Finite-Difference Frequency-Domain Algorithm for Modeling Guided-Wave Properties of Substrate Integrated Waveguide

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***Abstract*—In multilayer microwave integrated circuits such as low-temperature co-fired ceramics or multilayered printed circuit boards, waveguide-like structures can be fabricated by using periodic metallic via-holes referred to as substrate integrated waveguide (SIW). Such SIW structures can largely preserve the advantages of conventional rectangular waveguides such as high-Q factor and high power capacity. However, they are subject to leakage due to periodic gaps, which potentially results in wave attenuation. Therefore, such a guided-wave modeling problem becomes a very complicated complex eigenvalue problem. Since the SIW are bilaterally unbounded, absorbing boundary conditions should be deployed in numerical algorithms. This often leads to a difficult complex root-extracting problem of a transcend equation. In this paper, we present a novel finite-difference frequency-domain algorithm with a perfectly matched layer and Floquet's theorem for the analysis of SIW guided-wave problems. In this scheme, the problem is converted into a generalized matrix eigenvalue problem and finally transformed to a standard matrix eigenvalue problem that can be solved with efficient subroutines available. This approach has been validated by experiment.**

***Index Terms*—Eigenvalue problem, finite difference frequency domain (FDFD), open periodic structure, perfectly matched layer (PML), substrate integrated waveguide (SIW).**

I. Introduction

A MICROWAVE system generally requires the use of many different technologies to achieve the preplanned performance although integrated planar circuits such as monolithic microwave integrated circuits (MMICs) are still the mainstream building blocks. Active devices in the form of chips are often surface mounted on a planar carrier substrate while high-Q passive components such as a diplexer and filter are usually designed on the basis of rectangular waveguide or other nonplanar technologies. The concept of integrated rectangular waveguide has recently been studied and developed [1]–[4]. The idea of synthesizing nonradiative dielectric (NRD) guides and other nonplanar guided-wave structures in planar form on a single substrate leads to the design and development of low-cost millimeter-wave integrated circuits (ICs) and systems [5]. In this way, a system can be integrated even in a package [i.e., system on package (SOP)], reducing size, weight, and cost, and greatly enhancing manufacturing repeatability and reliability. Fig. 1 shows the structure of a substrate integrated waveguide (SIW) that is synthesized with linear arrays of metallic via-holes on a low-loss substrate [e.g., low-temperature co-fired ceramic (LTCC)].

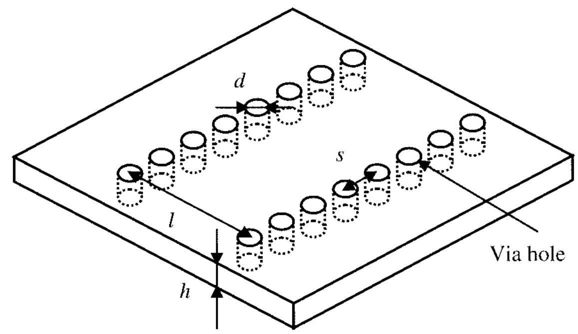

Fig. 1. Configuration of SIW synthesized using metallic via-hole arrays.

Since the bilateral walls of such a periodic waveguide formed are laterally open, absorbing boundary conditions (ABCs) should be used in numerical algorithms for modeling this kind of structure. This often results in a difficult complex root-extracting problem of a transcend equation, which is also a very tedious procedure. Accurate prediction of propagation characteristics of such periodic structures is essential in any successful design of SIW-based circuits and systems. A variety of numerical techniques have been used to analyze periodic structure problems. Basically, there are two groups of numerical technique, i.e., 1) methods that determine the propagation constants of Floquet modes from a transcend equation [6], [7] and 2) methods that determine the propagation constant of Floquet modes on the basis of classical eigenvalues of a matrix [8]–[11].

The finite-difference frequency-domain (FDFD) method has advantages to handle periodic structures with complicated geometries and anisotropy. When the existing FDFD method is

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used to calculate the propagation constant, roots of the eigenvalue equation are related to frequencies for a given value of the propagation constant [8], [11]. As a result, the existing FDFD method is not suitable for modeling complex propagation constant problems, particularly for the analysis of periodic structures [12], [13].

