Efficient Calculation of the Dispersion Diagram of Planar Electromagnetic Band-Gap Structures by the MoM/BI-RME Method

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***Abstract*—The characterization of planar electromagnetic band-gap structures requires the calculation of the dispersion diagram of the modes supported by the periodic structure and the phase of the reflection coefficient under plane-wave illumination. In this paper, we present a novel method for the calculation of the dispersion diagram. The electromagnetic analysis is based on the method of moments/boundary integral-resonant mode expansion (MoM/BI-RME) method and leads to the formulation of a homogeneous matrix problem. The solution of this problem is performed by an iterative procedure: for a given value of the propagation phase constant, the frequency range is scanned to find the frequencies where the field equation has a nontrivial solution. The search of these frequencies is based on the tracking of the eigenvalues in the complex plane, and proved more efficient than other classical methods (direct search of the determinant zeros, singular value decomposition). The reflection coefficient can be readily determined by using the MoM/BI-RME method, already developed for the analysis of the scattering from frequency selective surfaces. The method is applied to the characterization of the classical uniplanar compact photonic-bandgap structure, and analysis results show the accuracy of the method, its efficiency, and its convergence properties.**

***Index Terms*—Eigenvalues and eigenfunctions, electromagnetic band-gap structures, integral equations, periodic structures.**

I. Introduction

ELECTROMAGNETIC band-gap (EBG) structures are widely used for improving the electromagnetic performance in microwave circuits and antennas [1]–[4]. Structures operating in the microwave range commonly consist of periodic arrays of metal patches patterned on a grounded dielectric substrate, possibly connected to the ground plane through metal pins. By properly choosing their geometrical properties, these structures exhibit two different kinds of performance. On the one hand, they can be used to prevent the propagation of substrate waves in a frequency band (EBG behavior), thus reducing the leakage in guided-wave circuits [1] and improving the efficiency of patch antennas [2]. On the other hand, they can behave as an artificial perfect magnetic conductor (PMC), reflecting an incident plane wave without phase reversal. This permits the design of transverse electromagnetic TEM waveguides [3] and low-profile patch antennas [4].

Therefore, the analysis of electromagnetic band-gap structures should ideally provide two kinds of information: the dispersion diagram (DD) of the modes supported by the periodic structure and the phase of the reflection coefficient of the structure under plane-wave illumination. However, in many practical cases [4], the most useful information is the frequency range where the propagation of electromagnetic waves is forbidden (bandgap). This information can be retrieved by considering only the boundary of the irriducible Brillouin zone [5], [6].

The phase of the reflection coefficient can be calculated like the scattering from frequency selective surfaces (FSSs). Different numerical methods and hybrid techniques have been proposed: the integral equation method (IEM) [7], the finite element method (FEM) [8], the finite difference time domain (FDTD) method [9], the method of moments/boundary integral-resonant mode expansion (MoM/BI-RME) method [10], [11], and the finite element/boundary integral (FE/BI) method [12].

Most of the techniques used for the analysis of FSSs can be adapted to the determination of the DD of EBG structures. Anyway, there is a substantial difference between the two analyzes: in the calculation of the plane-wave response of FSSs, the fields inside the structure are generated by an incident plane wave, whereas in the determination of the DD there is no incident field, and the analysis leads to a homogeneous problem. When using the FEM in conjunction with the perfectly matched layer (PML), it is possible to determine all mode frequencies through the solution of a single eigenvalue problem for each value of the propagation constant [13]. However, it requires a volume mesh, thus typically resulting in large matrix problems. When using the FDTD method, time iterations are required, until the steady-state is reached. The Fourier transform of the steady-state solution yields a peak at the frequency of each eigenmode [14]. The FEM and the FDTD method provide flexibility in the geometry of the components, but can be time consuming. Conversely, when using the IEM, the matrix of the resulting homogeneous problem depends on both the propagation constant and the frequency, and it is not possible to separate the two dependencies [15]. In this case, for a given value of the propagation constant, a frequency iteration is needed to find the nontrivial solutions of the problem and the resulting points

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of the dispersion diagram. The most common iterative strategies are based either on the search of the determinant zeros of the matrix [15] or on the search of its minimum singular value [16]. However, these approaches require a fine frequency scan, as discussed in Section III. Moreover, the identification of multiple zeros can be critical, and the human intervention can be required for distinguishing real zeros from local minima.

