Method-of-Moments Solver for All-Metal Corrugated Structures: Leaky-Mode Analysis and Open-Stopband Suppression

Dejian Zhang, *Member, IEEE*, Davide Comite, *Senior Member, IEEE*, Paolo Baccarelli, *Member, IEEE*, Xiaojiao Deng, Xiaoping Zheng, Alessandro Galli, *Member, IEEE*, and Paolo Burghignoli, *Senior Member, IEEE*

***Abstract*— A method-of-moments (MoMs) approach is proposed for the modal analysis of two-dimensional (2-D) all-metal leaky-wave structures periodic along one direction. The electric-field integral equations (EFIEs) for both TM and TE polarizations are formulated in the unit cell (UC) by employing a rapidly convergent Ewald representation for the relevant 2-D periodic Green's function. The integral equations are discretized using triangular subsectional basis functions and a Galerkin testing scheme. Effective modal dispersive analyses are developed for a variety of leaky-wave structures. Real proper (bound) modes, as well as complex proper (backward leaky) and improper (forward leaky) modes, are investigated. The proposed method allows for the accurate design and tuning of asymmetric UCs required to suppress the open stopband (OSB), which cannot be easily carried out by means of full-wave simulations with commercial software. Validations are provided through alternative methods and in all cases the presented formulation shows high accuracy and versatility.**

***Index Terms*— All-metal antennas, corrugated structures, leaky wave antennas (LWAs), method-of-moments (MoMs).**

I. Introduction

CORRUGATED all-metal structures periodic along one direction are a key constituent of a variety of radiating and waveguiding devices thanks to their rich and versatile electromagnetic response. Prominent examples of corrugated structures have been proposed since the 1940s to enhance the field confinement in the vicinity of metal plates even assuming an ideally infinite conductivity, thus allowing for the realization of *surface waveguides* and *antennas* [1].

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Dejian Zhang, Xiaojiao Deng, and Xiaoping Zheng are with the Department of Automation, Beijing National Research Center for Information Science and Technology, Tsinghua University, Beijing 100084, China (e-mail: asean@mail.tsinghua.edu.cn).

Davide Comite, Alessandro Galli, and Paolo Burghignoli are with the Department of Information Engineering, Electronics and Telecommunications, Sapienza University of Rome, 00184 Rome, Italy (e-mail: davide.comite/etc@uniroma1.it).

Paolo Baccarelli is with the Department of Industrial, Electronic, and Mechanical Engineering, "Roma Tre" University of Rome, 00146 Rome, Italy (e-mail: paolo.baccarelli@uniroma3.it).

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Corrugated metal surfaces have also been employed to synthesize various classes of *anisotropic impedance boundaries*, including the so-called artificially soft and hard surfaces [2]. A classical application of such boundaries is the design of circular corrugated horn antennas operating on hybrid modes, which exhibit highly symmetric beams with low cross polarization and wideband performance (see, e.g., [3]).

A third feature of the considered class of structures, which provide the main motivation for the present study, is the possibility of using them as a constituent of transversely open waveguides that can support *leaky modes*. The relatively low material losses associated with their all-metal nature indeed make them a suitable platform for the realization of both one-dimensional (1-D) and radially periodic (or bull's-eye) leaky-wave antennas (LWAs) operating at microwaves, millimeter-wave and up to terahertz [4], [5], [6].

However, periodic LWAs suffer from the presence of a well-known and longstanding open stopband (OSB) problem (see, e.g., [7], [8]), resulting in a deterioration of the radiation pattern and of the input matching of the antenna. An essential step in the OSB design of any LWA is the accurate dispersive analysis of the structure, aimed at determining the complex wavenumber of the operating leaky mode (or modes) [9]. In this connection, it is worth noting that, in the case of bull's-eye geometries, the lack of translational invariance prevents a direct modal analysis, which is customarily performed considering their linearized and two-dimensional (2-D) 1-D periodic counterparts (i.e., invariant with respect to one coordinate) [10], [11].

As concerns all-metal corrugated 1-D periodic open waveguides, various methods have been used for their modal analysis. In [12], a standard corrugated plate with rectangular grooves is analyzed with a transverse resonance approach under the assumption that only the TEM mode exists in the grooves, thus generalizing to an open configuration the approach outlined in [13] for a closed configuration. The method offers control on the spectral nature (proper or improper) of the relevant space harmonics; however, besides being limited to a single specific geometry, it is approximated, because of the abovementioned assumption, and is accurate only for groove lengths smaller than half a free-space wavelength. Such limitations are overcome by the full-wave approach adopted in [14], which tracks the maxima of the electric field inside a single unit cell (UC). The method is versatile, as it allows for having an arbitrary geometry inside the UC. It is also computationally convenient with respect to alternative approaches based on the simulation of truncated structures with a finite (but typically large) number of UCs, such as those based on the matrix pencil method [14] or on Bloch analyses (see, e.g., [15], and refs. therein). However, as with the latter approaches, it does not offer the possibility of controlling the *spectral nature* of the space harmonics [16].

Therefore, at present, there is no analytical method or commercial software able to provide the needed accurate complex wavenumbers for the effective design of all-metal 1-D periodic corrugated LWAs, including those requiring suppression of the OSB for sustained broadside radiation.

In this article, we present a novel full-wave modal solver for the analysis and design of all-metal leaky-wave structures periodic along one direction. It is based on the formulation of an electric-field integral equation (EFIE) having the surface electric current density on the metal surfaces (assumed to be perfectly conducting) as an unknown and on its numerical discretization with the method-of-moments (MoMs). The method is particularly suited for the efficient and accurate derivation of the modal complex wavenumber of all-metal 1-D periodic corrugated structures and combines for the first time the following desirable features: 1) it allows for treating arbitrary geometries of the corrugation in the UC; 2) it is computationally fast, as the EFIE is formulated in terms of a periodic Green's function accelerated with the Ewald method (which offers Gaussian convergence) [17], [18] and thus requires discretizing the metal surfaces inside a single UC; and 3) it allows for controlling the spectral nature of the space harmonics and thus tracking the solution along the various Riemann sheets of the dispersion equation as a function of the complex wavenumber. All these characteristics make the present full-wave modal solver a fundamental tool for the accurate design of a wide class of corrugated LWAs with advanced radiating features, including the bull's-eye configurations.

The article is organized as follows. In Section II the MoM formulation is presented for both the TE and TM polarizations. In Section III, modal dispersive analyses are reported for a variety of structures, including air-filled Fabry–Perot cavity (FPC) antennas and corrugated plates, validating them with alternative approaches based on homogenized models (HMs) or full-wave approximate results obtained with commercial solvers. In Section V, a corrugated 1-D LWA is investigated, whose radiative features obtained with the proposed MoM approach (by means of reciprocity) are compared with those from independent full-wave commercial software (i.e., CST Microwave Studio and FEKO). Section V discusses the capability of the proposed MoM approach to assist the design of UCs of corrugated all-metal structures with the suppressed OSB, which must be based on the *accurate* knowledge of the complex wavenumber of the leaky modes supported by the relevant open waveguide. At the present state of the art, there are no other analytical methods nor commercial software able to provide an accurate evaluation of the complex wavenumber for this kind of corrugated structures. Finally, conclusions are drawn in Section VI.