In this paper, we propose a FDFD algorithm that can model periodic structures with complicated geometries and anisotropy. By the use of Floquet's theorem for periodic structures on electric and magnetic field boundaries, the computational domain is restricted to a single period. The open periodic propagation problem can be solved by introduce a perfectly matched layer (PML) ABC, which yields an eigenvalue problem after eliminating the longitudinal field components. Different from the existing finite-difference (FD) methods, the roots of an eigenvalue equation are directly the propagation constants for a given value of frequency. With this method, open SIW periodic structures have been modeled, and a complex eigenvalue of the propagation problem has been obtained. Numerical results are compared with measurements in the frequency range of 26.5–40 GHz.

II. Formulation of the Eigenvalue-Based FDFD

According to the Floquet's theorem for periodic structures [14], electric and magnetic fields for the periodic guided waves can be expressed as

$$G(x,y,z) = g(x,y,z)e^{-\gamma z} \tag{1}$$

where γ is the propagation constant $(\gamma = \alpha + j\beta), g(x,y,z)$ is a periodic function with respect to z and represents the periodic functions of electric field $e(x,y,z)$, or magnetic field $h(x,y,z), G(x,y,z)$ represents electric field $E(x,y,z)$ or magnetic field $H(x,y,z)$. To simulate the open periodic structures in the x- and y-directions, a PML ABC is introduced. Maxwell's curl equations can be written in a split form [15], where each component of electromagnetic fields in split into two parts in the PML medium. In Cartesian coordinates, the six field components yield 12 subcomponents. Although the proposed approach is applicable to a general anisotropic type of periodic guided structures, the following derivation will be limited to the case of electric anisotropic media with diagonal dielectric constant tensor for the simplicity of illustration. Substituting (1) into Maxwell equations with a split form and replacing the derivative with respect to t by $j\omega$ yields

$$j\omega\mu_0 h_{xy} + \sigma_y^*(x)h_{xy} = -\frac{\partial(e_{zx}+e_{zy})}{\partial y} \tag{2a}$$

$$j\omega\mu_0 h_{xz} = \frac{\partial(e_{yz}+e_{yx})}{\partial z} - \gamma(e_{yz}+e_{yx}) \tag{2b}$$

$$j\omega\mu_0 h_{yz} = -\frac{\partial(e_{xy}+e_{xz})}{\partial z} + \gamma(e_{xy}+e_{xz}) \tag{2c}$$

$$j\omega\mu_0 h_{yx} + \sigma_x^*(y)h_{yx} = \frac{\partial(e_{zx}+e_{zy})}{\partial x} \tag{2d}$$

$$j\omega\mu_0 h_{zx} + \sigma_x^*(z)h_{zx} = -\frac{\partial(e_{yz}+e_{yx})}{\partial x} \tag{2e}$$

$$j\omega\mu_0 h_{zy} + \sigma_y^*(z)h_{zy} = \frac{\partial(e_{xy}+e_{xz})}{\partial y} \tag{2f}$$

$$j\omega\varepsilon_x e_{xy} + \sigma_y(x)e_{xy} = \frac{\partial(h_{zx}+h_{zy})}{\partial y} \tag{2g}$$

$$j\omega\varepsilon_x e_{xz} + \sigma_z(x)e_{xz} = -\frac{\partial(h_{yz}+h_{yx})}{\partial z} + \gamma(h_{yz}+h_{yx}) \tag{2h}$$

$$j\omega\varepsilon_y e_{yz} + \sigma_z(y)e_{yz} = \frac{\partial(h_{xy}+h_{xz})}{\partial z} - \gamma(h_{xy}+h_{xz}) \tag{2i}$$

$$j\omega\varepsilon_y e_{yx} + \sigma_x(y)e_{yx} = -\frac{\partial(h_{zx}+h_{zy})}{\partial x} \tag{2j}$$

$$j\omega\varepsilon_z e_{zx} + \sigma_x(z)e_{zx} = \frac{\partial(h_{yz}+h_{yx})}{\partial x} \tag{2k}$$

$$j\omega\varepsilon_z e_{zy} + \sigma_y(z)e_{zy} = -\frac{\partial(h_{xy}+h_{xz})}{\partial y} \tag{2l}$$

where σ and σ^* denote the electric conductivity and magnetic loss, respectively, and they satisfy the impedance matching condition [16]. Next, we combine the electromagnetic fields with a split form and decrease the number of variables. Furthermore, we find that if the longitudinal components (e_z and h_z) are eliminated, the guided-wave problem can be transformed into a generalized eigenvalues problem. This step not only decreases the number of variables further, but also results in an easily solved eigenvalue problem. From (1) and (2), we can obtain (see the Appendix)