In this paper, we present a novel approach for the efficient characterization of planar metallo-dielectric EBG structures, consisting of a periodic array of either isolated or interconnected metal patches, printed on a dielectric substrate (Fig. 1). In the calculation of the DD, the electromagnetic analysis is based on the MoM/BI-RME method, which is described in Sections II-A and B. The novelty of the proposed approach is the particular technique used for the computation of the dispersion diagram, which is based on the tracking of the eigenvalues in the complex plane. This approach, discussed in Section II-C, is very reliable and allows for a coarse frequency scanning, thus leading to a limited number of electromagnetic full-wave analysis. As far as the phase of the reflection coefficient is concerned, the MoM/BI-RME method described in [10] and [11] is applied. The proposed method is applied to the analysis of the classical uniplanar compact photonic-bandgap (UC-PBG) structure [1]. Some results reported in Section III show the accuracy of the method, its efficiency, and its convergence properties.

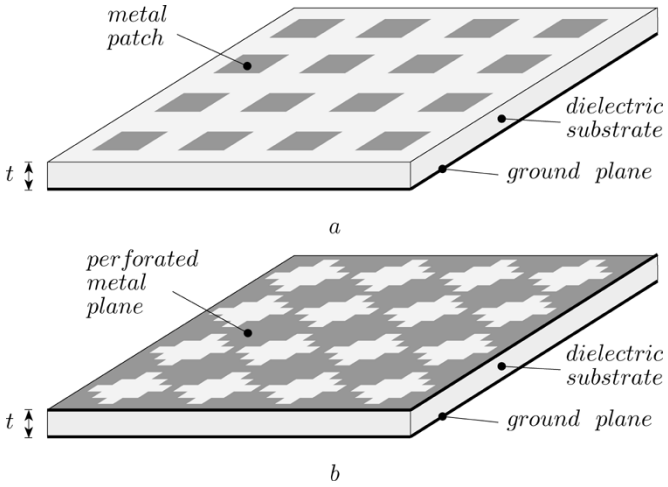

Fig. 1. Examples of planar electromagnetic band-gap structures. (a) Array of isolated metal patches on a grounded dielectric substrate. (b) Array of interconnected metal patches on a grounded dielectric substrate.

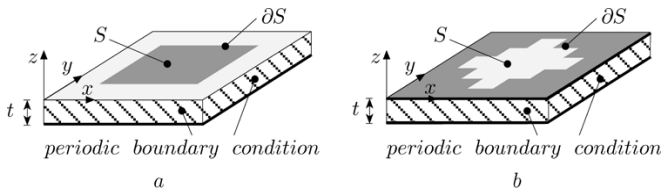

Fig. 2. Unit cells of planar electromagnetic band-gap structures of Fig. 1. (a) Array of metal patches. (b) Perforated metal plate.

II. Calculation of the Dispersion Diagram

Let us consider a periodic structure, consisting of either an array of arbitrarily-shaped isolated metal patches, patterned on an (un)grounded dielectric substrate [Fig. 1(a)], or a periodically-perforated metal plate, with an (un)grounded dielectric substrate underneath [Fig. 1(b)]. Due to the periodicity of the structure, the Floquet theorem can be applied and the electromagnetic problem reduces to the investigation of the unit cell of the periodic structure (Fig. 2). In our approach, the calculation is performed by choosing a propagation constant $\vec{\beta} = \beta_x \vec{u}_x + \beta_y \vec{u}_y$ (which determines the boundary conditions of the unit cell), and scanning the frequency f, to find the pairs $\{\vec{\beta}, f\}$ that correspond to points of the DD.