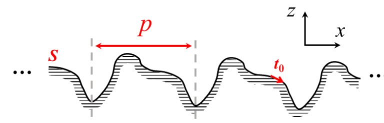

Fig. 1. All-metal corrugated structures with an arbitrary periodic profile along x-axis with period p. t_0 is the unit vector tangent to the surface profile s.

II. Electric Field Integral Equation

Consider the general all-metal corrugated structure sketched in Fig. 1, characterized by an arbitrary profile periodic along the x axis with period p and invariant along the y axis. Assuming that the field is independent of the y coordinate as well, and that the metal profile is modeled as a perfect electric conductor (PEC), the modal spectrum of the structure decouples into TE^{xz} and TM^{xz} disjoint subsets, having nonzero field components (E_y, H_x, H_z) and (H_y, E_x, E_z), respectively. By virtue of the Floquet's theorem, each TE or TM mode can be represented as a superposition of an infinite number of space harmonics with (generally complex) propagation wavenumbers $k_{xn} = \beta_{xn} - j\alpha_x = \beta_{x0} + 2\pi n/p - j\alpha_x$, where $n = 0, \pm 1, \pm 2, \ldots$

Sections II-A and II-B, the modal problem for the two polarizations is formulated in terms of an EFIE, along with the relevant discretization with the MoM.

A. Transverse Electric Case

In the TE case, the surface current $\mathbf{J}_s$ existing on the PEC boundary is y-directed: $\mathbf{J}_s = J\mathbf{y}_0$ and is independent of y, therefore no electric charge density exists on the PEC boundary. Then it yields the following EFIE which, thanks to the periodicity, can be enforced on a single UC:

$$jk_0\zeta_0 \int_s J(\mathbf{r}')G_\mathrm{p}(\mathbf{r},\mathbf{r}')d\mathbf{r}' = 0 \tag{1}$$

where k_0 and ζ_0 are the free-space wavenumber and impedance, respectively; s is the PEC contour within the UC; $\mathbf{r} \equiv (x, z)$ is the observation point; $\mathbf{r}' \equiv (x', z')$ is the source point; and the relevant 1-D periodic Green's function is

$$G_\mathrm{p}(\mathbf{r},\mathbf{r}') = \frac{1}{4j}\sum_{m=-\infty}^{\infty} H_0^{(2)}(k_0 R_m)e^{-jmk_{x0}p} \tag{2}$$

where $H_0^{(2)}(\cdot)$ is the zero-order Hankel function of the second kind; $R_m = ((z-z')^2 + (x-x'-mp)^2)^{1/2}$; k_{x0} is the propagation wavenumber of the considered TE or TM mode. However, the expression in (2) is known to be extremely slowly convergent and thus in this work, the relevant 1-D periodic Green's function is evaluated through the Ewald representation [17], [18], which is rapidly convergent for both real and complex wavenumbers k_{x0}, and is expressed in terms of modified spectral and spatial series.

As well known, the accuracy of the Ewald representation is related to the summation limits M, P, and the splitting parameter $\mathcal{E}$ [19]. Among them, M and P are the numbers of summation terms in the spatial and spectral series, respectively. Obviously, a large number of P and M is required to get a high accuracy but also a heavy computation burden. To reduce the burden, the splitting parameter $\mathcal{E}$ is thus introduced to minimize the required total number $P+M$. However, one should be very careful about the choice of $\mathcal{E}$ because an improper splitting parameter $\mathcal{E}$ might lead to cancellation errors known as "high frequency breakdown" [19]. Therefore, in this article, $M=P=15$ has been used and the splitting parameter $\mathcal{E}$ is prudently chosen from three options in [19]. Moreover, as mentioned later, an iterative algorithm based on Müller's method is applied in the MoM to solve the eigenvalue problem, and the choice of $\mathcal{E}$ is thus updated based on the iterative eigenvalue during each iteration.

The EFIE (1) is discretized with the MoM by representing the unknown current density through a set of scalar triangle basis functions $f_n(\mathbf{r})$ [20]

$$J(\mathbf{r}) = \sum_{n=1}^{N} a_n f_n(\mathbf{r}) \tag{3}$$

where N is the total number of basis functions and

$$f_n(\mathbf{r}) = \begin{cases} \|r-r_{n-1}\|/\|r_n-r_{n-1}\|, & [r_{n-1}, r_n] \\ \|r-r_{n+1}\|/\|r_{n+1}-r_n\|, & [r_n, r_{n+1}] \\ 0, & \text{otherwise} \end{cases} \tag{4}$$

where r_n is the start or endpoint for each segment. Clearly, each triangle basis function is defined on two adjacent segments and a_n are the unknown weighting coefficients. To ensure calculation speed, a nonuniform mesh is applied here to achieve satisfactory discretization accuracy with about 100 segments. A Galerkin testing scheme is applied to obtain a linear system of equations for the unknown coefficients, whose coefficient-matrix elements are

$$z_{mn} = jk_0\zeta_0 \int_{f_m} f_m(\mathbf{r}) \int_{f_n} f_n(\mathbf{r}') G_\mathrm{p}(\mathbf{r},\mathbf{r}') d\mathbf{r}' d\mathbf{r}. \tag{5}$$

Since each triangle function is defined over a pair of segments, the foregoing integrals need to be evaluated only for four segments related to f_m and f_n.

When the segments on which f_m and f_n are defined overlap, the distance between the observation point $\mathbf{r}$ and the source point $\mathbf{r}'$ is electrically small and thus the periodic Green's function G_p shows a singular behavior. The extraction of the Green's function singularity and the effective integration in the MoM approach are fundamental for the accurate convergence of the modal solver when complex wavenumbers are sought and proper and improper determinations are considered [21], [22]. In the Ewald representation, such a singular behavior is present in the spatial series only, which can be written as

$$\begin{aligned} G_\mathrm{p}(\mathbf{r},\mathbf{r}') &= \hat{G}_\mathrm{p}(\mathbf{r},\mathbf{r}') + \frac{1}{4\pi} E_1\left(R_0^2\mathcal{E}^2\right) \\ &= G_\mathrm{spectral}(\mathbf{r},\mathbf{r}') + \hat{G}_\mathrm{spatial}(\mathbf{r},\mathbf{r}') + \frac{1}{4\pi} E_1\left(R_0^2\mathcal{E}^2\right) \end{aligned} \tag{6}$$

where $E_1(\cdot)$ is the first-order exponential integral function. As shown, the singularity occurs in E_1, and it is absent in the regular term, given in the following:

$$\begin{aligned} \hat{G}_\mathrm{spatial}(\mathbf{r},\mathbf{r}') &= \frac{1}{4\pi} \sum_{q=1}^{\infty} \left(\frac{k_0}{2\mathcal{E}}\right)^{2q} \frac{1}{q!} E_{q+1}\left(R_0^2\mathcal{E}^2\right) \\ &\quad + \frac{1}{4\pi} \sum_{\substack{m=-\infty \\ (m,q)\neq(0,0)}}^{\infty} e^{-jmk_{x0}p} \\ &\quad \times \sum_{q=0}^{\infty} \left(\frac{k_0}{2\mathcal{E}}\right)^{2q} \frac{1}{q!} E_{q+1}\left(R_m^2\mathcal{E}^2\right). \end{aligned} \tag{7}$$