$$\begin{aligned}\gamma h_x = {}& \frac{\partial h_x}{\partial z} - \frac{j\omega\varepsilon_y+\sigma_z(y)}{j\omega\varepsilon_y+\sigma_x(y)}\cdot\frac{1}{j\omega\mu_0+\sigma_y^*}\cdot\frac{\partial^2 e_x}{\partial x\partial y}\\ & - \big(j\omega\varepsilon_y+\sigma_z(y)\big)e_y + \frac{j\omega\varepsilon_y+\sigma_z(y)}{j\omega\varepsilon_y+\sigma_x(y)}\\ & \cdot\frac{1}{j\omega\mu_0+\sigma_x^*}\cdot\frac{\partial^2 e_y}{\partial x^2}\end{aligned} \tag{3a}$$

$$\begin{aligned}\gamma h_y = {}& \frac{\partial h_y}{\partial z} + \big(j\omega\varepsilon_x+\sigma_z(x)\big)e_x - \frac{j\omega\varepsilon_x+\sigma_z(x)}{j\omega\varepsilon_x+\sigma_y(x)}\\ & \cdot\frac{1}{j\omega\mu_0+\sigma_x^*}\cdot\frac{\partial^2 e_x}{\partial y^2} + \frac{j\omega\varepsilon_x+\sigma_z(x)}{j\omega\varepsilon_x+\sigma_y(x)}\\ & \cdot\frac{1}{j\omega\mu_0+\sigma_x^*}\cdot\frac{\partial^2 e_y}{\partial x\partial y}\end{aligned} \tag{3b}$$

$$\begin{aligned}\gamma e_x = {}& \frac{j\omega\mu_0}{j\omega\mu_0+\sigma_x^*}\cdot\frac{1}{j\omega\varepsilon_z+\sigma_y(z)}\cdot\frac{\partial^2 h_x}{\partial x\partial y} + j\omega\mu_0 h_y\\ & - \frac{j\omega\mu_0}{j\omega\mu_0+\sigma_x^*}\cdot\frac{1}{j\omega\varepsilon_z+\sigma_x(z)}\cdot\frac{\partial^2 h_y}{\partial x^2} + \frac{\partial e_x}{\partial z}\end{aligned} \tag{3c}$$

$$\begin{aligned}\gamma e_y = {}& -j\omega\mu_0 h_x + \frac{j\omega\mu_0}{j\omega\mu_0+\sigma_y^*}\cdot\frac{1}{j\omega\varepsilon_z+\sigma_y(z)}\cdot\frac{\partial^2 h_x}{\partial y^2}\\ & - \frac{j\omega\mu_0}{j\omega\mu_0+\sigma_y^*}\cdot\frac{1}{j\omega\varepsilon_z+\sigma_x(z)}\cdot\frac{\partial^2 h_y}{\partial x\partial y} + \frac{\partial e_y}{\partial z}.\end{aligned} \tag{3d}$$

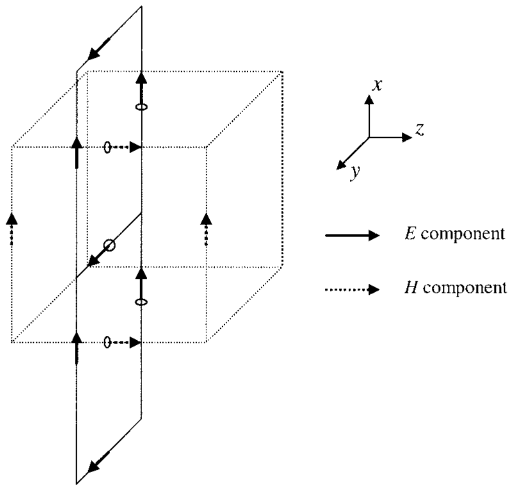

Fig. 2. Yee's three-dimensional lattice.

