

# Efficient Analysis of Radiation From a Dipole Source in Woodpile EBG Structures

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**Abstract** — An efficient and rigorous approach to analyze the radiation from an electric dipole placed in three-dimensional woodpile EBG structures consisting of layered periodic arrays of circular cylinders is presented. The response in the spectral domain of the woodpile EBGs excited by a dipole source is calculated using one-dimensional lattice sums, the three-dimensional transition matrix of a circular cylinder, and the generalized reflection and transmission matrices of the layered system. The closed-form expression of the radiated field is rigorously obtained by a finite integral of the spectral response. Comparisons with the far fields calculated by using reciprocity both with the present approach and with a fully independent commercial EM software are presented showing excellent agreement with the proposed method and demonstrating the high computational efficiency of the semi-analytical formalism.

**Index Terms** — Electromagnetic bandgap materials, dipole source, periodic structure, radiation, woodpile structure.

## I. INTRODUCTION

ELECTROMAGNETIC bandgap materials (EBGs) are multilayered periodic dielectric or metallic structures in which electromagnetic filtering is achieved and wave propagation is allowed or forbidden within a possibly large frequency range (electromagnetic bandgap) [1], [2]. In the field of antenna technology, the EBG structures, with their unique ability to block the electromagnetic wave propagation within a limited frequency range, have been used as substrates and covers for the sources to improve their radiating characteristics

[3]. The full potential of EBG structures can be utilized with a three-dimensional (3-D) bandgap. In this regard, the woodpile EBG can be considered as a promising candidate because of relatively easy fabrication for applications at microwave and mmWave frequencies, with a possible extension up to THz frequencies [4]–[11]. Woodpile materials find applications as: resonator antennas with high directivity over a narrow bandwidth [12], EBG sectoral horn antennas [13]; passive waveguides and waveguide components [14]. Many of these devices operate at submillimeter-wave or THz frequencies, where the woodpile EBG material is particularly attractive, since it may be accurately fabricated. Furthermore, even in low-index materials woodpile structures exhibit large bandgaps for many rod sizes or in-plane spacings, which is promising performance, especially in the perspective of using additive manufacturing technologies based on innovative 3-D printing techniques and polymer jetted devices [10], [15].

The design of Fabry-Perot cavity antennas with 3-D EBG superstrates is often performed by using time consuming simulations with commercial EM software based on the Finite-Difference Time-Domain (FDTD) method or the Finite Element Method (FEM) in the frequency domain [5],[8],[9]. Other techniques are available for the computation of dispersion band diagrams of infinite two-dimensional (2-D) EBG lattices [1],[6]. Unlike similar Fabry-Perot antennas based on printed superstrates, where semi-analytical methods based on the Method of Moments and the Array Scanning Method (ASM) have been used [16]–[20], there is a lack of tailored numerical tools for the analysis of dipole excited infinite periodic dielectric woodpile antennas, as well as for the study of hybrid leaky modes propagating in this type of 3-D dielectric EBG structure.

In this paper, we propose a novel efficient, self-contained and rigorous method to analyze the radiation from a horizontal electric dipole (HED) coupled to a woodpile EBG structure consisting of a layered periodic arrays of cylinders. The cylinders in each layer are infinitely long and parallel to each other, while the cylinders' axes are rotated by  $90^\circ$  in each successive layer along the transverse direction ( $y$ -axis in Fig.1). The formalism is general and it rigorously takes into account all the space-harmonics of the co-polarized and cross-polarized fields and the interaction between them.

The approach is based on the ASM, which is an analytical technique that allows an aperiodic function to be expressed as an integral superposition of Floquet-periodic functions in the spectral domain [16], [21]–[25]. The HED is transformed into an infinite periodic array of linearly phased HEDs in the

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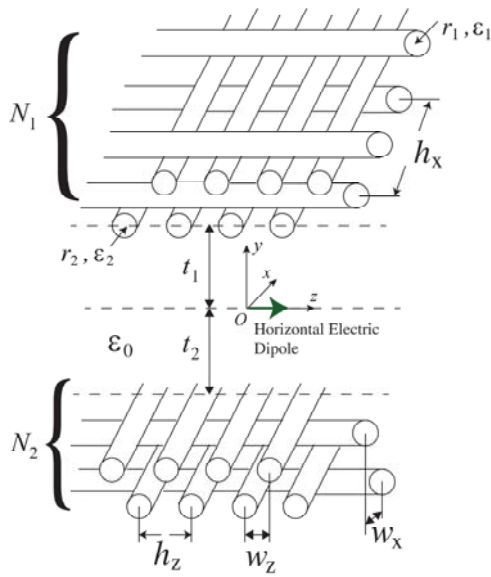


Fig. 1. A localized HED located at the global origin and sandwiched by woodpile EBG structures formed from layered crossed arrays of circular cylinders. The cylinders in each layer are infinitely long and parallel to each other, while the cylinders' axes are rotated by  $90^\circ$  in each successive layer along  $y$ -axis.

spectral domain and for each of the spectral components, the HED arrays excite a corresponding set of space-harmonic fields. The spectral response of the EBG structures to the HED excitation is calculated using the one-dimensional (1-D) periodic lattice sums (LSs) [26]–[28], the 3-D transition matrix (T-matrix) of a circular cylinder in isolation [1], and the generalized reflection and transmission matrices associated with the multilayered structure [26],[27],[29]. This approach is particularly suitable for the analysis of periodic and bandgap structures formed by array of circular cylinders, where the effective representation of the scattered fields from the periodic objects in terms of cylindrical harmonics naturally involves the LSs formulation [26]. Moreover, the proposed approach can be applied to various configurations of layered arrays with different types and locations of the excitation sources.

A closed form expression for the radiated field is obtained from the finite integral of the spectral response by using the method of the steepest descent [23], [30]. Three configurations of the woodpile EBG structures are studied in the manuscript. The normalized radiation patterns of the HED coupled to woodpile structures are calculated and compared with the results obtained by means of CST Microwave Studio [31]. A very good agreement is observed for all the configurations, whereas our proposed formalism is much faster than the commercial software CST showing a speed improvement of one to two order of magnitudes. The apparent computational advantage makes the proposed quasi analytical approach extremely useful in those optimization procedures that require extensive parametric analysis of such kind of EBG structures.

The main goal of the paper is to demonstrate, for the first time, a numerically very efficient and flexible full-wave approach applicable to a wide class of 3-D woodpile structures. Although only the far-field is analyzed in the manuscript, also for validation purposes, the quasi analytical formalism can be

suitably modified to allow for tailored numerical analyses, which cannot be provided by general purpose EM software. In particular, the rigorous study of the complex and leaky hybrid modes supported by 3-D EBG waveguides [25], as well as the calculation of the near field excited by aperiodic sources [19] in the woodpile EBG antennas, can be considered as possible extension of the adopted numerical tool, which therefore results in a very versatile approach for a systematic and insight design of such kind of woodpile EBG radiators.

This paper is organized as follows. In Section II the field solutions for a localized HED coupled to the woodpile EBG structures is found. An analytical expression for the electric field of the HED is rigorously expressed in terms of the generalized reflection and transmission matrices of the woodpile structures. In Section III, the generalized reflection and transmission matrices for the periodic arrays along the  $z$ - and  $x$ -axis (Fig. 1) for the co-polarized and crossed-polarized fields are separately derived. We note that unlike a 2-D problem [24], [25], the crossed-polarized terms contribute to the formation of the radiation pattern. In Section IV the reciprocity theorem to calculate the radiated far-field is briefly discussed and the expression for the scattered fields at the location of the HED is rigorously obtained. The normalized radiation patterns of the HED coupled to three different configurations of the woodpile EBG structures at different excitation frequencies are demonstrated in Section V. Here, results obtained with the formalism presented in Section II have been validated with those obtained with the same approach, but using reciprocity, as described in Section IV, and with a fully independent commercial software, i.e., CST Microwave Studio. Finally, the computation times for all three configurations are summarized. Concluding remarks are given in Section VI. A time-harmonic dependence of  $e^{j\omega t}$  is assumed in the manuscript.