For computational purposes, since the length of each mesh segment is small with respect to the wavelength, the singularity can be expressed by using the small-argument approximation to the exponential integral function [17], [23, Eq. (6.6.2)]

$$E_1(x) \approx -\gamma - \ln x, \quad x \to 0 \tag{8}$$

where γ is the Euler's constant.

Therefore, for the self term where any of the segments overlap, the integral in (5) can be expressed as

$$\begin{aligned} z_{mn} &= I_1 + I_2 \\ &= jk_0\zeta_0 \int_{f_m} f_m(\mathbf{r}) \int_{f_n} f_n(\mathbf{r}') \hat{G}_\mathrm{p}(\mathbf{r},\mathbf{r}') d\mathbf{r}' d\mathbf{r} \\ &\quad + \frac{jk_0\zeta_0}{4\pi} \int_{f_m} f_m(\mathbf{r}) \int_{f_n} f_n(\mathbf{r}') \left[-\gamma - \ln\left(R_0^2\mathcal{E}^2\right)\right] d\mathbf{r}' d\mathbf{r}. \end{aligned} \tag{9}$$

where only the integral I_2 is related to the contribution of the overlapped segment. If the slope of the testing functions $f_m(\mathbf{r})$ is positive, after some algebraic manipulations, the final expression for performing numerically the last integral I_2, where the singularity terms only exist, evaluates to (cf., [20, Eq. (6.19)])

$$\begin{aligned} I_2 = -\frac{jk_0\zeta_0}{4\pi} \int_0^\Delta f_m(t) \bigg[&\frac{\Delta}{2}(\gamma + 2\ln\mathcal{E} - 1) - t \\ &+ \Delta \ln(\Delta - t) + \frac{t^2}{\Delta} \ln\left(\frac{t}{\Delta - t}\right) \bigg] dt \end{aligned} \tag{10}$$

in which Δ is the length of the overlapped segment. The substituted observation point t lies within the limits of integration and the function $f_m(t)$ is the half of the triangle having a positive or negative slope. If the slope of the testing functions is negative, due to symmetry, one can evaluate I_2 by simply replacing $f_m(t)$ with $1 - f_m(t)$ in (10).

B. Transverse Magnetic Case

In the TM case, the scattered electric field and the surface electric currents lie in the xz plane and, hence, there is a nonzero surface charge density on the PEC surface. The EFIE can be written as

$$jk_0\zeta_0 \hat{\mathbf{t}}_0(\mathbf{r}) \cdot \int_s \left[\mathbf{1} + \frac{1}{k_0^2}\nabla\nabla\right] \cdot \mathbf{J}(\mathbf{r}') G_\mathrm{p}(\mathbf{r},\mathbf{r}') d\mathbf{r}' = 0 \tag{11}$$

where the unit vector $\hat{\mathbf{t}}_0(\mathbf{r})$ is tangent to s. Similarly, the contour s is divided into a number of linear segments, and the current $\mathbf{J}(\mathbf{r})$ is expanded using vector triangle basis functions

[24], which are tangent to each segment. Using the Galerkin's method and the nonzero divergence of triangle functions [20] yield the matrix elements for the TM case

$$z_{mn} = jk_0\zeta_0 \int_{f_m} \boldsymbol{f}_m(\mathbf{r}) \cdot \int_{f_n} \boldsymbol{f}_n(\mathbf{r}')G_{\mathrm{p}}(\mathbf{r},\mathbf{r}')d\mathbf{r}' - \frac{j\zeta_0}{k_0}\int_{f_m} \nabla\cdot \boldsymbol{f}_m(\mathbf{r})\int_{f_n}\nabla'\cdot \boldsymbol{f}_n(\mathbf{r}')G_{\mathrm{p}}(\mathbf{r},\mathbf{r}')d\mathbf{r}'. \quad (12)$$

Again, special attention should be paid to overlapped terms; the series expansion is written as

$$\begin{aligned} z_{mn} &= I_1 + I_2 - I_3 - I_4 \\ &= jk_0\zeta_0 \int_{f_m} \boldsymbol{f}_m(\mathbf{r}) \cdot \int_{f_n} \boldsymbol{f}_n(\mathbf{r}')\hat{G}_{\mathrm{p}}(\mathbf{r},\mathbf{r}')d\mathbf{r}'d\mathbf{r} \\ &\quad + \frac{jk_0\zeta_0}{4\pi}\int_{f_m} \boldsymbol{f}_m(\mathbf{r}) \cdot \int_{f_n} \boldsymbol{f}_n(\mathbf{r}')\left[-\gamma - \ln\left(R_0^2\mathcal{E}^2\right)\right]d\mathbf{r}'d\mathbf{r} \\ &\quad - \frac{j\zeta_0}{k_0}\int_{f_m}\nabla\cdot \boldsymbol{f}_m(\mathbf{r})\int_{f_n}\nabla'\cdot \boldsymbol{f}_n(\mathbf{r}')\hat{G}_{\mathrm{p}}(\mathbf{r},\mathbf{r}')d\mathbf{r}'d\mathbf{r} \\ &\quad - \frac{j\zeta_0}{4\pi k_0}\int_{f_m}\nabla\cdot \boldsymbol{f}_m(\mathbf{r})\int_{f_n}\nabla' \\ &\quad \cdot \boldsymbol{f}_n(\mathbf{r}')\left[-\gamma - \ln\left(R_0^2\mathcal{E}^2\right)\right]d\mathbf{r}'d\mathbf{r} \end{aligned} \quad (13)$$

where both the integral I_2 and the integral I_4 are related to the contribution of the overlapped segment. Since the integral I_2 is similar to the integral indicated with the same symbol occurring in the TE case, the same expression in (10) can be used for its numerical calculation. As for the integral I_4, with (8) it results (cf., [20, Eq. (6.12)])

$$I_4 = \pm\frac{j\zeta_0}{4\pi k_0\Delta}\int_0^{\Delta}\left[-(\gamma + 2\ln\mathcal{E} - 2) - \frac{2t}{\Delta}\ln\frac{t}{\Delta - t} - 2\ln(\Delta - t)\right]dt \quad (14)$$

in which the sign depends on the slope of the basis and testing functions on each segment.