As shown in Fig. 2, if uniform meshes are used in the periodic direction, the difference equation of (3a) is

$$\begin{aligned}
&\gamma \frac{h_x\left(i,j,k-\frac{1}{2}\right)+h_x\left(i,j,k+\frac{1}{2}\right)}{2} \\
&\quad= \frac{1}{\Delta z}\left[h_x\left(i,j,k+\frac{1}{2}\right)-h_x\left(i,j,k-\frac{1}{2}\right)\right] \\
&\quad - \frac{1}{\Delta x \Delta y}\cdot\frac{j\omega\varepsilon_y(i,j,k)+\sigma_z^y(i,j,k)}{j\omega\varepsilon_y(i,j,k)+\sigma_z^x(i,j,k)} \\
&\quad \cdot \frac{1}{j\omega\mu_0+\sigma_y^*\left(i+\frac{1}{2},j,k\right)} \\
&\quad \cdot \left[e_x\left(i+\frac{1}{2},j+\frac{1}{2},k\right)-e_x\left(i+\frac{1}{2},j-\frac{1}{2},k\right)\right] \\
&\quad + \frac{1}{\Delta x \Delta y}\cdot\frac{j\omega\varepsilon_y(i,j,k)+\sigma_z^y(i,j,k)}{j\omega\varepsilon_y(i,j,k)+\sigma_z^x(i,j,k)} \\
&\quad \cdot \frac{1}{j\omega\mu_0+\sigma_y^*\left(i-\frac{1}{2},j,k\right)} \\
&\quad \cdot \left[e_x\left(i-\frac{1}{2},j+\frac{1}{2},k\right)-e_x\left(i-\frac{1}{2},j-\frac{1}{2},k\right)\right] \\
&\quad - \left[j\omega\varepsilon_y(i,j,k)+\sigma_z^y(i,j,k)\right] \\
&\quad \cdot e_y(i,j,k) + \frac{1}{\Delta x \Delta x}\cdot\frac{j\omega\varepsilon_y(i,j,k)+\sigma_z^y(i,j,k)}{j\omega\varepsilon_y(i,j,k)+\sigma_z^x(i,j,k)} \\
&\quad \cdot \frac{1}{j\omega\mu_0+\sigma_x^*\left(i+\frac{1}{2},j,k\right)} \\
&\quad \cdot \left[e_y(i+1,j,k)-e_y(i,j,k)\right] - \frac{1}{\Delta x \Delta x} \\
&\quad \cdot \frac{j\omega\varepsilon_y(i,j,k)+\sigma_z^y(i,j,k)}{j\omega\varepsilon_y(i,j,k)+\sigma_z^x(i,j,k)} \\
&\quad \cdot \frac{1}{j\omega\mu_0+\sigma_x^*\left(i-\frac{1}{2},j,k\right)} \\
&\quad \cdot \left[e_y(i,j,k)-e_y(i-1,j,k)\right]. \qquad (4)
\end{aligned}$$

Since the discrete node is on the e_y component, the left-sided h_x component of the equation is obtained by taking the average of two-sided h_x components under the condition of uniform longitudinal meshes. The corresponding difference equations of the remained equations (3b)–(d) are similar to (4). As shown in Fig. 3, according to Floquet's theorem, e_0 equals e_p and h_1 equals h_{p+1} if the periodic length is discretized into p segments along the z-direction; the computational domain is then restricted to a single period of the structure. When all the boundary conditions are applied, the guided-wave problem can be converted into a generalized eigenvalue problem as follows:

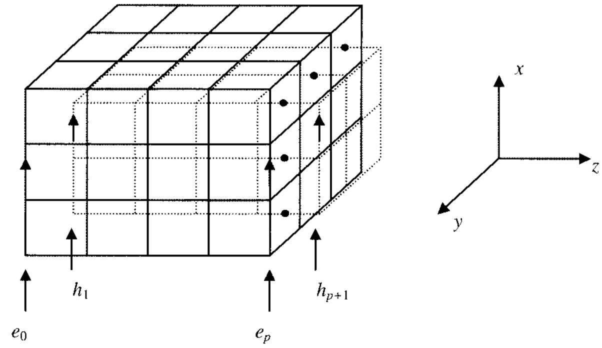

Fig. 3. Yee meshes of single period.

$$\gamma \mathbf{B}\begin{bmatrix}\mathbf{h_x}\\ \mathbf{h_y}\\ \mathbf{e_x}\\ \mathbf{e_y}\end{bmatrix} = \mathbf{A}\begin{bmatrix}\mathbf{h_x}\\ \mathbf{h_y}\\ \mathbf{e_x}\\ \mathbf{e_y}\end{bmatrix}. \qquad (5)$$