A. Application of the IEM to EBG With Metal Patches

In the case of an array of metal patches [Fig. 2(a)], for a given propagation constant $\vec{\beta}$ and frequency f, the boundary integral equation is obtained by imposing the electric wall condition on the metal patch with shape S [11]

$$\int_S \overleftrightarrow{G}_E(x,y,x',y') \cdot \vec{J}(x',y')dS' = 0 \quad \text{on } S \tag{1}$$

where $\vec{J}$ is the unknown electric current density on the patch and $\overleftrightarrow{G}_E$ is the Green's function which relates the transverse-to-z electric field at $z = 0$ to the transverse-to-z electric current density on the patch, and is given by

$$\overleftrightarrow{G}_E(x,y,x',y') = \sum_m Z_m \vec{\mathcal{E}}_m(x,y)\vec{\mathcal{E}}_m^*(x',y'). \tag{2}$$

In (2), $\vec{\mathcal{E}}_m$ is the electric modal vector of the mth Floquet mode [17], the asterisk denotes the complex conjugate, and Z_m is the parallel of the impedance seen by the mth Floquet mode, looking from the section $z = 0$ in the positive and negative direction of the z-axis.

The integral equation (1) is solved by using the MoM. The electric current density is expressed as a combination of vector basis functions $\vec{e}_j$

$$\vec{J}(x,y) = \sum_{j=1}^{N} \xi_j \vec{e}_j(x,y) \tag{3}$$

where ξ_j are unknown coefficients. In our approach, entire-domain basis functions are adopted, which span the whole domain S of the patch and are tangential to its boundary ∂S, thus satisfying the same boundary condition as the current density $\vec{J}$. A suitable set of basis functions are the electric modal vectors of a waveguide with cross-section S and magnetic-wall boundary condition on ∂S. In the case of arbitrarily-shaped patches, the entire-domain basis functions are calculated numerically by using the BI-RME method [18]. The advantage of using entire-domain basis functions is that few tens of functions are typically enough to guarantee the convergence of the method, as will be shown in Section III.

By substituting (3) in (1) and testing the integral equation (1) with the same set of functions $\vec{e}_i$, the following matrix problem is obtained

$$\mathbf{AX} = \mathbf{0} \tag{4}$$

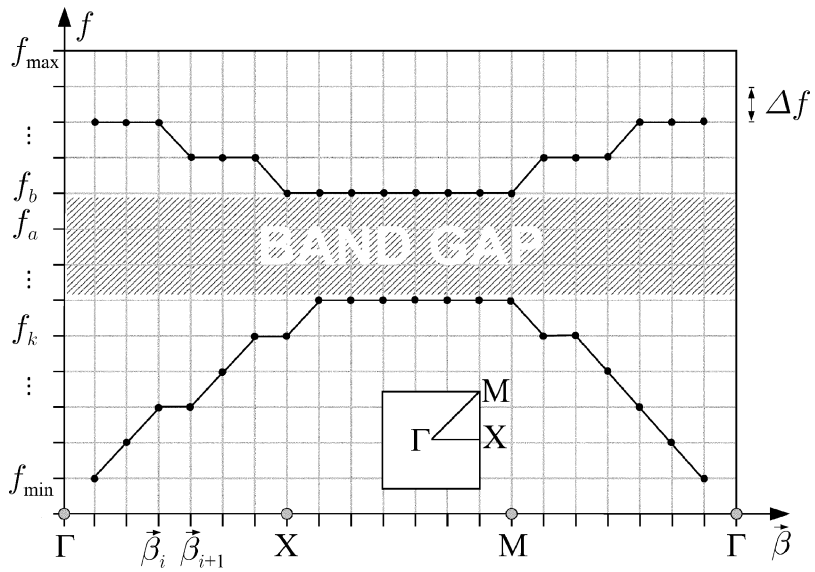

Fig. 3. Scheme of the grid used for identifying the bandgap, in the case of a structure with a square lattice.

where $\mathbf{X}$ is the vector of the unknown coefficients ξ_j and the entries of $\mathbf{A}$ are given by

$$A_{ij} = \sum_m Z_m \int_S \vec{e}_i(x,y) \cdot \vec{\mathcal{E}}_m(x,y) dS \times \int_S \vec{e}_j(x',y') \cdot \vec{\mathcal{E}}_m^*(x',y') dS'. \quad (5)$$

Due to the particular choice of the basis functions, surface integrals appearing in (5) can be transformed into line integrals [19], thus significantly increasing the computational efficiency of the method.