After transforming original continuous-domain problems in (1) and (11) to the discrete forms in (5) and (12), and extracting the singular terms using (9) and (13), the final matrix equation can be obtained for both TM and TE cases

$$[\mathbf{Z}][\mathbf{J}] = \mathbf{0} \quad (15)$$

where $[\mathbf{J}]$ is the unknown surface current vector and $[\mathbf{Z}]$ is the impedance matrix with the coefficient-matrix element z_{mn}. Obviously, the only unknowns in z_{mn} are the *propagation wavenumbers* k_{x0} of the TE or TM modes. To get enough accuracy, the regular integrals in z_{mn} are evaluated using Gaussian quadrature rule with four integration points in all cases, whereas the integrals related to the singularity are evaluated using the analytical expression in (14) (or (10) in the TE case).

A final remark concerns the case in which, in (5) and (12), either f_m or f_n are associated with an edge lying on the unit-cell boundaries. They should carefully be handled to ensure that the current flows correctly from one unit cell to the next. Thanks to Floquet's theorem and the symmetry imposed on the mesh at the UC boundaries, the half-triangle attached to one border can be shifted by a factor $e^{\pm jk_{x0}p}$ with respect to the opposite border, where the $\pm$ sign refers to backward or forward shift, respectively. In this way, the domains of the basis and the test function can be completed [21].

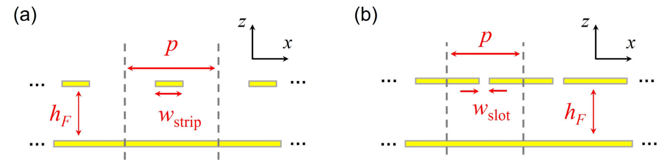

Fig. 2. (a) UC of the FPC antenna with $p = 3$ mm, $h_{\mathrm{F}} = 14.1$ mm, and $w_{\mathrm{strip}} = 0.5$ mm. (b) UC of the FPC antenna with $p = 3$ mm, $h_{\mathrm{F}} = 14.1$ mm, and $w_{\mathrm{slot}} = 0.1$ mm.

III. Dispersive Analyses

The *propagation wavenumbers* of the leaky modes correspond to *zeros* of the determinant of the linear system matrix

$$\det[\mathbf{Z}(k_{x0})] = 0. \quad (16)$$

By means of the methodology described in this article, the proposed MoM code is able to evaluate the dispersion behavior of the bound and leaky modes of periodic all-metal structures with arbitrary geometry of the corrugation in the UC using (16). To find the complex propagation wavenumbers k_{x0}, which lead to effective designs of corrugated metallic LWAs, an iterative algorithm based on the Müller's method is used [25, Sec. 9.5.2] in MATLAB. The current approach is validated on FPC antennas by comparing the results with those obtained with the HM [15]. Validation has also been performed for bound and leaky modes of 1-D periodic all-metal corrugated structures with previously published results. A proper choice of initial guesses in Müller's method provides a quick convergence. If HMs were efficient, one can easily choose the HM results as initial points, while in the all-metal corrugated structures the initial point $\beta = k_0 - 2\pi/p$ was found to be efficient in finding the expected leaky mode.

A. FPC Antenna in Air With a MSG

The dispersion behavior of a FPC antenna has been considered as a benchmark for demonstrating the capability of finding improper leaky-mode solutions of the dispersion equation.

The FPC antenna is periodically modulated as depicted in Fig. 2, which represents the unit-cell geometry. It should be noted that both the strip grating and the ground plane were discretized and thus the Green's function in the previous section can be used for the FPC antenna. Due to the metallic covering, the unperturbed guiding structure can be described as a parallel-plate waveguide (PPW), which propagates in an unimodal regime with the phase constant β_u.

In Fig. 3(a), dispersion behaviors of the fundamental TE_1 mode and the first higher-order TE_2 mode are reported for the FPC antenna with a narrow strip. While, in Fig. 3(b), the dispersion behavior of the fundamental TM_1 mode is reported for the FPC antenna with a much wider strip. The validation is performed by means of the transverse resonance technique (see, e.g., [15]). (The normalized phase constant of

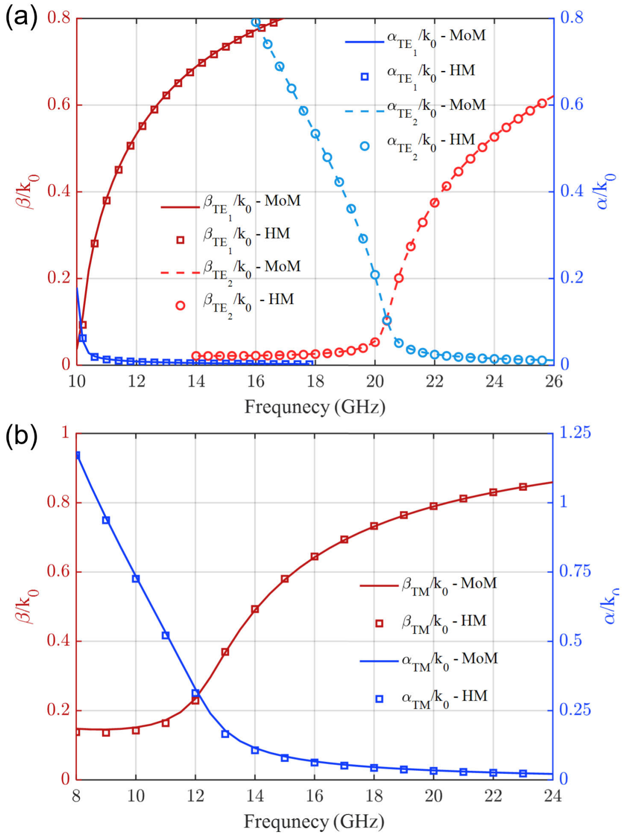

Fig. 3. Normalized phase and attenuation constants versus the frequency for (a) $\mathrm{TE}_{1,2}$ and (b) TM_1 improper leaky modes for the UC in Fig. 2.

the $n = 0$ improper harmonic is shown for the MoM case.) The agreement between the two different methods is excellent.

B. Corrugated Planes

Two infinite 1-D PEC corrugated planes, depicted in Fig. 4, are considered. The problem is assumed to be invariant along the y direction and 1-D periodic along the x direction. As discussed in [26], both the proper surface-wave regime and the proper and improper leaky-wave regimes can be supported by the corrugated surfaces.