From (4), it is obvious that matrix $\mathbf{B}$ is a block diagonal matrix

$$\mathbf{B} = \begin{bmatrix}\mathbf{B_c} & 0 & \cdots & 0\\ 0 & \ddots & \ddots & \vdots\\ \vdots & \ddots & \ddots & 0\\ 0 & \cdots & 0 & \mathbf{B_c}\end{bmatrix} \qquad (6)$$

where $\mathbf{B_c}$ is a square matrix with p rows, and its inverse matrix can be obtained analytically if the number p of longitudinal nodes is odd. The resulting inverse matrix has a Toeplitz form with the first row written as

$$\mathbf{b_c}\left(\mathbf{B_c^{-1}}\right) = [1 \quad -1 \quad 1 \quad -1 \quad \cdots \quad 1 \quad -1 \quad 1]. \qquad (7)$$

Finally, the generalized eigenvalue problem (5) can be simplified to a standard matrix eigenvalue problem as

$$\mathbf{Cx} = \gamma \mathbf{x} \qquad (8)$$

where $\mathbf{C} = \mathbf{B}^{-1}\mathbf{A}$. Hence, we always divide the longitudinal period into odd segments.

Due to the similarity between finite-difference time-domain (FDTD) and FDFD methods, many well-established algorithms for the FDTD method can be used in the eigenvalue-based FDFD method to enhance efficiency and accuracy.

A. Nonuniform Orthogonal Grids

The modeling accuracy of the SIW guided-wave problem can be further improved by using an orthogonal nonuniform grid. This allows the geometry to be modeled more accurately than in the case of a uniform scheme. To apply this technique, the growth (or scaling) factor of the mesh, i.e., the ratio of spatial steps of two adjacent cells, should be kept approximately below 1.2–1.3 to prevent artificial field discontinuities due to the abrupt change in cell size [17]. Hence, we make use of the same $\mathbf{B}^{-1}$ to simplify the problem. However, the nonuniform

TABLE I
EIGENVALUE COMPARISON OF CALCULATED SIW PROPAGATION CONSTANTS BETWEEN LONGITUDINAL SEGMENTS $P = 5$ AND $P = 7$ ($f = 12$ GHz)

Complex propagation constant ($\gamma = \alpha + j\beta$)	
Longitudinal segments $p = 5$	Longitudinal segments $p = 7$
1.1559e-002 + 6.4005e+002i	1.1612e-002 + 6.4005e+002i
6.8991e-003 - 2.8435e+003i	4.8100e-002 + 2.6623e+003i
3.5579e-003 + 4.4201e+003i	6.1284e-002 - 4.0821e+003i
5.7656e-003 - 1.3341e+004i	7.5643e-002 +7.7719e+003i
5.9631e-003 + 1.7433e+004i	1.2206e-002 +2.7914e+004i

grid can no longer preserve the second-order accuracy obtained by the uniform grid. If the mesh spacing changes slowly, the resulting error looks more similar to that of a second-order method than that of a first-order counterpart.

B. Locally Conformed FD Technique

The locally conformed FD technique adequately makes use of the integral form of Maxwell's equations for modeling arbitrarily shaped curved surfaces [18]. We can also apply such a technique to the FDFD method to improve the simulation accuracy in connection with via-holes.

Since the number of eigenvalues is equal to the rank of matrix $\mathbf{C}$ after solving (8), it is important to extract real solutions and get rid of pseudosolutions. The most efficient method we have found is to solve the eigenvalue problem twice with different mesh sizes and compare the two groups of obtained eigenvalues. If an eigenvalue remains the same in the two cases, it is a real solution, otherwise it will be a pseudosolution. The correctness of this rule has been shown by numerical experiments. Table I lists five eigenvalues of matrix $\mathbf{C}$ when the longitudinal segments number p is 5 and 7, respectively. The eigenvalue problem is related to the propagation constant of the periodic SIW, where $f = 12$ GHz, $l = 7.112$ mm, $h = 2$ mm, $s = 2.0$ mm, $d = 0.8$ mm, and the relative permittivity of dielectric $\epsilon_r = 10.2$, as shown in Fig. 1. Real parts of those selected eigenvalues are mostly close to zero because we mainly concern the guided-wave problem with less leakage and loss. From Table I and following the rule, $\gamma = 1.16\text{e} - 02 + j6.40\text{e} + 02$ is obviously the solution we search for. When the range of the imaginary part of the propagation constant is defined, however, it is not necessary to solve the problem twice for extracting the real eigenvalues at different frequencies.