## II. FORMULATION OF THE PROBLEM

The geometry for the problem is shown in Fig. 1. A HED oriented in the  $z$ -direction and located at the origin  $O$  of the reference system is sandwiched in between two EBG structures with the separation distances  $t_1$  and  $t_2$  from the source plane. Although in the manuscript only a HED is analyzed, the proposed formulation can be easily extended to the radiation produced by a horizontal magnetic dipole (HMD). The cylinders in each layer are infinitely long and parallel to each other, while the cylinders' axes are rotated by  $90^\circ$  in each successive layer separated along the  $y$ -axis. The periodic spacing, the radius, and the dielectric permittivity of the circular rods are  $h_x$  and  $h_z$ ,  $r_1$  and  $r_2$ ,  $\epsilon_1$  and  $\epsilon_2$ , respectively. The number of the layers comprising the upper and lower EBGs are  $N_1$  and  $N_2$ , and the shifts of the cylinders from the global origin  $O$  along the  $z$ - and  $x$ -axis are  $w_z$  and  $w_x$ , respectively. The cylinders must be identical along each layer of the arrays but those in different layers can differ in material properties and dimensions. Without loss of generality, the background medium is assumed to be free space. Since the source is localized, the entire system is not periodic and the conventional ray-tracing technique of space-harmonics along the  $y$ -axis cannot be applied. Instead, we express the localized HED in terms of a spectral

wavenumber integration over a phased array of dipole sources along the  $z$ - and  $x$ -axis [16],[21]-[25]:

$$\delta(x)\delta(z) = \frac{h_x h_z}{4\pi^2} \int_{-\pi/h_x}^{\pi/h_x} dk_x \left[ \sum_{m=-\infty}^{\infty} \delta(z - mh_z) e^{-jmh_z k_z} \right] \times \int_{-\pi/h_x}^{\pi/h_x} dk_x \left[ \sum_{\ell=-\infty}^{\infty} \delta(x - \ell h_x) e^{-j\ell h_x k_x} \right] \quad (1)$$

where  $k_x$  and  $k_z$  are the spectral parameters and  $\delta$  denotes the Dirac delta function. From (1) it follows that if  $E_{z,P}(x, y, z, k_x, k_z)$  is the electric field radiated from a periodic array of HEDs with a current density defined as:

$$J_{z,P}(x, y, z, k_x, k_z) = \mathcal{I} \sum_{m=-\infty}^{\infty} \sum_{\ell=-\infty}^{\infty} \delta(x - \ell h_x) \delta(y) \delta(z - mh_z) e^{-j\ell h_x k_x} e^{-jm h_z k_z} \quad (2)$$

where  $\mathcal{I}$  denotes the current dipole moment, then the radiated field  $E_z(x, y, z)$  from the localized HED:

$$J_z(x, y, z) = \mathcal{I} \delta(x) \delta(y) \delta(z) \quad (3)$$

is given in the following form:

$$E_z(x, y, z) = \frac{h_x h_z}{4\pi^2} \int_{-\pi/h_x}^{\pi/h_x} dk_x \int_{-\pi/h_z}^{\pi/h_z} dk_z E_{z,P}(x, y, z, k_x, k_z) \quad (4)$$

Thus the problem is reduced to the calculation of the electric field  $E_{z,P}(x, y, z, k_x, k_z)$  radiated from the infinite periodic array of the linearly phased HEDs  $J_{z,P}(x, y, z, k_x, k_z)$ . To calculate  $E_{z,P}(x, y, z, k_x, k_z)$ , the primary field  $E_z^{inc}(x, y, z, k_x, k_z)$  radiated from the periodic HEDs must be defined (the  $z$ -component of the magnetic field  $H_z$  of the HED is equal to zero). The primary field is expressed in the following form:

$$E_z^{inc}(x, y, z, k_x, k_z) = \frac{-j\omega\mu_0 \mathcal{I}}{4\pi} \sum_{m=-\infty}^{\infty} \sum_{\ell=-\infty}^{\infty} \frac{e^{-jk_0 r_{lm}}}{r_{lm}} e^{-j\ell h_x k_x} e^{-jm h_z k_z} \quad (5)$$

where  $r_{lm} = \sqrt{(x - \ell h_x)^2 + y^2 + (z - mh_z)^2}$  and  $k_0$  is the free-space wavenumber. After applying the plane-wave integral representation of the spherical wave:

$$\frac{e^{-jk_0 r_{lm}}}{r_{lm}} = \frac{-j}{2\pi} \int_{-\infty}^{\infty} d\tilde{k}_x \int_{-\infty}^{\infty} d\tilde{k}_z \frac{1}{k_y} e^{-j\tilde{k}_x(x - \ell h_x) - j\tilde{k}_z(z - mh_z) - jk_y|y|} \quad (6)$$

and taking into account the identity:

$$\sum_{\ell=-\infty}^{\infty} e^{-j\ell h_x k_x} e^{j\ell h_x \tilde{k}_x} = \frac{2\pi}{h_x} \sum_{\ell=-\infty}^{\infty} \delta\left(k_x - \tilde{k}_x + \frac{2\pi\ell}{h_x}\right) \quad (7)$$

the primary field  $E_z^{inc}(x, y, z, k_x, k_z)$  from the periodic dipole sources can be expressed in the following form:

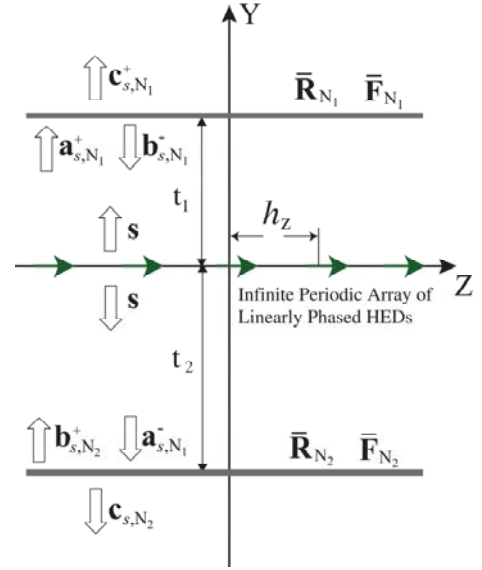


Fig.2. Schematic view of multiple scattering processes for the space-harmonic waves primarily radiated from the infinite periodic arrays of linearly phased HEDs. The upper and lower EBG structures are replaced by two plane boundaries characterized by the generalized reflection and transmission matrices  $(\bar{\mathbf{R}}_{N_1}, \bar{\mathbf{F}}_{N_1})$  and  $(\bar{\mathbf{R}}_{N_2}, \bar{\mathbf{F}}_{N_2})$ .

$$E_z^{inc}(x, y, z, k_x, k_z) = -\frac{\omega\mu_0 \mathcal{I}}{2h_x h_z} \sum_{\ell=-\infty}^{\infty} \sum_{m=-\infty}^{\infty} \frac{e^{-jk_{x\ell}x - jk_{zm}z - jk_{y,\ell m}|y|}}{k_{y,\ell m}} \quad (8)$$

with  $k_{x\ell} = k_x + 2\ell\pi/h_x$ ,  $k_{zm} = k_z + 2m\pi/h_z$  and  $k_{y,\ell m} = (k_0^2 - k_{x\ell}^2 - k_{zm}^2)^{1/2}$ . The periodically arranged HEDs (Fig. 2) located on the plane  $y=0$  excite space harmonic fields of the plane waves which are up-going in the region  $y > 0$  and down-going in region  $y < 0$ . Employing  $e^{-jk_{x\ell}x - jk_{zm}z - jk_{y,\ell m}|y|}$  as the basis for the field representation, the up-going and down-going space harmonics given by (8) are expressed in terms of the amplitude vector  $\mathbf{s}(k_x, k_z)$  defined as follows:

$$\mathbf{s}(k_x, k_z) = \begin{Bmatrix} \mathbf{s}_E(k_x, k_z) \\ \mathbf{0} \end{Bmatrix}, \quad \mathbf{s}_E = \begin{Bmatrix} -\frac{\omega\mu_0 \mathcal{I}}{2h_x h_z k_{y,\ell m}} \end{Bmatrix} \quad (9)$$

Here, the scattering problem is formulated by employing the  $z$ -components of the fields as the leading fields and a zero vector in (9) corresponds to the amplitude of the  $z$ -component of the incident magnetic field. In the manuscript,  $\{\cdot\}$  and  $[\cdot]$  denote the vectors and matrices, respectively. Besides, the vectors and matrices are marked as boldface symbols. When the spectral amplitude of the infinite, periodically arranged, linearly phased HEDs is defined, the radiated field can be calculated using the technique of reflection and transmission matrices for layered periodic arrays subject to plane wave incidence. The two EBG structures located in the upper and lower space with respect to the HEDs may be replaced by two plane boundaries located at  $y = t_1$  and  $y = -t_2$ . They are

characterized by the generalized reflection and transmission matrices  $(\bar{\mathbf{R}}_{N_1}, \bar{\mathbf{F}}_{N_1})$  and  $(\bar{\mathbf{R}}_{N_2}, \bar{\mathbf{F}}_{N_2})$  as illustrated in Fig. 2. The reflection and transmission matrices of the upper (lower) EBG structures describe a relation between the reflected and transmitted space-harmonics and an amplitude of the incident field, which is given as a sum of the original source amplitude vector  $\mathbf{s}(k_x, k_z)$  and the amplitude vector of the reflected field from the lower (upper) EBG structures.