1) TM Proper Surface Waves: If the grating period is less than half of the free-space wavelength ($p < \lambda_0/2$), the corrugated planes are able to support TM surface waves.

Firstly, the transverse resonance technique is applied here again, as a benchmark for the surface-wave mode. To this aim, the UC is assumed to be much smaller than the wavelength and the groove width to be much smaller than the period, so that the TM dispersion equation can obtained as [1]

$$Z_s + Z_0^{\mathrm{TM}} = j\zeta_0 \bar{X}_s(\omega) + \frac{k_z}{k_0}\zeta_0 = 0 \tag{17}$$

where $\bar{X}_s(\omega) = w_g/p \cdot \tan(k_0 h_g)$. If $h_g < \lambda_0/4$, propagation constants of the excited TM proper surface wave are

$$k_x = k_0\sqrt{1 + \bar{X}_s^2(\omega)}. \tag{18}$$

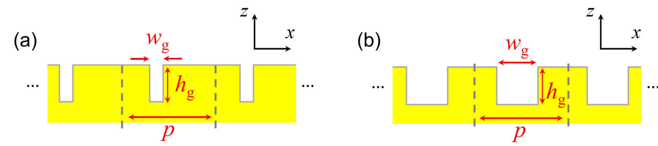

Fig. 4. (a) UC of the corrugated plane with narrow grooves. (b) UC of the corrugated plane with wide grooves.

As shown in Fig. 5(a), good agreement is obtained for the phase constants calculated with the two different methods (i.e., the MoM presented here and the HM), where the $n = 0$ proper harmonic is reported for the MoM in Fig. 5(a). A slight deviation among the results is visible at higher frequencies, when the homogenization regime gradually loses accuracy.

2) TM Proper and Improper Leaky Waves: If the period is larger than half of the wavelength ($p > \lambda_0/2$) the corrugated planes can support TM leaky modes and typically the geometry of the structure is optimized to support only one fast spatial harmonic (i.e., the $n = -1$ harmonic). A structure with a wide groove ($w_g > \lambda_0/2$) is considered in the following to validate the proposed MoM.

Since the period of the corrugation is comparable with the wavelength, and the structure radiates through the $n = -1$ fast spatial harmonic, a surface impedance cannot accurately model the structure, and the HM is not anymore applicable. Hence, the dispersive curves achieved with the proposed MoM have been compared with those presented in [14], which are calculated by means of the matrix-pencil method (MPM) applied to results obtained with a full-wave software (i.e., CST Microwave Studio). In CST, the frequency domain solver using the finite element method is applied. As known, the MPM is based on the approximation of the Hankel function (cf., [11, Eq. (2)]). However, such an approximation is not so accurate for a small wavenumber and thus results from the MPM cannot be used as a good reference around broadside. Instead, results are then further validated with a Bloch analysis (see, e.g., [15], [27]), which uses CST full-wave simulations of a finite number of UCs. Although the Bloch method might be inaccurate at higher frequencies due to higher-order Bloch modes, it provides reliable results for small wavenumbers. In the following figures, the Bloch method is performed by using 25 UCs for corrugated structures. Typically, 25 UCs are sufficient to evaluate accurately the mutual coupling effects among UCs and provide stable and promising references around broadside.

In Fig. 5(b), the dispersion behavior around the OSB region is reported; when the normalized phase constant of the $n = -1$ harmonic is negative the nature of solution is proper (backward leaky wave), whereas for positive β/k_0 values it is improper (forward leaky wave) [8], [28]. As shown, our results are in good agreement with the MPM in the off-broadside frequency region and also with the Bloch analysis in the broadside frequency region. The agreement in the entire frequency region also proves that our method can overcome the limitations of both MPM and Bloch methods.

To evaluate the MoM performance on the OSB suppression around broadside, Fig. 6 provides a convergence analysis

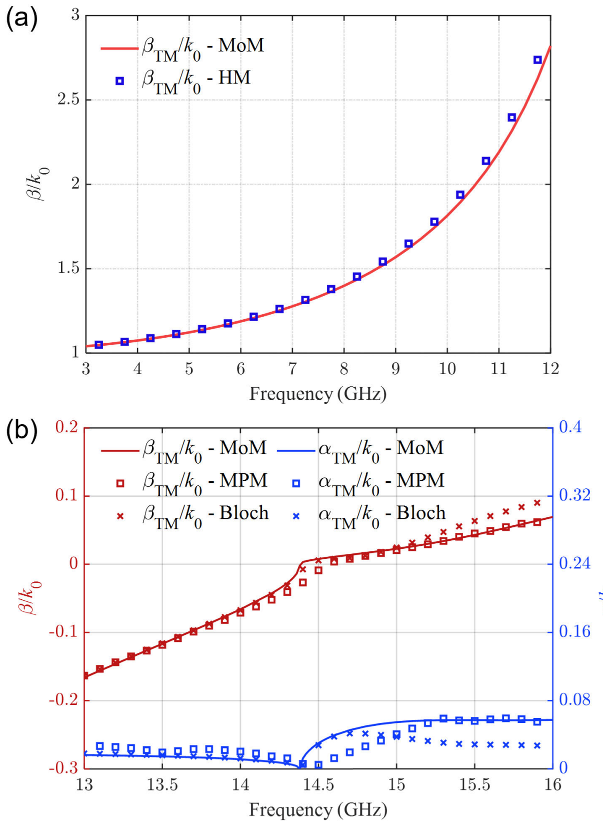

Fig. 5. Normalized phase and attenuation constant as a function of frequency for the UC of the corrugated plane in Fig. 4 supporting (a) surface wave with $p = 0.75$ mm, $w_g = 0.1p$, $h_g = 5$ mm and (b) leaky wave, with $p = 19.62$ mm, $w_g = 13.32$ mm, $h_g = 3.315$ mm. The $n = -1$ harmonic is proper when $\beta/k_0 < 0$ and improper for $\beta/k_0 > 0$.

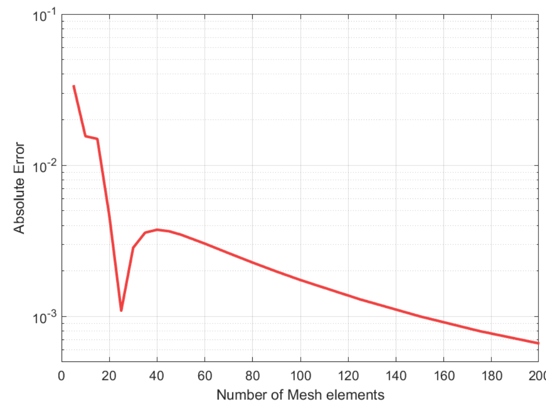

Fig. 6. Absolute error as a function of the number of mesh elements.