III. SIMULATION AND MEASUREMENT RESULTS

A. Simulation

The structure under consideration is a periodic SIW, as shown in Fig. 1, where only a periodic cell is considered. The other walls are either perfect electric conductors (PECs) (upper and bottom planes) or PML absorbing boundaries (lateral). Simulation and measurement results show that the attenuation constant of the SIW is rather small. Note that the attenuation constant is too small to be measured accurately. We make use of HFSS software to solve the same problem for the validation of the proposed FDFD method.

The geometry parameters of the SIW in Fig. 1 are set as $l = 7.112$ mm, $h = 2$ mm, and $d = 0.8$ mm, the relative permittivity ϵ_r is 10.2, and frequency is $f = 12$ GHz. When the via-spacing s changes from 2.0 to 4.0 mm, the insertion loss and phase constant of the SIW have been calculated using the HFSS software, where 11 periods are considered in the calculations. The comparison of the attenuation constant obtained from the FDFD method and HFSS is illustrated in Fig. 4. Clearly, simulation results from the two methods are in very good agreement. Numerical results of the phase constant obtained by the two techniques are shown in Fig. 5, which also illustrates a good agreement. When applying HFSS to simulate an SIW, we calculate the propagation constant as a determinate problem with an excitation, and the higher order modes resulting from port discontinuity cause the additional loss. As a result, the insertion loss based on HFSS is slightly larger than that from the proposed FDFD method. The distribution of the vector electric field drawn by HFSS is shown in Fig. 6. It is observed that the energy is mainly restricted between the two rows of holes. With the proposed FDFD method, the typical computation time is only 5 to 7 min, while the CPU time for HFSS simulation is usually over 30 min. In the meantime, the memory requirement in HFSS is much more than the proposed technique.

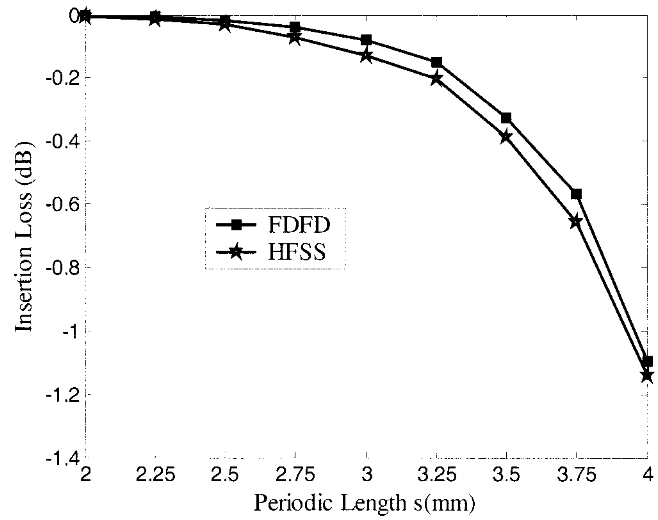

Fig. 4. Comparison between insertion-loss results obtained from the FDFD method and the HFSS software with 11 periods.

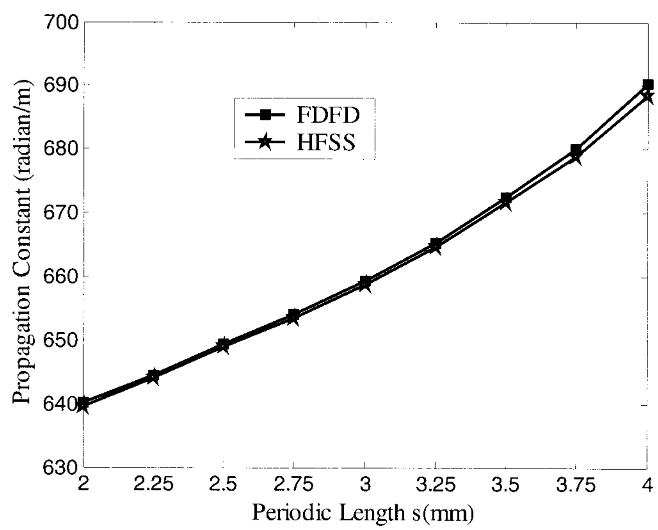

Fig. 5. Comparison between the phase-constant results obtained from the FDFD method and the HFSS software with 11 periods.