It should be noted that the generalized reflection  $\bar{\mathbf{R}}_{N_{1(2)}}$  and transmission  $\bar{\mathbf{F}}_{N_{1(2)}}$  matrices (overlined matrices) consist of

four block matrices, namely  $\bar{\mathbf{R}}_{N_{1(2)}} = \begin{bmatrix} \mathbf{R}_{N_{1(2)}}^{(EE)} & \mathbf{R}_{N_{1(2)}}^{(EH)} \\ \mathbf{R}_{N_{1(2)}}^{(HE)} & \mathbf{R}_{N_{1(2)}}^{(HH)} \end{bmatrix}$

and  $\bar{\mathbf{F}}_{N_{1(2)}} = \begin{bmatrix} \mathbf{F}_{N_{1(2)}}^{(EE)} & \mathbf{F}_{N_{1(2)}}^{(EH)} \\ \mathbf{F}_{N_{1(2)}}^{(HE)} & \mathbf{F}_{N_{1(2)}}^{(HH)} \end{bmatrix}$ . The diagonal block matrices

$(\mathbf{R}_{N_1}^{(EE)}, \mathbf{R}_{N_1}^{(HH)})$ ,  $(\mathbf{R}_{N_2}^{(EE)}, \mathbf{R}_{N_2}^{(HH)})$  and  $(\mathbf{F}_{N_1}^{(EE)}, \mathbf{F}_{N_1}^{(HH)})$ ,  $(\mathbf{F}_{N_2}^{(EE)}, \mathbf{F}_{N_2}^{(HH)})$  represent the reflection and transmission into the co-polarized electric and magnetic fields, while the non-diagonal block matrices  $(\mathbf{R}_{N_1}^{(EH)}, \mathbf{R}_{N_1}^{(HE)})$ ,  $(\mathbf{R}_{N_2}^{(EH)}, \mathbf{R}_{N_2}^{(HE)})$  and  $(\mathbf{F}_{N_1}^{(EH)}, \mathbf{F}_{N_1}^{(HE)})$ ,  $(\mathbf{F}_{N_2}^{(EH)}, \mathbf{F}_{N_2}^{(HE)})$  describe the reflection and transmission into the cross-polarized space-harmonic components. The expressions for block matrices are briefly given in Section III.

Let  $\mathbf{a}_{s,N_1}^+ = \begin{Bmatrix} \mathbf{a}_{s,N_1}^{+(E)} \\ \mathbf{a}_{s,N_1}^{+(H)} \end{Bmatrix}$  and  $\mathbf{b}_{N_1}^- = \begin{Bmatrix} \mathbf{b}_{s,N_1}^{-(E)} \\ \mathbf{b}_{s,N_1}^{-(H)} \end{Bmatrix}$  be the

amplitude vectors of the electric and magnetic fields of the scattered space-harmonics incoming and outgoing for the

plane  $y = t_1$ ,  $\mathbf{a}_{s,N_2}^- = \begin{Bmatrix} \mathbf{a}_{s,N_2}^{-(E)} \\ \mathbf{a}_{s,N_2}^{-(H)} \end{Bmatrix}$  and  $\mathbf{b}_{s,N_2}^+ = \begin{Bmatrix} \mathbf{b}_{s,N_2}^{+(E)} \\ \mathbf{b}_{s,N_2}^{+(H)} \end{Bmatrix}$  be

those for the plane  $y = -t_2$ , respectively. Then, a relation between the scattering amplitudes and an amplitude  $\mathbf{s}(k_x, k_z)$  of the incident field can be written in the following form:

$$\mathbf{a}_{s,N_1}^+(k_x, k_z) = \bar{\mathbf{\Lambda}}(k_x, k_z) \cdot \mathbf{b}_{s,N_2}^+(k_x, k_z) \quad (10)$$

$$\mathbf{a}_{s,N_2}^-(k_x, k_z) = \bar{\mathbf{\Lambda}}(k_x, k_z) \cdot \mathbf{b}_{s,N_1}^-(k_x, k_z) \quad (11)$$

where

$$\bar{\mathbf{\Lambda}}_1 = \begin{bmatrix} \mathbf{\Lambda}_1(k_x, k_z) & 0 \\ 0 & \mathbf{\Lambda}_1(k_x, k_z) \end{bmatrix}, \mathbf{\Lambda}_1 = \begin{bmatrix} e^{-jk_{y,\ell m} t_1} \end{bmatrix} \quad (12)$$

$$\bar{\mathbf{\Lambda}}_2 = \begin{bmatrix} \mathbf{\Lambda}_2(k_x, k_z) & 0 \\ 0 & \mathbf{\Lambda}_2(k_x, k_z) \end{bmatrix}, \mathbf{\Lambda}_2 = \begin{bmatrix} e^{-jk_{y,\ell m} t_2} \end{bmatrix} \quad (13)$$

$$\bar{\mathbf{\Lambda}} = \begin{bmatrix} \mathbf{\Lambda}(k_x, k_z) & 0 \\ 0 & \mathbf{\Lambda}(k_x, k_z) \end{bmatrix}, \mathbf{\Lambda} = \begin{bmatrix} e^{-jk_{y,\ell m} (t_1+t_2)} \end{bmatrix} \quad (14)$$

$$\mathbf{b}_{s,N_1}^-(k_x, k_z) = \bar{\mathbf{R}}_{N_1}(k_x, k_z) \cdot \{\mathbf{a}_{s,N_1}^+(k_x, k_z) + \bar{\mathbf{\Lambda}}_1(k_x, k_z) \cdot \mathbf{s}(k_x, k_z)\} \quad (15)$$

$$\mathbf{b}_{s,N_2}^+(k_x, k_z) = \bar{\mathbf{R}}_{N_2}(k_x, k_z) \cdot \{\mathbf{a}_{s,N_2}^-(k_x, k_z) + \bar{\mathbf{\Lambda}}_2(k_x, k_z) \cdot \mathbf{s}(k_x, k_z)\} \quad (16)$$

$$\mathbf{c}_{s,N_1}^+(k_x, k_z) = \bar{\mathbf{F}}_{N_1}(k_x, k_z) \cdot \{\mathbf{a}_{s,N_1}^+(k_x, k_z) + \bar{\mathbf{\Lambda}}_1(k_x, k_z) \cdot \mathbf{s}(k_x, k_z)\} \quad (17)$$

$$\mathbf{c}_{s,N_2}^-(k_x, k_z) = \bar{\mathbf{F}}_{N_2}(k_x, k_z) \cdot \{\mathbf{a}_{s,N_2}^-(k_x, k_z) + \bar{\mathbf{\Lambda}}_2(k_x, k_z) \cdot \mathbf{s}(k_x, k_z)\} \quad (18)$$

We note that in 2-D structures, the non-diagonal block matrices become zero and the problem can be easily solved for TE and TM waves separately [26], [27]. The scattering amplitude  $\mathbf{c}_{s,N_1}^+(k_x, k_z)$  of the transmitted field in the uppermost region of the EBG structure and the scattering amplitude  $\mathbf{c}_{s,N_2}^-(k_x, k_z)$  of the transmitted field in the lowermost region of the EBG structure are expressed through the incident spectral amplitudes  $\mathbf{s}(k_x, k_z)$  using the following relations:

$$\mathbf{c}_{s,N_1}^+ = \bar{\mathbf{F}}_{N_1} \cdot \left[ (\bar{\mathbf{I}} - \bar{\mathbf{\Lambda}} \cdot \bar{\mathbf{R}}_{N_2} \cdot \bar{\mathbf{\Lambda}} \cdot \bar{\mathbf{R}}_{N_1})^{-1} \cdot \bar{\mathbf{\Lambda}} \cdot \bar{\mathbf{R}}_{N_2} \cdot \bar{\mathbf{\Lambda}}_2 \cdot (\bar{\mathbf{I}} + \bar{\mathbf{\Lambda}}_1 \cdot \bar{\mathbf{R}}_{N_1} \cdot \bar{\mathbf{\Lambda}}_1) + \bar{\mathbf{\Lambda}}_1 \right] \cdot \mathbf{s} \quad (19)$$