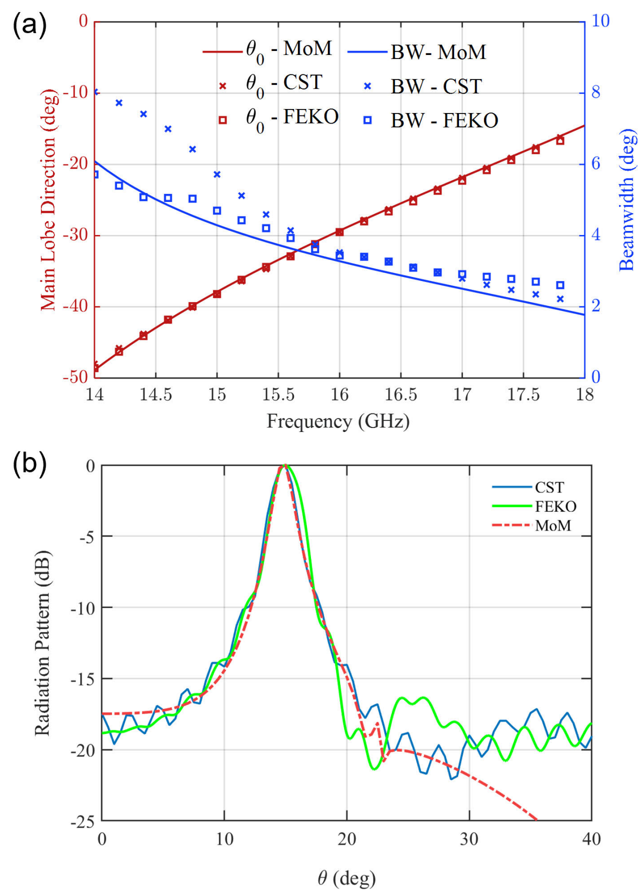

Fig. 7. (a) Comparison between the main lobe direction θ_0 and the beamwidth (BW) obtained through full-wave simulations in commercial software and the proposed MoM code. (b) Comparison between pattern obtained with commercial software and the proposed MoM via reciprocity at $f = 18$ GHz. Parameters of the UC: $p = 12$ mm, $w_g = 11.52$ mm, $h_g = 3.4$ mm.

showing the absolute error as a function of the number of mesh elements. The convergence analysis was performed for the corrugated structure in Fig. 5(b) at 14.3 GHz and the reference propagation constant is $k_{x,\mathrm{CST}}/k_0 = -0.031 - 0.0066j$ using the Bloch method. As shown, with about 100 segments, the absolute errors $|k_{x,\mathrm{CST}} - k_{x,\mathrm{MoM}}|$ are the order of 10^{-3} and relative errors $|(k_{x,\mathrm{CST}} - k_{x,\mathrm{MoM}})/k_{x,\mathrm{CST}}|$ are about 5%. Moreover, in periodic leaky-wave antennas, the dimensions of the UC are typically comparable to wavelengths and thus the low-frequency breakdown would not occur in most cases. As a result, the errors can be reduced steadily by increasing the number of mesh elements and one can use the MoM solver for the OSB suppression with 100 segments or more.

IV. Radiation Features of an 1-D Periodic All-Metal Corrugated LWA

As a further test, the radiation features of the LWA based on a 1-D periodic all-metal corrugated structure presented in [26] are analyzed here with the proposed MoM approach and validated by means of CST Microwave Studio and FEKO. In FEKO, the MoM with surface equivalence principle is adopted to analyze the antenna, where the surfaces of metal and other media are meshed by triangles. Compared to the design in Fig. 5(b), the period is smaller ($p \approx 0.6\lambda_0$ instead of $p \approx \lambda_0$) and only the $n = -1$ spatial harmonic is fast in the backward leaky regime.

The structure consists of a linear corrugated metal plate excited by a magnetic line source placed on the air-metal interface. On each side of the source, 35 UCs are present, for a total length of around 820 mm; the number of UCs is selected to radiate more than 90% of the injected power, based on the attenuation constant α provided by the MoM.

As shown in Fig. 7(a), a very good agreement has been obtained between the full-wave results for θ_0 and BW obtained with CST and FEKO for the truncated LWA and ones calculated from the complex wavenumbers given by the MoM

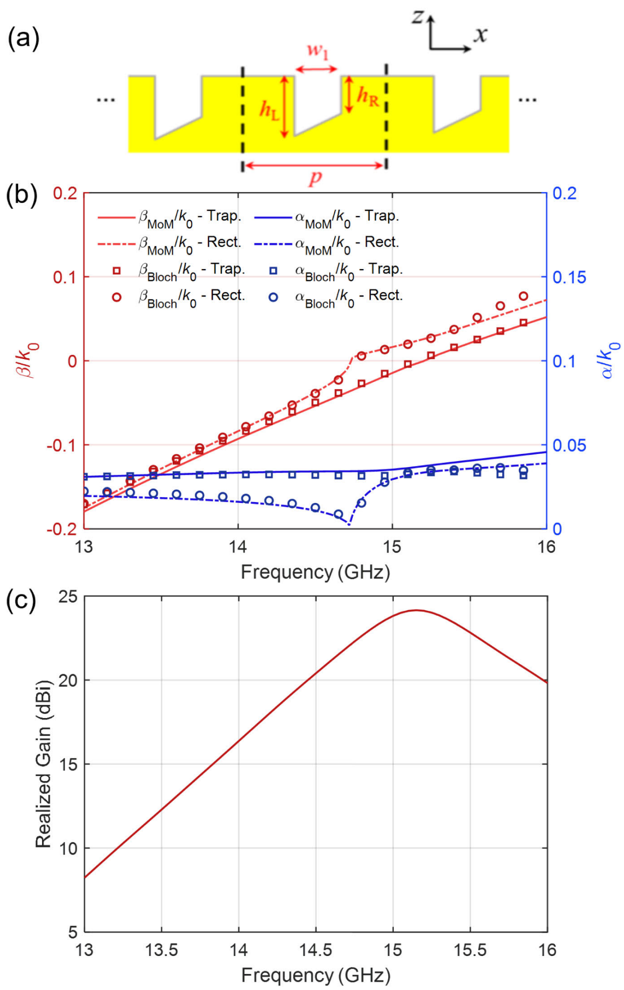

Fig. 8. (a) Asymmetric UCs for the OSB suppression with a (single) trapezoidal groove. (b) Normalized phase and attenuation constant supported by rectangular and trapezoidal UCs. (c) Simulated gain at broadside of the related 1-D corrugated LWA with trapezoidal UCs. Parameters of the rectangular UCs: $p = 19.62$ mm, $w_g = 12.32$ mm, $h_L = h_R = 3.315$ mm. Parameters of the trapezoidal UCs: $p = 19.62$ mm, $w_g = 12.32$ mm, $h_L = 4.575$ mm, $h_R = 2.055$ mm.