B. Application of the IEM to EBG With Perforated Metal Plates

In the case of a thin metal plate perforated with holes [Fig. 2(b)], for a given propagation constant $\vec{\beta}$ and frequency f, the boundary integral equation is obtained by imposing the continuity of the tangential component of the fields across the aperture with shape S [10]. In particular, by applying the equivalence theorem, the aperture is closed by a conductive sheet, and the continuity of the tangential component of the electric field is guaranteed by defining two unknown magnetic current densities $\vec{M}$ and $-\vec{M}$ on the opposite sides of the conductive sheet. Conversely, the continuity of the tangential component of the magnetic field is enforced, thus obtaining

$$\int_S \overleftrightarrow{G}_H(x,y,x',y') \cdot \vec{M}(x',y') dS' = 0 \quad \text{on } S \quad (6)$$

where $\overleftrightarrow{G}_H$ is the Green's function which relates the transverse-to-z magnetic field at $z = 0$ to the transverse-to-z magnetic current density on the aperture, and is given by

$$\overleftrightarrow{G}_H(x,y,x',y') = \sum_m Y_m \vec{\mathcal{H}}_m(x,y) \vec{\mathcal{H}}_m^*(x',y') \quad (7)$$

where $\vec{\mathcal{H}}_m$ is the magnetic modal vector of the mth Floquet mode, and Y_m is the parallel of the admittance seen by the mth Floquet mode, looking from the section $z = 0$ in the positive and negative direction of the z-axis.

In the application of the MoM, the magnetic current density is expressed as a combination of vector basis functions $\vec{h}_j$

$$\vec{M}(x,y) = \sum_{j=1}^{N} \xi_j \vec{h}_j(x,y) \quad (8)$$

where ξ_j are unknown coefficients, and $\vec{h}_j$ are entire-domain basis functions, corresponding to the magnetic modal vectors of a waveguide with the same cross-section S of the aperture and electric-wall boundary condition on ∂S.

The testing procedure by the same set of functions $\vec{h}_i$ leads to a matrix problem like (4), where

$$A_{ij} = \sum_m Y_m \int_S \vec{h}_i(x,y) \cdot \vec{\mathcal{H}}_m(x,y) dS \times \int_S \vec{h}_j(x',y') \cdot \vec{\mathcal{H}}_m^*(x',y') dS'. \quad (9)$$

Also in this case, the transformation of surface integrals into line integrals is applicable [19].

C. Search of the Points of the Dispersion Diagram

The approach proposed in this paper for finding the nontrivial solutions of (4) is based on the tracking of the eigenvalues of $\mathbf{A}(\vec{\beta}, f)$ in the complex plane. A solution is found if one or more eigenvalues cross the origin.

By considering a given propagation constant $\vec{\beta}_i$, the frequency range $f_{\min} \div f_{\max}$ is scanned with a given frequency step Δf (Fig. 3). For each interval with extremes f_a and f_b, the eigenvalues $\mathbf{\Lambda}_a$, $\mathbf{\Lambda}_b$ and the eigenvectors $\mathbf{V}_a$, $\mathbf{V}_b$ of matrix $\mathbf{A}$ at f_a and f_b, respectively, are calculated

$$\mathbf{A}(f_a, \vec{\beta}_i) = \mathbf{V}_a^{\mathrm{T}*} \mathbf{\Lambda}_a \mathbf{V}_a \quad (10)$$

$$\mathbf{A}(f_b, \vec{\beta}_i) = \mathbf{V}_b^{\mathrm{T}*} \mathbf{\Lambda}_b \mathbf{V}_b \quad (11)$$

where superscripts T denotes the transpose, and the eigenvectors are normalized according to

$$\mathbf{V}_a^{\mathrm{T}}\mathbf{V}_a^* = \mathbf{I} \quad \mathbf{V}_b^{\mathrm{T}}\mathbf{V}_b^* = \mathbf{I} \tag{12}$$

and $\mathbf{I}$ is the identity matrix. Note that the computational effort for finding eigenvalues and eigenvectors is very light, since the dimension of matrix $\mathbf{A}$ is typically of few tens [10], [11].