$$\mathbf{c}_{s,N_2}^- = \bar{\mathbf{F}}_{N_2} \cdot \left[ (\bar{\mathbf{I}} - \bar{\mathbf{\Lambda}} \cdot \bar{\mathbf{R}}_{N_1} \cdot \bar{\mathbf{\Lambda}} \cdot \bar{\mathbf{R}}_{N_2})^{-1} \cdot \bar{\mathbf{\Lambda}} \cdot \bar{\mathbf{R}}_{N_1} \cdot \bar{\mathbf{\Lambda}}_1 (\bar{\mathbf{I}} + \bar{\mathbf{\Lambda}}_2 \cdot \bar{\mathbf{R}}_{N_2} \cdot \bar{\mathbf{\Lambda}}_2) + \bar{\mathbf{\Lambda}}_2 \right] \cdot \mathbf{s} \quad (20)$$

where  $\bar{\mathbf{I}}$  is the block unit matrix. Hence, the electric field  $E_{z,P}(x, y, z, k_x, k_z)$  radiated from the infinite periodic array of linearly phased HEDs can be expressed as:

$$E_{z,P}(x, y, z, k_x, k_z) = \begin{cases} \sum_{\ell=-\infty}^{\infty} \sum_{m=-\infty}^{\infty} c_{s,N_1,\ell m}^{+(E)}(k_x, k_z) e^{-jk_{x\ell}x - jk_{zm}z - jk_{y,\ell m}y}, & y > 0 \\ \sum_{\ell=-\infty}^{\infty} \sum_{m=-\infty}^{\infty} c_{s,N_2,\ell m}^{-(E)}(k_x, k_z) e^{-jk_{x\ell}x - jk_{zm}z + jk_{y,\ell m}y}, & y < 0 \end{cases} \quad (21)$$

Substituting (21) into (4), the radiation field from the localized dipole source is obtained. For the radiated field in the far-zone  $k_0 r \gg 1$ , the integral with the spectral parameters in (4) can be evaluated using the conventional asymptotic method [30]. The radiation pattern  $P(\theta, \varphi)$  can be calculated in the following form:

$$P(\theta, \varphi) = (|E_\theta|^2 + |E_\varphi|^2) = (1 - \sin^2 \theta \sin^2 \varphi) (|E_z|^2 + |\tilde{H}_z|^2) \quad (22)$$

where  $\tilde{H}_z = \sqrt{(\mu_0 / \epsilon_0)} H_z$  is the normalized magnetic field,  $0 < \theta < 90^\circ$ ,  $0 < \varphi < 360^\circ$ . All the space-harmonics radiating in the free space are properly taken into account in the calculation of  $\{C_{S,N1,\ell m}^+\}$  and  $\{C_{S,N2,\ell m}^-\}$ .

In this paper, we focus only on the far-field radiated by a HED placed in a woodpile EBG structure; however, it should be noted that the near field can be evaluated by calculating the integral in (4) along appropriate paths in the  $k_z$  and  $k_x$  complex planes. The singularities of the integrand (i.e., corresponding either to the branch points or bound and leaky modes of the 3-D structure) are important to determine the proper numerical treatment of the integrals [16]-[20] and, hence, need to be deeply discussed in a different and appropriate framework. For this type of calculation highly efficient acceleration schemes may be used to assure a fast convergence of the LSs, in particular when complex values of  $k_z$  and  $k_x$  are required [26].

### III. SCATTERING FROM THE WOODPILE PERIODIC STRUCTURE – DERIVATION OF REFLECTION AND TRANSMISSION MATRICES

Let us assume that the  $z$  components of the incident electric and magnetic fields are expressed as follows:

$$E_z^{inc} = E_0 e^{-j(k_{x0}x - k_{y,00}y + k_{z0}z)} \quad (23)$$

$$\tilde{H}_z^{inc} = \tilde{H}_0 e^{-j(k_{x0}x - k_{y,00}y + k_{z0}z)} \quad (24)$$

where

$$E_0 = -\cos \psi, \quad \tilde{H}_0 = \sin \psi \quad (25)$$

with  $k_{x0} = -k_0 \sin \theta \cos \varphi$ ,  $k_{z0} = k_0 \sin \theta \sin \varphi$ ,  $k_{y,00} = k_0 \cos \theta$ ,  $\tilde{H}_z^{inc}$  is the normalized incident magnetic field,  $(\theta, \varphi)$  is the angle of incidence and  $\psi$  is the polarization angle of the incident electric field as depicted in Fig. 3. Since the 3-D woodpile structure is periodic in both the  $x$  and  $z$  directions, the incident field is scattered into a set of space harmonics varying as  $e^{-jk_{x\ell}x - jk_{zm}z}$ . Here  $\ell = \pm L$  and  $m = \pm M$  denote the orders of truncation of the space harmonics. Note that in the 2-D problem [26], [27], we do not have field variations along the  $z$ -axis. We formulate first the reflection and transmission matrices for the structures periodic along the  $x$ - and  $z$ -axis, separately. The results are then used to derive the reflection and transmission matrices for a single-layered woodpile periodic structure (Fig. 4). Finally, the generalized reflection and transmission matrices for the multilayered system along the transverse direction ( $y$  - axis) are obtained.

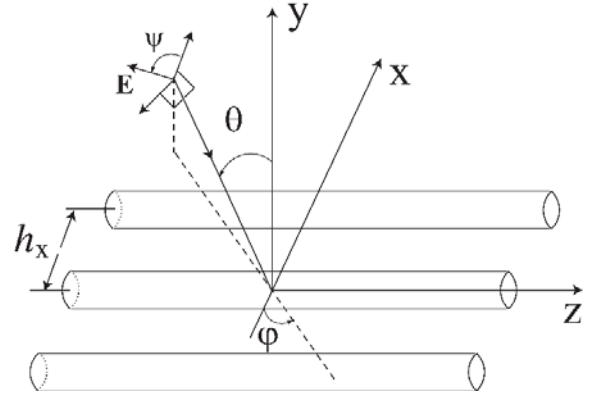


Fig. 3. Plane waves impinging on a structure composed of circular cylinders periodically distributed along the  $x$ -axis.

#### A. Reflection and transmission matrices for a single-layered structure periodic along the $x$ -axis

The  $z$ -oriented cylinders are periodically distributed along the  $x$ -axis with a period  $h_x$ . The scattering problem is formulated by employing  $E_z$  and  $\tilde{H}_z$  as the leading fields. The amplitude vectors of the incident electric and magnetic fields are denoted by  $\mathbf{e}_z^{inc} = \{e_{z,\ell,m}^{inc}\}$  and  $\tilde{\mathbf{h}}_z^{inc} = \{\tilde{h}_{z,\ell,m}^{inc}\}$ , respectively. Note that the size of the amplitude vectors is  $(2M+1)$  times larger than that for the 2-D problem [26], [27]. This is because the scattering amplitude vectors should be calculated for each space-harmonic  $m$ , associated with the field variations along the  $z$ -axis due to the crossed nature of the structure. The size of each sub-vector is  $(2L+1)$ . Thus, the amplitude column vectors of the incident fields can be written in the following form:

$$\mathbf{e}_z^{inc}(m) = \{e_{z,-L,m}, e_{z,-L+1,m} \cdots e_{z,0,m} \cdots e_{z,L-1,m}, e_{z,L,m}\}^T \quad (26)$$

$$\tilde{\mathbf{h}}_z^{inc}(m) = \{\tilde{h}_{z,-L,m}, \tilde{h}_{z,-L+1,m} \cdots \tilde{h}_{z,0,m} \cdots \tilde{h}_{z,L-1,m}, \tilde{h}_{z,L,m}\}^T$$

where “ $T$ ” represents the transpose of the vector. Following a similar analytical procedure as for the 2-D problem [26], [27], the amplitude vectors of the reflected  $\{\mathbf{e}_z^r(m), \tilde{\mathbf{h}}_z^r(m)\}$  and transmitted  $\{\mathbf{e}_z^t(m), \tilde{\mathbf{h}}_z^t(m)\}$  waves are related to the incident waves  $\{\mathbf{e}_z^{inc}(m), \tilde{\mathbf{h}}_z^{inc}(m)\}$  through the following expressions:

$$\begin{Bmatrix} \mathbf{e}_z^r \\ \tilde{\mathbf{h}}_z^r \end{Bmatrix} = \begin{bmatrix} \mathbf{R}_z^{(EE)} & \mathbf{R}_z^{(EH)} \\ \mathbf{R}_z^{(HE)} & \mathbf{R}_z^{(HH)} \end{bmatrix} \begin{Bmatrix} \mathbf{e}_z^{inc} \\ \tilde{\mathbf{h}}_z^{inc} \end{Bmatrix} \quad (27)$$

$$\begin{Bmatrix} \mathbf{e}_z^t \\ \tilde{\mathbf{h}}_z^t \end{Bmatrix} = \begin{bmatrix} \mathbf{F}_z^{(EE)} & \mathbf{F}_z^{(EH)} \\ \mathbf{F}_z^{(HE)} & \mathbf{F}_z^{(HH)} \end{bmatrix} \begin{Bmatrix} \mathbf{e}_z^{inc} \\ \tilde{\mathbf{h}}_z^{inc} \end{Bmatrix} \quad (28)$$