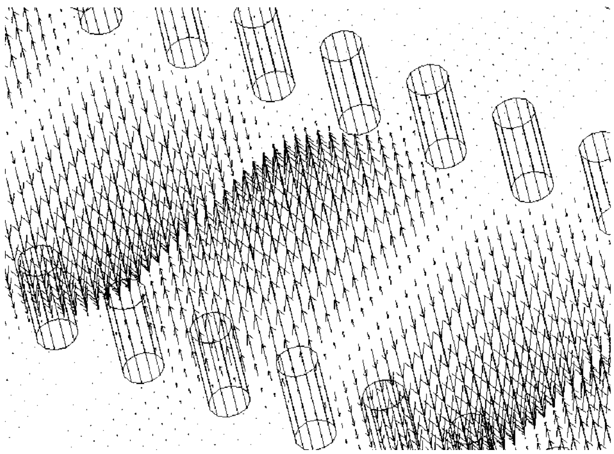

Fig. 6. Distribution diagram for the electric-field vector (from HFSS), where $f = 12$ GHz, $s = 2$ mm, $d = 0.8$ mm, $h = 2$ mm, and $l = 7.112$ mm.

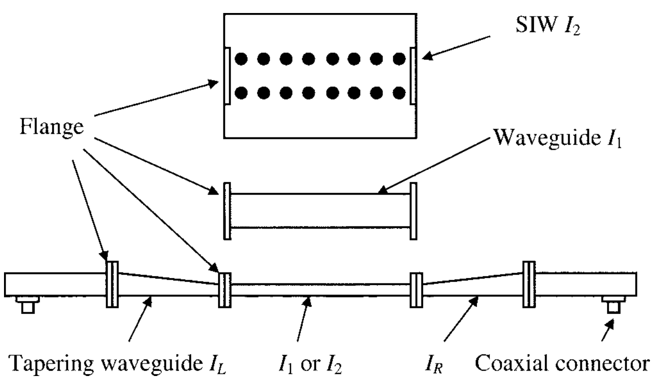

Fig. 7. SIW experimental configuration.

B. Experiment

In the measurement setup, the dimensions of the SIW geometry are chosen as $l = 7.112$ mm, $d = 0.8$ mm, $s = 1.5$ mm, and $h = 2$ mm, and the relative permittivity of dielectric substrate is 10.2, as shown in Figs. 1 and 7. In Fig. 7, parts I_L and I_R are taper rectangular waveguides in highness, which act as the adapters between the SIW and Ka-band standard waveguide of size 7.112 mm × 3.556 mm, and the middle part is denoted as I_1 or I_2. Part I_1 or Part I_2 are connected with Parts I_L and I_R by the mounting flanges, as shown in Fig. 7.

Part I_1 SIW, synthesized with linear arrays of metallized via-holes on a low-loss substrate covered by metallized walls, as shown in Fig. 1.

Part I_2 Rectangular waveguide filled with the same medium as the SIW substrate. Its basic geometry parameters are 7.112 mm × 2 mm.

The experiments are performed in the frequency range of 26.5–40 GHz. Since Part I_2 is a conventional rectangular waveguide with a reduced height, the inherent phase values of the experimental system at various frequencies can be obtained by measuring the scattering S parameters of Part I_2. In the measurement, we first insert Part I_2 between Parts I_L and I_R, and the S-parameters are measured to calibrate the system error. Then insert Part I_1 to measure the S-parameters of the SIW. Since the attenuation constant is very small, only phase constant β, i.e., the imaginary part of the propagation constant γ, can be derived from the measured S-parameter data.

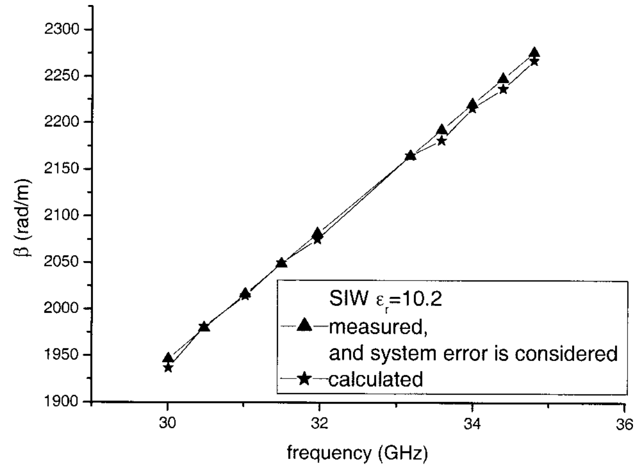

Fig. 8. Simulation results of the FDFD method and measurement results of the SIW after calibration.

The SIW is measured at ten frequencies, and the comparison of measurement results with numerical simulations from the proposed FDFD method is shown in Fig. 8. It can be seen that the numerical results are in good agreement with the experiment results.