In order to follow the path of the eigenvalues in the complex plane from f_a to f_b, a correspondence between eigenvalues $\mathbf{\Lambda}_a$ and $\mathbf{\Lambda}_b$ is needed. This is obtained by defining the matrix

$$\boldsymbol{\mathcal{P}} = \mathbf{V}_a^{\mathrm{T}}\mathbf{V}_b^* \tag{13}$$

which exploits the correlation between the eigenvectors. In particular, due to the normalization condition (12), the element $\mathcal{P}_{ij}$ is close to unity if the ith eigenvector at f_a slightly differs from the jth eigenvector at f_b and, therefore, there is a strong correlation between Λ_{ai} and Λ_{bj}. Conversely, $\mathcal{P}_{ij}$ is close to zero if there is no correlation between Λ_{ai} and Λ_{bj}. A threshold procedure applied to $\boldsymbol{\mathcal{P}}$ leads to a correlation matrix, where any row or column comprises one element equal to unity, being all other elements equal to zero. This permits to track the path of each eigenvalue on the complex plane when varying f and, eventually, to detect the zero-crossing of one or more of them.

It may happen that the entries of $\boldsymbol{\mathcal{P}}$ cannot be clearly identified as one or zero by the threshold procedure. Two different cases may arise. If some modes (say D) are degenerate, an identical number of eigenvalues cross the origin in the complex plane at the same frequency, and the corresponding eigenvectors are garbled. In this case, D^2 elements of $\boldsymbol{\mathcal{P}}$ differ from zero or one, and can be recognized as the elements of a rotation matrix. Applying such a rotation to an eigenvector matrix (e.g., to $\mathbf{V}_a$) and reusing (13) permits to obtain, after the threshold procedure, a correlation matrix whose entries are only one or zeros. The second case happens when the frequency step Δf is too large, and the eigenvectors are weakly correlated. In this case, a large number of elements of $\boldsymbol{\mathcal{P}}$ differ from zero or one. Therefore, the frequency interval $f_a \div f_b$ is subdivided in two identical parts, and the procedure is applied to each sub-interval. It is noted that both cases are easily detected and solved in a fully automated way, with no need of human intervention.

Once a zero-crossing is detected, the frequency f_z of the zero-crossing is estimated by a linear interpolation. Its accuracy is then automatically checked by a refinement procedure: the frequency range $f_a \div f_b$ is divided into the sub-intervals $f_a \div f_z$ and $f_z \div f_b$, and the whole procedure is repeated for each sub-interval. This requires the electromagnetic analysis at f_z, the calculation of the eigenvalues and eigenvectors of matrix $\mathbf{A}$ at f_z, and the calculation of the correlation matrix $\boldsymbol{\mathcal{P}}$ for the two sub-intervals. The refinement stops when the relative distance between the new estimated zero-crossing frequency and f_z is smaller than a prescribed value ε (typically $\varepsilon = 0.1\%$). It is worth observing that only few refinement steps (typically less than three) are needed.

This iterative procedure is applied to all the values $\vec{\beta}_i$, thus obtaining a list of points on the dispersion diagram.

The same correlation concept can also be applied for interconnecting these points, in order to draw the dispersion curves and find a possible bandgap. Also in this case, we exploit the correlation of the eigenvectors by calculating matrix $\boldsymbol{\mathcal{P}}$ for all the zeros found for two closely spaced values of $\vec{\beta}$ (namely $\vec{\beta}_i$ and $\vec{\beta}_{i+1}$). Note that this task is almost straightforward, since the needed eigenvectors have been already calculated and stored.