Here  $\mathbf{R}_z^{(\nu)}(m)$  and  $\mathbf{F}_z^{(\nu)}(m)$  ( $\nu = EE, EH, HE, HH$ ) are the reflection and transmission diagonal matrices of the periodic structure along the  $x$ -axis. For each space-harmonic  $m$ , the reflection and transmission matrices can be written in the following form:

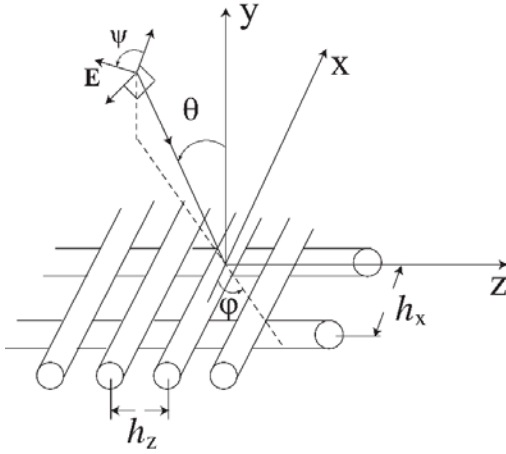


Fig. 4. Plane waves impinging on a single-layered woodpile periodic structure.

$$\mathbf{R}_z^{(\nu)}(m) = \mathbf{W}_z^- \cdot \mathbf{U}_z^+(m) \cdot \bar{\mathbf{T}}_z(m) \cdot \mathbf{P}_z(m) \cdot \mathbf{W}_z^+ \quad (\nu = EE, EH, HE, HH) \quad (29)$$

$$\mathbf{F}_z^{(\nu)}(m) = \mathbf{W}_z^- \cdot [\mathbf{I} + \mathbf{U}_z^-(m) \cdot \bar{\mathbf{T}}_z(m) \cdot \mathbf{P}_z(m)] \cdot \mathbf{W}_z^+ \quad (\nu = EE, HH) \quad (30)$$

$$\mathbf{F}_z^{(\nu)}(m) = \mathbf{W}_z^- \cdot \mathbf{U}_z^-(m) \cdot \bar{\mathbf{T}}_z(m) \cdot \mathbf{P}_z(m) \cdot \mathbf{W}_z^+ \quad (\nu = EH, HE) \quad (31)$$

with

$$\mathbf{P}_z(m) = \left[ (-j)^s e^{-js \cos^{-1}(k_{xq}/\sqrt{k_0^2 - k_{zm}^2})} \right] \quad (32)$$

$$\mathbf{U}_z^\pm(m) = \left[ \frac{2j^s}{h_x k_{y,\ell m}} e^{\mp js \cos^{-1}(k_{x\ell}/\sqrt{k_0^2 - k_{zm}^2})} \right] \quad (33)$$

$$\mathbf{W}_z^\pm = [e^{\pm jk_{x\ell} w_x}] \quad (34)$$

$$\bar{\mathbf{T}}_z(m) = \begin{bmatrix} \mathbf{T}_z^{(1)}(m) & \mathbf{T}_z^{(2)}(m) \\ \mathbf{T}_z^{(3)}(m) & \mathbf{T}_z^{(4)}(m) \end{bmatrix} \quad (35)$$

Note that  $\mathbf{P}_z(m)$  is a matrix which transforms the downgoing  $q$ -th incident space-harmonic wave to the  $s$ -th cylindrical harmonic wave for each of the indicated space-harmonic  $m$ ,  $\mathbf{U}_z^\pm(m)$  are matrices that transform the  $s$ -th cylindrical harmonic wave back to the upgoing and downgoing  $\ell$ -th space-harmonic waves varying as  $e^{-jk_{x\ell} x \pm jk_{y,\ell m} y}$ ,  $\mathbf{W}_z^\pm$  is a diagonal matrix that denotes the shift of the layer from the global origin. The matrix  $\bar{\mathbf{T}}_z(m)$  is expressed in terms of the 3-D T-matrix (see Appendix) of the  $z$ -directed cylinder for obliquely incident plane waves and the one-dimensional LSs, whose coefficients are given as [26],[27],[32]:

$$L_z^q(m) = \sum_{n=1}^{\infty} H_q^{(2)}[n\kappa(m)h_x] \left[ e^{-jn\kappa(m)h_x} + (-1)^q e^{jn\kappa(m)h_x} \right] \quad (36)$$

where  $\kappa(m) = \sqrt{k_0^2 - k_{zm}^2}$  and  $H_q^{(2)}$  is the Hankel function of the second kind. Recently, we have developed a formulation based on the Ewald method for a fast and accurate calculation of the LSs applicable to both real and complex-valued wavenumbers [26]. Note that when  $w_x = 0$ ,  $\varphi = 0$  and  $m = 0$  (no field variation along the  $z$ -axis), (29)-(35) directly reduce to (6)-(9) of the 2-D problem [26].

### B. Reflection and transmission matrices for a single-layered structure periodic along the $z$ -axis

When the  $x$ -directed cylinders are periodically distributed along the  $z$ -axis with a period  $h_z$ , employing the  $E_x$  and  $\tilde{H}_x$  fields as the leading fields, the scattering problem can be formulated in a similar way as in the previous case. Amplitude vectors of reflected  $\{\mathbf{e}_x^r(\ell), \tilde{\mathbf{h}}_x^r(\ell)\}$  and transmitted  $\{\mathbf{e}_x^t(\ell), \tilde{\mathbf{h}}_x^t(\ell)\}$  waves are related to the incident waves  $\{\mathbf{e}_x^{inc}, \tilde{\mathbf{h}}_x^{inc}\}$  as follows:

$$\begin{bmatrix} \mathbf{e}_x^r \\ \tilde{\mathbf{h}}_x^r \end{bmatrix} = \begin{bmatrix} \mathbf{R}_x^{(EE)} & \mathbf{R}_x^{(EH)} \\ \mathbf{R}_x^{(HE)} & \mathbf{R}_x^{(HH)} \end{bmatrix} \begin{bmatrix} \mathbf{e}_x^{inc} \\ \tilde{\mathbf{h}}_x^{inc} \end{bmatrix} \quad (37)$$

$$\begin{bmatrix} \mathbf{e}_x^t \\ \tilde{\mathbf{h}}_x^t \end{bmatrix} = \begin{bmatrix} \mathbf{F}_x^{(EE)} & \mathbf{F}_x^{(EH)} \\ \mathbf{F}_x^{(HE)} & \mathbf{F}_x^{(HH)} \end{bmatrix} \begin{bmatrix} \mathbf{e}_x^{inc} \\ \tilde{\mathbf{h}}_x^{inc} \end{bmatrix} \quad (38)$$

where

$$\mathbf{R}_x^{(\nu)}(\ell) = \mathbf{W}_x^- \cdot \mathbf{U}_x^+(\ell) \cdot \bar{\mathbf{T}}_x(\ell) \cdot \mathbf{P}_x(\ell) \cdot \mathbf{W}_x^+ \quad (\nu = EE, EH, HE, HH) \quad (39)$$

$$\mathbf{F}_x^{(\nu)}(\ell) = \mathbf{W}_x^- \cdot [\mathbf{I} + \mathbf{U}_x^-(\ell) \cdot \bar{\mathbf{T}}_x(\ell) \cdot \mathbf{P}_x(\ell)] \cdot \mathbf{W}_x^+ \quad (\nu = EE, HH) \quad (40)$$

$$\mathbf{F}_x^{(\nu)}(\ell) = \mathbf{W}_x^- \cdot \mathbf{U}_x^-(\ell) \cdot \bar{\mathbf{T}}_x(\ell) \cdot \mathbf{P}_x(\ell) \cdot \mathbf{W}_x^+ \quad (\nu = EH, HE) \quad (41)$$

with

$$\mathbf{P}_x(\ell) = \left[ (-j)^s e^{-js \cos^{-1}(k_{xq}/\sqrt{k_0^2 - k_{x\ell}^2})} \right] \quad (42)$$