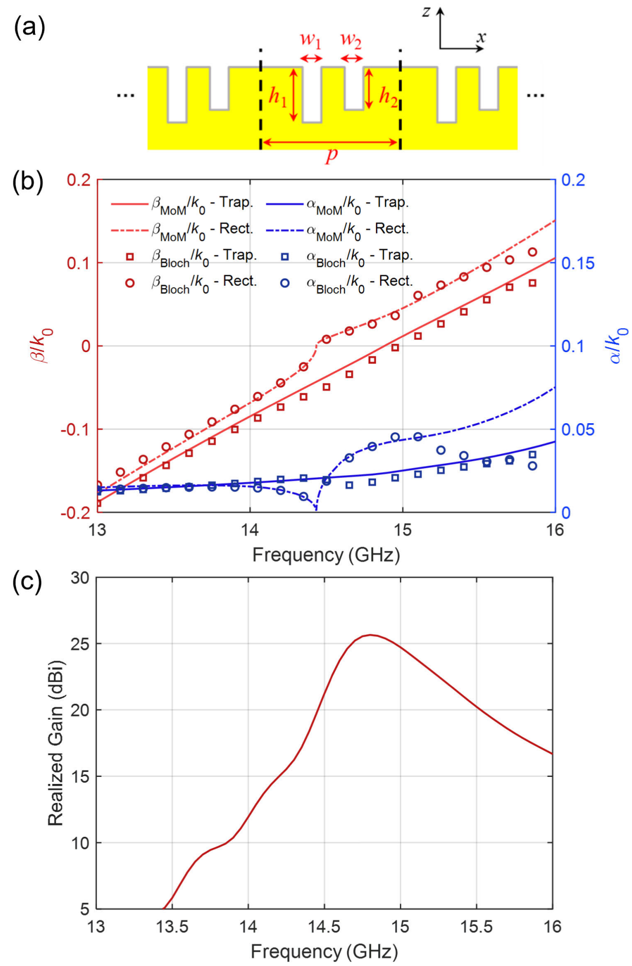

Fig. 9. (a) Asymmetric UC made by a double trapezoidal groove for the OSB suppression. (b) Normalized phase and attenuation constant supported by rectangular and trapezoidal UCs. (c) Simulated gain at broadside of the related 1-D corrugated LWA with two unequal UCs. Parameters of the symmetric UCs: $p = 19$ mm, $w_1 = w_2 = 2.62$ mm, $h_1 = h_2 = 3$ mm. Parameters of the asymmetric UCs: $p = 19$ mm, $w_1 = 2.62$ mm, $w_2 = 2.23$ mm, $h_1 = 3$ mm, $h_2 = 2.76$ mm.

simulation [9], which provides an independent validation of the proposed method. A slight mismatch is visible at lower frequencies due to the truncation enforced by the full-wave model, which unavoidably introduces diffraction effects.

The far-field pattern produced by the considered line source in an infinite LWA can be calculated using the conventional approach based on the reciprocity theorem, i.e., by letting a TM plane wave impinge on the structure and calculating the reaction of the total field calculated with the MoM and the source. As shown in Fig. 7(b), a good agreement is also obtained between the radiation pattern achieved by reciprocity (MoM) and full-wave models (CST and FEKO) for the truncated LWA, thus confirming the usefulness of the proposed full-wave modal solver. Minor differences are related to diffraction effects at the antenna truncation, not modeled in the reciprocity-based approach.

V. OSB Suppression

To illustrate the proposed MoM approach in challenging OSB suppression design, where accurate knowledge of the leaky-mode complex wavenumber is required, corrugated LWAs with optimized UCs are shown here for the first time. To this aim, a trapezoidal groove UC as shown in Fig. 8(a) is considered. It should be noted that, although the principle of introducing the relevant asymmetric shape to suppress the OSB [7] is well known, one can barely take this idea from concept to a prototype without an accurate and robust modal solver. Thanks to the proposed modal solver, as shown in Fig. 8(b), a proper selection and fine-tuning of the widths at both sizes (h_L and h_R) can be realized to produce a flat curve having nonzero attenuation constants and linear phase constants around the broadside frequency, which proves that a complete suppression of the OSB is achieved [8]. The conventional design with rectangular grooves similar to [14], whose dispersive curves have been also reported in Fig. 8(b), clearly shows an OSB. Results are also further validated with a Bloch analysis. Good agreement between the two methods can be observed.

Fig. 8(c) reports the simulated gain at the broadside of a 1-D periodic all-metal corrugated LWA using trapezoidal UCs. Again, 25 UCs on each side of the line source are used to ensure more than 90% efficiency at the broadside. Thanks to the OSB suppression, we do not observe any deterioration of the gain between 13 and 16 GHz.

A different asymmetric UC, consisting of two unequal narrow grooves, is also presented in Fig. 9. Again, compared to the conventional design with obvious OSB effects, the proposed modal solver is capable of finding a proper selection of the physical parameters which allows for completely suppressing the OSB, as proven by the consistently nonzero attenuation constants, as well as by the stable broadside gain of the relevant 1-D periodic corrugated LWA.

VI. Conclusion

An efficient numerical tool for the OSB suppression of 2-D all-metal structures, periodic along one direction, has been developed. A formulation for both the TM and TE polarizations is proposed, which employs a rapidly convergent Ewald representation for the relevant 2-D periodic Green's function with a detailed and accurate treatment of both the regular and singular parts of Green's function in the self-reaction terms.

The MoM approach has been validated and tested by means of data taken from the literature and considering independent full-wave simulations performed by means of state-of-the-art commercial software. Both the complex-valued propagation constant of the relevant leaky modes, as well as real propagation constants of surface waves, and the radiation features of a corresponding antenna, have been validated.

The approach has been proven to be accurate and versatile on different asymmetric UCs. In this way, the proposed approach is capable of designing corrugated LWAs with OSB suppression. Furthermore, its computational burden is quite limited both in terms of memory requirements, due to the small number of unknowns (typically, of the order of one hundred), and of calculation time (of the order of 2 s for the calculation of a modal wavenumber at a single frequency on an Intel[1] Xeon[1] Gold 5218 CPU). The proposed tool can be very useful for the design and optimization of innovative and compact 1-D and 2-D all-metal leaky-wave antennas, based on arbitrary shapes and configurations of the UC, offering high-gain broadside beams or scanning beams, at all frequencies.

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Dejian Zhang (Member, IEEE) received the B.S. degree from Central South University, Changsha, China, in 2017. He is currently pursuing the Ph.D. degree in control science and control engineering with Tsinghua University, Beijing, China.

He was a Visiting Ph.D. Student with the Department of Electronic Engineering, La Sapienza University of Rome, Rome, Italy, from February to July 2019. His research interests include the analysis and design of leaky-wave antennas and other terahertz and antenna technologies.