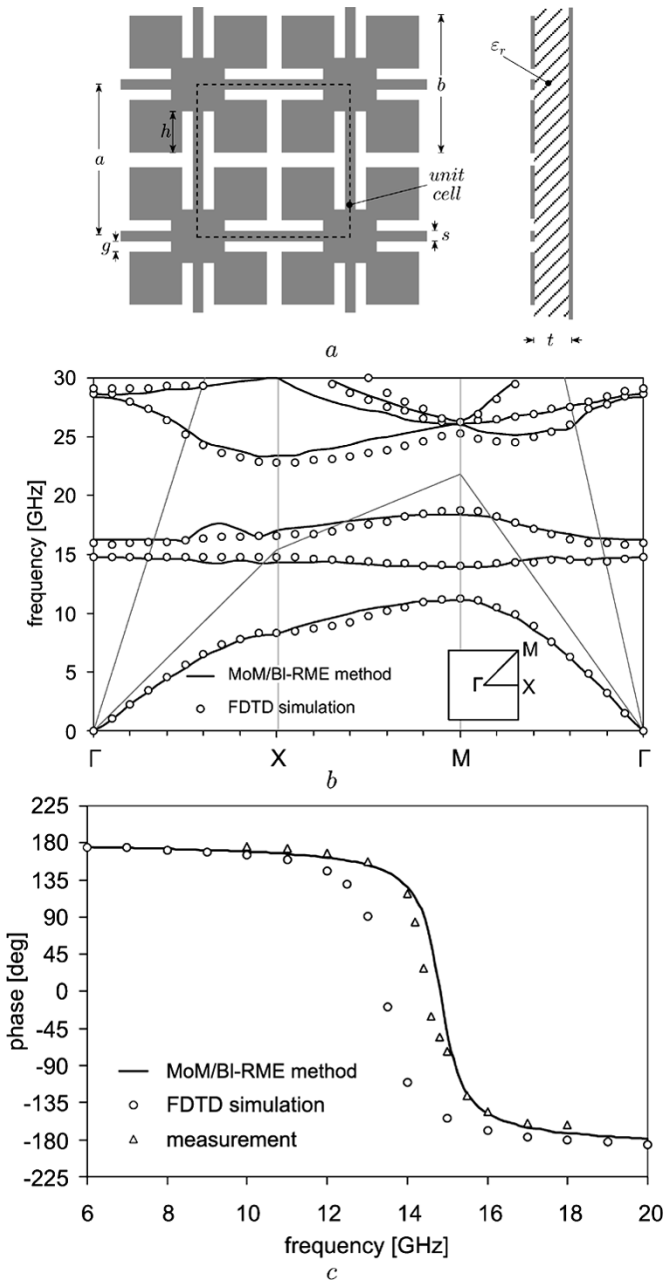

Fig. 4. Analysis of the UC-PBG structure. (a) Geometry of the unit cell (dimensions: $a = 3.048$ mm, $b = 2.794$ mm, $g = s = 0.254$ mm, $h = 0.6985$ mm). (b) Dispersion diagram; c) phase of the reflection coefficient.

III. Numerical Results

The classical UC-PBG structure [1] is used for validating the proposed method. It consists of a dielectric layer metallized on both sides (thickness $t = 0.635$ mm, relative dielectric constant $\varepsilon_r = 10.2$), with Jerusalem-cross shaped holes etched in the top side [Fig. 4(a)]. Since the portion of metal is larger than the holes, in the electromagnetic analysis it is more convenient

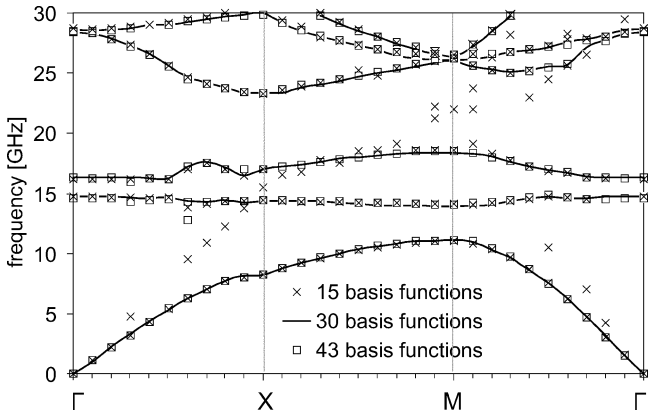

Fig. 5. Convergence properties of the MoM/BI-RME method versus the number of basis functions in the calculation of the dispersion diagram of Fig. 4(b).