$$\mathbf{U}_x^\pm(\ell) = \left[ \frac{2j^s}{h_z k_{y,\ell m}} e^{\mp js \cos^{-1}(k_{zm}/\sqrt{k_0^2 - k_{x\ell}^2})} \right] \quad (43)$$

$$\mathbf{W}_x^\pm = [e^{\pm jk_{zm} w_z}] \quad (44)$$

### C. Reflection and transmission matrices for a crossed structure

When the periodic arrays along the  $z$ -axis and  $x$ -axis are stacked as shown in Fig. 4, thus forming a woodpile structure, the use of common leading fields is required to describe the multiple

scattering process. Without loss of generality, we choose  $E_z$  and  $\tilde{H}_z$  as the leading fields and express the reflection and transmission matrices in (39)-(41) in terms of these field components [33],[34]. Consequently, following this process results in a new set of reflection  $\tilde{\mathbf{R}}_x^{(\nu)}$  and transmission  $\tilde{\mathbf{F}}_x^{(\nu)}$  matrices. This can be accomplished using the formulas from electromagnetics, where the transverse components of the electric and magnetic fields are expressed in terms of the z-components of the fields [33],[34]. Finally, the generalized reflection and transmission matrices for the  $N$ -layered woodpile periodic structure (Fig. 1) can be easily obtained based on the recursive algorithm using  $(\tilde{\mathbf{R}}_x^{(\nu)}, \tilde{\mathbf{F}}_x^{(\nu)})$  and  $(\mathbf{R}_z^{(\nu)}, \mathbf{F}_z^{(\nu)})$  concatenating the reflection and transmission matrices corresponding to the  $N$ -layers along the y-axis [24],[26],[27],[29].

#### IV. EVALUATION OF THE RADIATED FAR FIELD THROUGH THE RECIPROCITY THEOREM

The problem of calculating the far-field radiated in an arbitrary direction  $(\theta, \varphi)$  by a z-directed HED, can be reduced by means of the reciprocity theorem to that of calculating the z-polarized electric field produced at the source location by properly polarized plane waves, i.e., the leading fields  $E_z(x, y, z)$  and  $\tilde{H}_z = \sqrt{(\mu_0 / \varepsilon_0)} H_z$ , impinging on the periodic structure from the same direction [35], [36]. The z-component of the electric field  $E_z$  at the global origin due to the incidence of the plane waves with  $E_z^{inc} \neq 0$ ,  $\tilde{H}_z^{inc} = 0$  and  $\tilde{H}_z^{inc} \neq 0$ ,  $E_z^{inc} = 0$  can be written in the following form:

$$E_z(x=0, y=0, z=0) \Big|_{\tilde{H}_z^{inc}=0, E_z^{inc} \neq 0} = \sum_{m=-\infty}^{\infty} M_{mn}^{(1)} E_0^n \quad (45)$$

$$E_z(x=0, y=0, z=0) \Big|_{E_z^{inc}=0, \tilde{H}_z^{inc} \neq 0} = \sum_{m=-\infty}^{\infty} M_{mn}^{(2)} \tilde{H}_0^n$$

where the  $\mathbf{M}$  matrix is defined as:

$$\mathbf{M} = \begin{bmatrix} \mathbf{M}^{(1)} & \mathbf{M}^{(2)} \\ \mathbf{M}^{(3)} & \mathbf{M}^{(4)} \end{bmatrix} = \bar{\mathbf{A}}_1 \cdot \left( \bar{\mathbf{I}} - \bar{\mathbf{R}}_{N_1} \cdot \bar{\mathbf{A}} \cdot \bar{\mathbf{R}}_{N_2} \cdot \bar{\mathbf{A}} \right)^{-1} \cdot \tilde{\mathbf{F}}_{N_1} + \bar{\mathbf{A}}_2 \cdot \bar{\mathbf{R}}_{N_2} \cdot \bar{\mathbf{A}} \cdot \left( \bar{\mathbf{I}} - \bar{\mathbf{R}}_{N_1} \cdot \bar{\mathbf{A}} \cdot \bar{\mathbf{R}}_{N_2} \cdot \bar{\mathbf{A}} \right)^{-1} \cdot \tilde{\mathbf{F}}_{N_1} \quad (46)$$

It should be mentioned that (46) can be obtained using (10)-(18) noting that an amplitude of the incident field impinging on the woodpile EBG structure from the free space is known. Incident amplitude vectors of the electric field and the magnetic field are  $\{E_0^n\} = (0 \cdots E_0 \cdots 0)^T$  and  $\{\tilde{H}_0^n\} = (0 \cdots \tilde{H}_0 \cdots 0)^T$ , respectively (see (23)-(25) in Section III).

The summation in (45) is taken over the scattered space-harmonics of matrix  $\mathbf{M}$  associated with the periodicity along both the z- and x-axis. Note that the block matrices  $\mathbf{M}^{(3)}$  and  $\mathbf{M}^{(4)}$  should be used to calculate the magnetic field at the global origin. However, since we are dealing here with the HED, only block matrices  $\mathbf{M}^{(1)}$  and  $\mathbf{M}^{(2)}$  are utilized. We also note that  $\tilde{\mathbf{F}}_{N_1}$  is the generalized transmission matrix of the woodpile EBG structure viewed from the perspective of the down-going space harmonics. In the general case,  $\tilde{\mathbf{F}}_{N_1}$  is not equal to  $\bar{\mathbf{F}}_{N_1}$  in (17), unless the structure is symmetric for up-going and down-going space harmonics.

#### V. NUMERICAL RESULTS AND DISCUSSIONS

The method described in Section II is applied to three numerical case studies. All the selected structures are typical examples of woodpile cavity antennas radiating at broadside with good radiation performances. In all the numerical results, the truncation of the reflection and the transmission matrices associated with the EBG structures are taken to be the same for the space-harmonics along both the x-axis and z-axis. In particular, the truncation order is equal to  $M = L = \pm 4$ , which means that the size of each block-matrix in (27), (28) and (37), (38) is  $81 \times 81$ . The numerical results that follow are obtained with the energy conservation of less than  $10^{-6}$ . We have proved the conservation of energy using the rigorous expressions of the reflection and transmission matrices given in Section III. The normalized radiation patterns in the E-plane ( $\varphi = 90^\circ$ ), in the H-plane ( $\varphi = 0^\circ$ ), and in cut planes apart from the principal planes versus angle  $\theta$ , obtained using (22) together with (19) and (20), are displaced. For comparison and validation, the

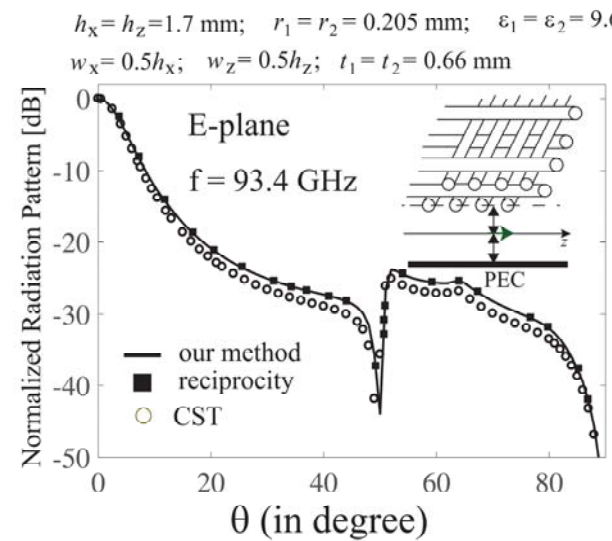


Fig. 5. Normalized radiation pattern (in dB) for a HED coupled to the 2-layered woodpile periodic structures in the E-plane obtained with the developed method (solid line), reciprocity (squares) and CST (circles). The operating frequency is  $f=93.4$  GHz and the structural parameters are given at the top of the figure.