IV. Conclusion

An FDFD algorithm has been presented for the efficient modeling of guided-wave properties of an SIW. Simulation and measurement results have validated the proposed numerical modeling scheme, which is suitable to solve arbitrary complex open periodic guided-wave problems. Both simulation and experimental results have shown that the performance of the SIW is very similar to that of a conventional waveguide (such as with a high-Q factor). The concepts of substrate integrated circuits (ISCs) including an SIW can be anticipated for the future design and development of low-cost millimeter-wave ICs and systems.

Appendix

From (2e) and (2f), we obtain

$$h_{zx} = -\frac{1}{j\omega\mu_0 + \sigma_x^*} \cdot \frac{\partial e_y}{\partial x} \tag{A1}$$

$$h_{zy} = \frac{1}{j\omega\mu_0 + \sigma_y^*} \cdot \frac{\partial e_x}{\partial y}. \tag{A2}$$

Combining (A1) and (A2) yields

$$h_z = \frac{1}{j\omega\mu_0 + \sigma_y^*} \cdot \frac{\partial e_x}{\partial y} - \frac{1}{j\omega\mu_0 + \sigma_x^*} \cdot \frac{\partial e_y}{\partial x}. \tag{A3}$$

Substituting (A3) into (2j) and (2g), we have

$$e_{yx} = -\frac{1}{j\omega\varepsilon_y + \sigma_x(y)} \cdot \left(\frac{1}{j\omega\mu_0 + \sigma_y^*} \cdot \frac{\partial^2 e_x}{\partial x \partial y} - \frac{1}{j\omega\mu_0 + \sigma_x^*} \cdot \frac{\partial^2 e_y}{\partial x^2}\right) \tag{A4}$$

$$e_{xy} = \frac{1}{j\omega e_x + \sigma_y(x)} \cdot \left(\frac{1}{j\omega\mu_0 + \sigma_y^*} \cdot \frac{\partial^2 e_x}{\partial y^2} - \frac{1}{j\omega\mu_0 + \sigma_x^*} \cdot \frac{\partial^2 e_y}{\partial x \partial y}\right). \tag{A5}$$

From (A4) and (2i), one obtains (3a). Similarly, from (A5) and (2h), we can get (3b).

Simplifying (2k) and (2l), we have

$$e_{zx} = \frac{1}{j\omega\varepsilon_z + \sigma_x(z)} \cdot \frac{\partial h_y}{\partial x} \tag{A6}$$

$$e_{zy} = -\frac{1}{j\omega\varepsilon_z + \sigma_y(z)} \cdot \frac{\partial h_x}{\partial y}. \tag{A7}$$

Combining the two equations yields

$$e_z = -\frac{1}{j\omega\varepsilon_z + \sigma_y(z)} \cdot \frac{\partial h_x}{\partial y} + \frac{1}{j\omega\varepsilon_z + \sigma_x(z)} \cdot \frac{\partial h_y}{\partial x}. \tag{A8}$$

Substituting (A8) into (2d) and (2a), we have

$$h_{yx} = \frac{1}{j\omega\mu_0 + \sigma_x^*} \cdot \left(-\frac{1}{j\omega\varepsilon_z + \sigma_y(z)} \frac{\partial^2 h_x}{\partial x \partial y} + \frac{1}{j\omega\varepsilon_z + \sigma_x(z)} \cdot \frac{\partial^2 h_y}{\partial x^2} \right) \tag{A9}$$

$$h_{xy} = -\frac{1}{j\omega\mu_0 + \sigma_y^*} \cdot \left(-\frac{1}{j\omega\varepsilon_z + \sigma_y(z)} \frac{\partial^2 h_x}{\partial y^2} + \frac{1}{j\omega\varepsilon_z + \sigma_x(z)} \cdot \frac{\partial^2 h_y}{\partial x \partial y} \right). \tag{A10}$$

From (A9) and (2c), (3c) can be obtained. Similarly, from (A10) and (2b), (3d) can be obtained.

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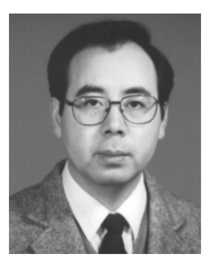

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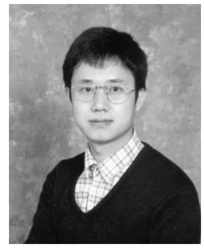

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