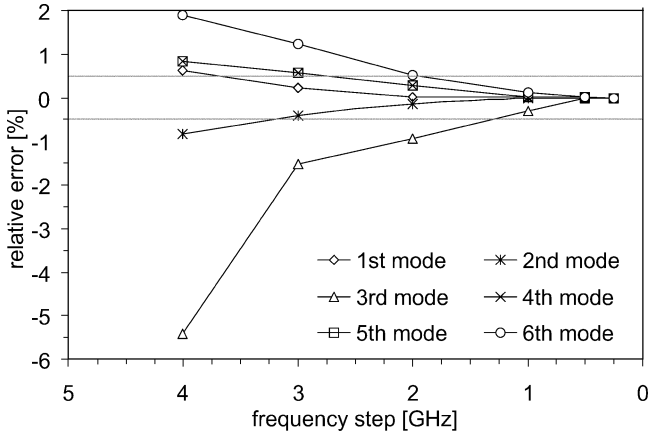

Fig. 6. Relative error in the calculation of the dispersion diagram versus the frequency step Δf used in the search of the eigenvalue zero-crossing (M-point: $\beta_x = \beta_y = \pi/a$).

to apply the theory described in Section II-B. Therefore, the unit cell is the one shown in Fig. 4(a), and the unknown of the problem is the magnetic current density defined on the hole.

The dispersion diagram was calculated by considering 30 basis functions and 1200 Floquet modes. The calculation was performed in 30 points on the boundary of the irreducible Brillouin zone, the frequency step adopted in the zero-search was $\Delta f = 1$ GHz, and the required accuracy was $\varepsilon = 0.1\%$. The overall process required about 1100 electromagnetic full-wave analyzes: 900 analyzes for the preliminary grid scan (30 steps in $\vec{\beta}$ times 30 steps in f, Fig. 3) plus about 200 analyzes for the refinement procedure. The results obtained by the MoM/BI-RME method are shown in Fig. 4(b) and compare well with the FDTD simulation reported in [20]. The overall computing time was about 18 min on a PC Pentium IV at 2.4 GHz.

The phase of the reflection coefficient in the case of normal incidence was calculated by using the same number of basis functions and Floquet modes, and is shown in Fig. 4(c). The reference plane was set to the top side of the dielectric layer, and this required to shift the measured phase reported in [3], where the reference plane was set to the ground plane. A very good agreement is found with the measurement results, whereas a frequency shift is noted with respect to the FDTD simulations