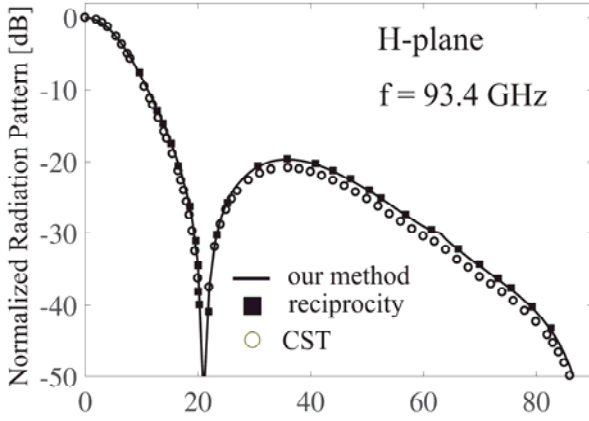


Fig. 6. The same as in Fig.5 but for the H-plane.

normalized radiation patterns using the reciprocity theorem with the proposed approach, as described in Section IV, and with independent simulations with CST are also shown. An important point of worth mentioning is a comparison of the computation times based on our method and CST. The computation time spent to calculate the normalized radiation patterns in the E-plane and in the H-plane at two extreme angles,  $\theta=0^\circ$  and  $\theta=89^\circ$  for all three numerical examples, is summarized in Table 1. The proposed method is considerably faster than the CST commercial software, showing a speed improvement of one to two orders of magnitudes. It is worth mentioning that the computation time using the reciprocity theorem (Section IV) is almost the same as that based on the ASM and the steepest descent method discussed in Section II. This is not surprising, since both approaches are based on the same formalism, namely, on the implementation of the reflection and transmission matrices and the incident source fields. The tests have been performed on Intel Xeon E5-2643 at 3.4 GHz with 384 GB of RAM.

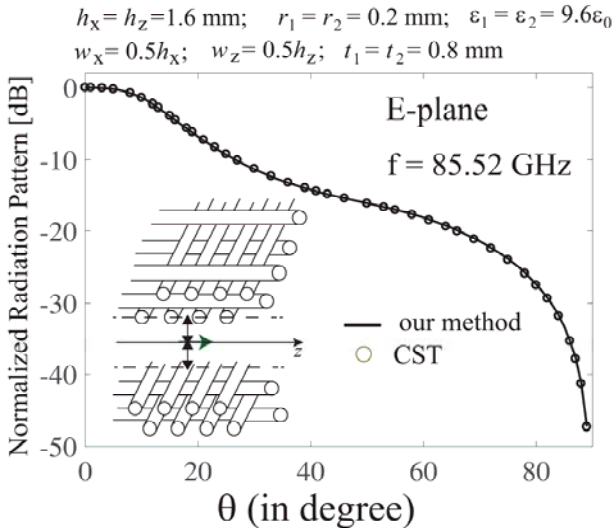


Fig. 7. Normalized radiation pattern (in dB) for a HED coupled to the 2-layered woodpile periodic structures in the E-plane obtained with the developed method (solid line) and CST (circles). The operating frequency is  $f=85.52$  GHz and the structural parameters are given at the top of the figure.

#### Case 1.

As a first example, we calculate the normalized radiation patterns of a HED coupled to a 2-layered woodpile EBG structure and PEC ground plane. The geometrical parameters are:  $h_x=h_z=1.7$  mm,  $r_1=r_2=0.205$  mm,  $w_x=0.5h_x$ ,  $w_z=0.5h_z$ ,  $\epsilon_1=\epsilon_2=9.6\epsilon_0$ ,  $t_1=t_2=0.66$  mm and  $f=93.4$  GHz (operating frequency). Note that the third and the fourth layers are shifted by half of the period ( $w_x=0.5h_x$  and  $w_z=0.5h_z$ ) with respect to the first and the second layers, thus forming a woodpile structure. The reflection matrix for the PEC ground plane can be easily implemented in our analysis assuming that only its diagonal elements are equal to -1, where all other matrix elements are zero. The normalized radiation patterns in the E-plane and H-plane versus the angle  $\theta$  using our method (solid line), the reciprocity theorem (squares) and CST (circles) are shown in Figs. 5 and 6, respectively.

#### Case 2.

In the second example we removed the PEC and put the woodpile structure as a substrate. We calculate the normalized radiation patterns of a HED coupled to a 2-layered woodpile EBG structures with the following parameters:  $h_x=h_z=1.6$  mm,  $r_1=r_2=0.2$  mm,  $w_x=0.5h_x$ ,  $w_z=0.5h_z$ ,  $\epsilon_1=\epsilon_2=9.6\epsilon_0$ ,  $t_1=t_2=0.8$  mm and  $f=85.52$  GHz. The results in the E-plane and H-plane, and in cut planes at  $\varphi=30^\circ$  and  $\varphi=45^\circ$  are shown in Figs. 7-9, respectively. We note that cross-polarization is present in the oblique cuts, i.e., at  $\varphi=30^\circ$  and  $\varphi=45^\circ$ , where both the leading fields  $E_z$  and  $\tilde{H}_z$  contribute to the total radiated power. An excellent agreement can be observed between the normalized radiation patterns in all cases. Regarding the computation time to calculate the normalized radiation pattern per one angle  $\theta$ , our method is about 16 times faster than CST for the E-plane and 16 times faster than CST for the H-plane.

#### Case 3.

In the third example we present the results for the normalized radiation patterns of a HED coupled to a 2-layered crossed EBG structure with the following parameters:  $h_x=h_z=1.6$  mm,  $r_1=r_2=0.2$  mm,  $w_x=0$ ,  $w_z=0$ ,  $\epsilon_1=\epsilon_2=9.6\epsilon_0$ ,  $t_1=t_2=0.25$  mm and  $f=110.86$  GHz. The results in the E-plane and H-plane are

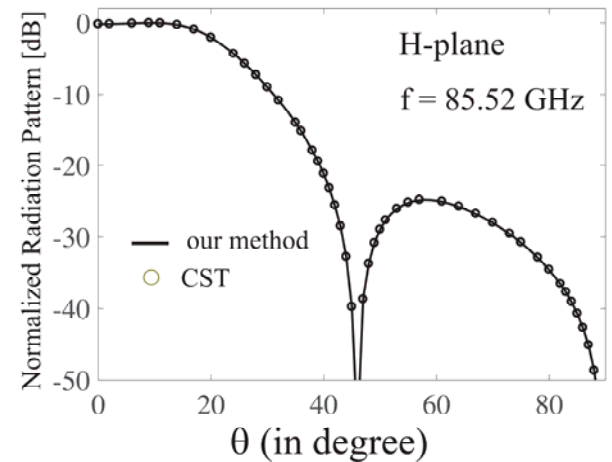


Fig. 8. The same as in Fig. 7 but for the H-plane.

shown in Fig. 10 and Fig. 11, respectively. The geometry of the problem is similar to Case 2 (but  $w_x=0, w_z=0$ ), however this is a challenging case, since the operating values  $h_{x,z}/\lambda_0$ , with  $\lambda_0$  the free-space wavelength, are much higher in comparison to the previous two cases, where only the fundamental space harmonic ( $\ell = m = 0$ ) becomes the propagating mode. In the present example, the higher-order space harmonic, namely the first-order space-harmonic ( $\ell = -1, m = 0$ ) in the E-plane and ( $\ell = 0, m = 1$ ) in the H-plane, starts to propagate at around  $\theta = 44^\circ$ . Nevertheless, a good agreement is observed between our formalism and the CST analysis for both principles planes, whereas our method is about 75 times faster than CST (see Table 1).

Finally, we note that the truncation number is mainly affected by the operating frequency. Computation times in Table 1 clearly confirms this. In Case 3 the operating frequency is very high (110.86 GHz) and therefore, the computation time

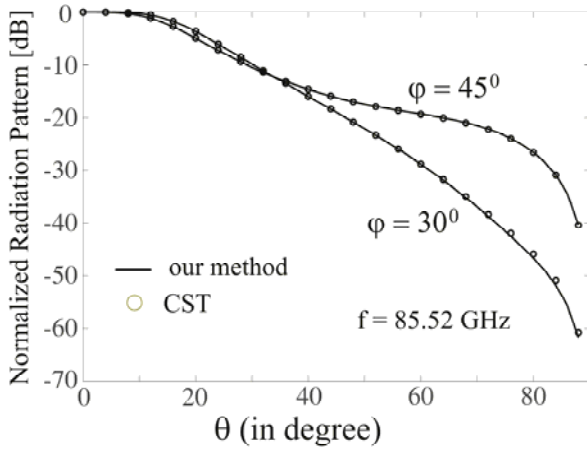


Fig. 9. The same as in Fig. 7 but in the cut planes at  $\varphi = 30^\circ$  and  $\varphi = 45^\circ$ .

