{min}}}^{\ell_{\text{max}}} p_{m, \ell} q_{M-m, \ell} \quad k = 0, \ldots, M$$

(30)

where $\ell_{\text{min}} = \max(0, k + m - M)$, $\ell_{\text{max}} = \min(k, m)$, and $\alpha_0$ is a normalization constant given by

$$\alpha_0 = \sum_{m=0}^{M} d_m q_{M-m, 0}$$

(31)

and the coefficients $d_m$ and $h'_m$ by

$$d_m = \frac{a_m}{2M-m \Delta p_i}$$

(32)

$$h'_m = \frac{b_m}{2M-m \Delta p_i}$$

(33)

The quantities $p_{m, \ell}$ and $q_{m, \ell}$ denote the coefficients of the polynomials $P_m(z)$ and $Q_m(z)$, respectively. These polynomials are given by

$$P_m(z) = \left(1 - z^{-1}\right)^m = \sum_{\ell=0}^{m} p_{m, \ell} z^{-\ell}$$

(34)

$$Q_m(z) = \left(1 + z^{-1}\right)^m = \sum_{\ell=0}^{m} q_{m, \ell} z^{-\ell}$$

(35)

so that the coefficients $p_{m, \ell}$ and $q_{m, \ell}$ are identified as

$$p_{m, \ell} = (-1)^{\ell} \frac{m!}{\ell!}$$

(36)

$$q_{m, \ell} = \frac{m!}{\ell!}$$

(37)

REFERENCES


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Roberto Sorrentino (M’77–SM’84–F’90), for photograph and biography, see this issue, p. 815.