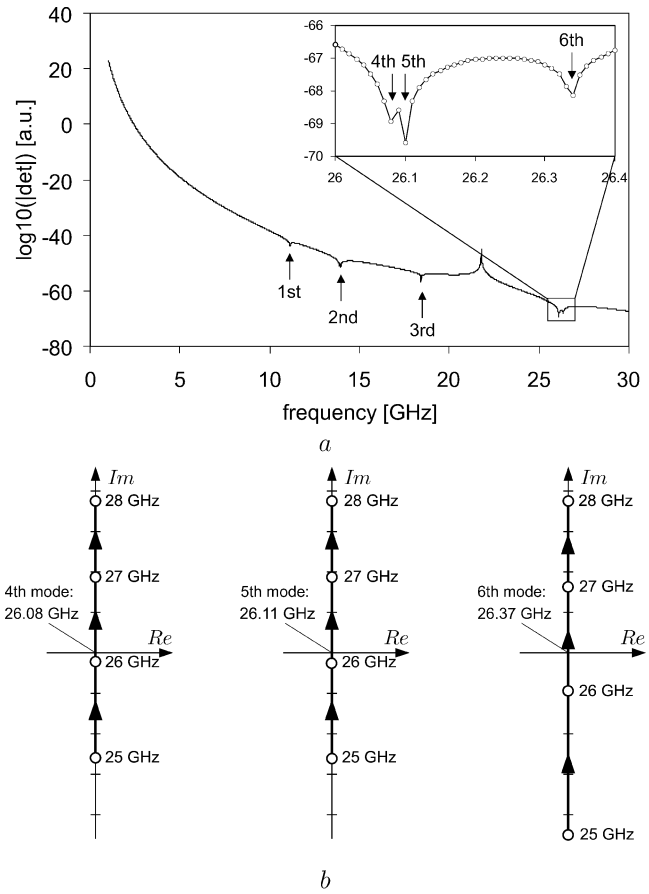

Fig. 7. Performance comparison between different technique for the calculation of the dispersion diagram. (a) Search of the determinant zeros. (b) Tracking of the path of some eigenvalues in the complex plane.

[3]. The overall computing time for calculating the phase of the reflection coefficient in 70 frequency points was 9 s on a PC Pentium IV at 2.4 GHz.

It is worth observing that the accuracy parameters used in these analyzes were chosen after some convergence tests. In particular, the effect of the number of basis functions on the dispersion diagram (Fig. 5) shows that 30 basis functions permit to reach the convergence. Actually, when using 15 functions, the curves are correctly identified, but some spurious solutions appear.

An important test is related to the reliability and the accuracy of the method, in the absence of the refinement procedure (i.e., by considering only the fixed frequency step Δf and the linear interpolation). To this aim, the M-point of the irreducible Brillouin zone [$\beta_x = \beta_y = \pi/a$; see Fig. 4(b)] was considered. This point is critical, since three modes are almost degenerate around 26 GHz. The frequency scan was performed with different values of Δf (namely, 4, 3, 2, 1, 0.5, and 0.1 GHz). Fig. 6 shows the relative error in the calculation of the mode frequencies (the reference value used for calculating the relative error is taken at a frequency step $\Delta f = 0.1$ GHz). It is observed that, even with a frequency step Δf as large as 4 GHz, the first six modes are found, and the accuracy is better than 6%. Moreover, an accuracy better than 0.5% is obtained with $\Delta f = 1$ GHz.

To better appreciate the robustness of the proposed method, we performed the calculation of the mode frequencies at the

M-point of the dispersion diagram by using the classical approach, based on the search of the zeros of the determinant of the MoM matrix. Fig. 7(a) shows that a frequency step of 10 MHz is required for identifying the zeros (in particular, the ones corresponding to the fourth and fifth modes). Conversely, by tracking the path of the MoM-matrix eigenvalues on the complex plane [Fig. 7(b)], these modes can be easily identified by using a frequency step of 1 GHz, even when their frequencies are almost coincident.

IV. Conclusion

This paper presented a novel method for the efficient, reliable, and fully automated calculation of the dispersion diagram of planar electromagnetic band-gap structures.

The efficiency of the method depends on the particular technique adopted for finding the points of the dispersion diagram, which requires a reduced number of electromagnetic analysis, and on the use of the MoM/BI-RME method for the electromagnetic characterization of the periodic structure. In fact, the MoM/BI-RME method yields the MoM matrix in a very short time, thanks to the use of entire-domain basis functions and to a special line-integral formulation of the entries of the matrix. The use of entire-domain basis functions also leads to a MoM matrix with small dimension, and therefore the calculation of its eigenvalues and eigenvectors is extremely fast.

Moreover, the novel technique based on the tracking of the eigenvalues in the complex plane takes advantage of some information on the correlation between the eigenvectors at different frequencies, which are not exploited in the classical direct search of the determinant zeros. This permits to detect the zeros even with a large frequency step, and to accurately determine their frequency with few refinement steps. Furthermore, the information on the correlation of the eigenvectors also permits to interconnect the points of the dispersion diagram at different values of the propagation constant, thus automatically obtaining the identification of the modes.

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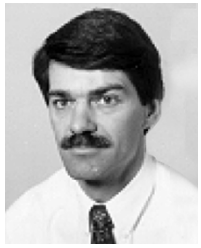

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