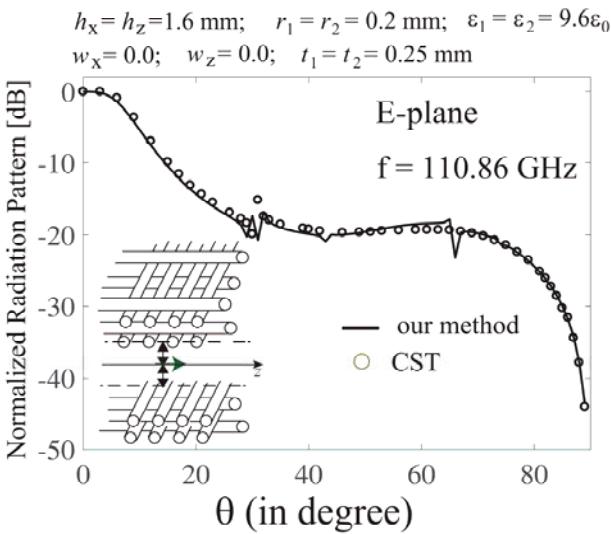


Fig. 10. Normalized radiation pattern (in dB) for a HED coupled to the 2-layered woodpile periodic structures in the E-plane obtained with the developed method (solid line) and CST (circles). The operating frequency is  $f=110.86$  GHz.

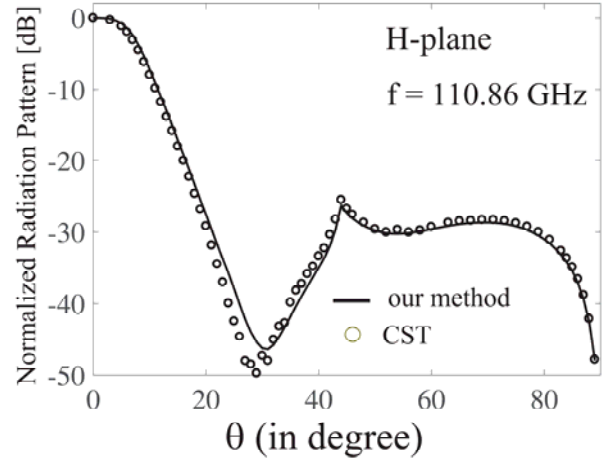


Fig. 11. The same as in Fig. 10 but for the H-plane.

is about 3 times higher than that in Case 2, where the operating frequency is 85.52 GHz. An increase of the layers and radius of the rods also effect the truncation number; however, the effect of the operating frequency is more substantial.

TABLE I  
COMPARISON OF COMPUTATION TIME TO CALCULATE THE NORMALIZED RADIATION PATTERNS IN E-PLANE AND H-PLANE AT  $\theta = 0^\circ$  AND  $\theta = 89^\circ$  FOR THREE NUMERICAL EXAMPLES USING THE PROPOSED METHOD IN SECTION II AND CST.

|        | $\theta$   | E-plane (Time [sec]) |     | H-plane (Time [sec]) |     |
|--------|------------|----------------------|-----|----------------------|-----|
|        |            | Proposed Method      | CST | Proposed Method      | CST |
| Case 1 | $0^\circ$  | 0.80                 | 24  | 0.75                 | 25  |
|        | $89^\circ$ | 0.95                 | 22  | 0.85                 | 21  |
| Case 2 | $0^\circ$  | 2.15                 | 39  | 2.05                 | 43  |
|        | $89^\circ$ | 2.50                 | 40  | 2.10                 | 34  |
| Case 3 | $0^\circ$  | 6.10                 | 583 | 6.60                 | 578 |
|        | $89^\circ$ | 6.90                 | 538 | 7.20                 | 546 |

## VI. CONCLUSION

The far-field radiated by a localized dipole source coupled to dielectric woodpile EBG structures, periodic in two dimensions and finite in the third one, has been investigated using a rigorous and self-contained semi-analytical approach. The spectral response of the 3-D EBG structure composed of multilayered crossed arrays of circular cylinders has been obtained using the transition matrix of the cylindrical objects and the lattice sums technique. A single crossed-array element of a woodpile structure is formed by concatenating two arrays whose cylinders' axes are orthogonal to each other. The adopted approach is very general since the period, as well as the geometrical and physical parameters of the cylindrical elements can be different along each layer. Radiation from an elementary dipole source coupled to three different configurations of woodpile structures has been investigated using the proposed

formalism. All the selected structures are typical examples of woodpile cavity antennas having good radiation performance at broadside. The obtained radiation patterns have been fully validated by using reciprocity and CST Microwave Studio. The proposed approach is very robust and exhibits high accuracy and exceptional efficiency, with simulation times from one to two orders lower than the commercial software. Thus, it can be considered as a very effective tool for the design of dielectric 3-D EBG cavity antennas.

#### ACKNOWLEDGEMENT

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#### APPENDIX

In the Appendix the details for 3-D T-matrix of the z-directed cylinder for obliquely incident plane waves is given. Consider a single periodic array of circular cylinders as shown in Fig. 3. The cylinders are parallel to each other and periodically spaced with a distance  $h_x$  along the x-direction (no field variations along the z-axis). The center of the  $q$ -th cylinder is located at  $x=qh_x$  and  $y=0$  and its local polar coordinates are  $\rho_q = \sqrt{[(x-qh_x)^2 + y^2]}$  and  $\cos\phi_q = (x-qh_x)/\rho_q$ . Using the Floquet theorem, the scattered fields may be expressed in the following form:

$$E_z^{sca} = e^{-jk_z z} \sum_{q=-\infty}^{\infty} e^{-jk_x qh_x} \Psi_q^T \cdot \mathbf{e}_z^{sca} \quad (47)$$

$$\tilde{H}_z^{sca} = e^{-jk_z z} \sum_{q=-\infty}^{\infty} e^{-jk_x qh_x} \Psi_q^T \cdot \tilde{\mathbf{h}}_z^{sca} \quad (48)$$

with

$$\Psi_q = \sum_{n=-\infty}^{\infty} H_n^{(2)}(\kappa\rho_q) e^{-jn\phi_q} \quad (49)$$

where  $\kappa = \sqrt{k_0^2 - k_z^2}$ . Here  $\mathbf{e}_z^{sca}$  and  $\tilde{\mathbf{h}}_z^{sca}$  are the column vectors, whose components represent the unknown amplitudes of the scattered fields from the 0-th cylinder. Using the addition theorem for cylindrical functions [37], the total fields outside the cylinders can be written as follows:

$$E_z^{total} = e^{-jk_z z} \left( \left\{ \Phi_0^T \cdot (\mathbf{e}_z^{inc} + \mathbf{L}_z \cdot \mathbf{e}_z^{sca}) \right\} + \Psi_0^T \cdot \mathbf{e}_z^{sca} \right) \quad (50)$$

$$\tilde{H}_z^{total} = e^{-jk_z z} \left( \left\{ \Phi_0^T \cdot (\tilde{\mathbf{h}}_z^{inc} + \mathbf{L}_z \cdot \tilde{\mathbf{h}}_z^{sca}) \right\} + \Psi_0^T \cdot \tilde{\mathbf{h}}_z^{sca} \right) \quad (51)$$

with

$$\Phi_0(m) = \sum_{n=-\infty}^{\infty} J_n(\kappa\rho_0) e^{-jn\phi_0} \quad (52)$$

and  $\mathbf{L}$  represents the lattice sums defined by (36). The first term on the right hand side of (50) and (51) can be viewed as an

incident field impinging on the 0-th cylinder, whereas the second term corresponds to the scattered field from the 0-th cylinder. Thus, the relation between the scattered and incident amplitudes can be expressed using the 3-D T-matrix of the circular cylinder in isolation:

$$\begin{Bmatrix} \mathbf{e}_z^{sca} \\ \tilde{\mathbf{h}}_z^{sca} \end{Bmatrix} = \bar{\mathbf{T}}_z \begin{Bmatrix} \mathbf{e}_z^{inc} \\ \tilde{\mathbf{h}}_z^{inc} \end{Bmatrix} = \begin{bmatrix} \mathbf{T}_z^{(1)} & \mathbf{T}_z^{(2)} \\ \mathbf{T}_z^{(3)} & \mathbf{T}_z^{(4)} \end{bmatrix} \begin{Bmatrix} \mathbf{e}_z^{inc} \\ \tilde{\mathbf{h}}_z^{inc} \end{Bmatrix} \quad (53)$$

At  $\varphi = 0$ ,  $\mathbf{T}_z^{(2)} = \mathbf{T}_z^{(3)} = 0$  and the problem is reduced to a 2-D problem with a separate T-matrix for TE and TM waves [26], [27]. It should be noted that for the crossed structure,  $\mathbf{T}_z^{(q)}$  and  $\mathbf{L}_z$  should be functions of  $m$ , which represents the space-harmonic of the field along the z-axis. The calculation procedure is similar when the cylinders are parallel to each other and periodically spaced with a distance  $h_z$  along the z-direction (no field variations along the x-axis). As a result, the expression for  $\mathbf{T}_x^{(q)}$  that appears in (39)-(41) can be derived.

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