Davide Comite (Senior Member, IEEE) received the master's degree (cum laude) in telecommunications engineering and the Ph.D. degree in electromagnetics and mathematical models for engineering from Sapienza University, Rome, Italy, in 2011 and 2015, respectively.

He was a Visiting Ph.D. Student with the Institute of Electronics and Telecommunications of Rennes, University of Rennes 1, Rennes, France, from March to June 2014, and a Post-Doctoral Researcher with the Center of Advanced Communications, Villanova University, Villanova, PA, USA, in 2015. He is currently a tenure track Assistant Professor with Sapienza University. In 2022, he was a Visiting Researcher with Sorbonne University, Paris, France, and with Toulouse III University/Center d'Etudes Spatiales de la Biosphère, Toulouse, France. His research interests include the study and design of antennas, including leaky-wave antennas and metasurfaces, as well as the generation of nondiffracting waves and pulses. He is also interested in remote sensing and the study of the scattering and interaction of electromagnetic waves with natural surfaces, as well as in the Global navigation satellite system (GNSS) reflectometry and Earth observation in general. He has authored about 150 articles in international journals, books, and conference proceedings.

Dr. Comite is a Senior Member of International Union of Radio Science (URSI). He was a recipient of Awards at National and International Conferences. In February 2023, he received the National Scientific Qualification for the role of a Full Professor of Electromagnetic Fields in Italian Universities. In 2019, 2020, 2021, and 2022, the IEEE Antennas and Propagation Society recognized him as an Outstanding reviewer for the IEEE Transaction on Antennas and Propagation. In 2020, he was awarded as the best reviewer for the IEEE Journal of Selected Topics in Applied Earth Observation and Remote Sensing. He currently serves as a reviewer for several international journals. He is an Associate Editor of the IEEE Journal of Selected Topics in Applied Earth Observation and Remote Sensing, the European Association on Antennas and Propagation (EuRAAP) Journal *Reviews of Electromagnetics*, the Institution of Engineering and Technology (IET) *Journal of Engineering*, the IET *Microwaves, Antennas, and Propagation* by the Institution of Engineering and Technology, and IEEE Access. He has been leading the GNSS-R modeling Working Group of the Geoscience and Remote Sensing Society Modeling in Remote Sensing (GRSS MIRS) technical committee since January 2021. In 2020 and 2021, he was one of the officers of the Young Professional Affinity Group of the IEEE Italian Chapter. Since 2022, he is the Chair of the Early Career in Antenna and Propagation (ECAP) Working Group of EuRAAP.

Paolo Baccarelli (Member, IEEE) received the Laurea degree in electronic engineering and the Ph.D. degree in applied electromagnetics from the La Sapienza University of Rome, Rome, Italy, in 1996 and 2000, respectively.

In 1996, he joined the Department of Electronic Engineering, La Sapienza University of Rome, where he has been an Assistant Professor since November 2010. From April 1999 to October 1999, he was a Visiting Researcher with the University of Houston, Houston, TX, USA. In 2017, he joined the Department of Engineering, University of Rome Tre, Rome, where he has been an Associate Professor since July 2017. In December 2017, he received the National Scientific Qualification for the role of a Full Professor of Electromagnetic Fields in Italian Universities. He is currently teaching courses and laboratories of electromagnetic fields, advanced antenna engineering, and complex variable techniques for wave phenomena for the B.Sc., M.Sc., and Ph.D. programs in electronics and communications engineering with the University of Rome Tre. He has coauthored about 240 articles in international journals, conference proceedings, and book chapters. His research interests include analysis and design of planar antennas and arrays, leakage phenomena in uniform and periodic structures, numerical methods for integral equations and periodic structures, propagation and radiation in anisotropic media, metamaterials, graphene, and electromagnetic band-gap structures.

Dr. Baccarelli is a member of the TPCs of several international conferences. He was a Secretary of the 2009-European Microwave Week (EuMW 2009) and a member of the Local Organizing Committee of the XXXIV General Assembly and Scientific Symposium of the International Union of Radio Science (URSI GASS 2021). He was a recipient of the Giorgio Barzilai Laurea Prize from 1994 to 1995 presented by the former IEEE Central and South Italy Section. He is currently an Associate Editor of the *Electronics* (MDPI), of the *Magnetism* (MDPI), and of the *International Journal of Antennas and Propagation* (Hindawi) and acts as a reviewer for more than 20 IEEE, Institution of Engineering and Technology (IET), OSA, and AGU journals.

Xiaojiao Deng received the B.S. degree in chemistry from Changchun Normal University, Changchun, China, in 2008, the master's degree in inorganic chemistry from Northeast Normal University, Changchun, in 2011, and the Ph.D. degree in physical chemistry from the Institute of Chemistry, Chinese Academy of Sciences, Beijing, China, in 2015.

From 2016 to 2019, she was a Post-Doctoral Research Fellow with the Department of Automation, Tsinghua University, Beijing. She joined with the Department of Automation, Tsinghua University, in 2019, where she is currently an Assistant Researcher. Her current research interests include terahertz devices and terahertz spectroscopy.

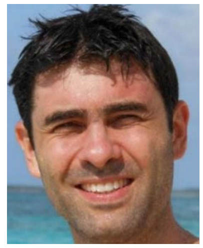
Xiaoping Zheng received the B.S. degree from the Chengdu University of TCM, Chengdu, China, in 1995, and the Ph.D. degree from Sichuan University, Chengdu, in 2003.

From 2004 to 2006, he was a Post-Doctoral Researcher with the School of Management, Fudan University, Shanghai, China. From 2006 to 2013, he was a Professor with the Institute of Safety Management, Beijing University of Chemical Technology, Beijing, China. He is currently a Professor with the Department of Automation, Tsinghua University, Beijing. His current research interests include detecting technology of dangerous chemicals and engineering management.

Alessandro Galli (Member, IEEE) received the Laurea degree in electronics engineering and the Ph.D. degree in applied electromagnetics from the Sapienza University of Rome, Rome, Italy, in 1990 and 1994, respectively.

Since 1990, he has been with the Department of Information Engineering, Electronics and Telecommunications, Sapienza University of Rome, where he became an Assistant Professor and an Associate Professor in the sector of electromagnetic fields in 2000 and 2002, respectively. In 2013, he passed the National Scientific Qualification, and in 2020, he definitively achieved the role of a Full Professor in the sector of electromagnetic fields with the Sapienza University of Rome. He is currently teaching or co-teaching the courses of electromagnetic fields, advanced antenna engineering, microwaves, and engineering electromagnetics, for the bachelor's and master's degrees in electronics engineering and in communications engineering with Sapienza University. He has authored more than 300 articles in international journals, books, and conference proceedings. He holds a patent for an invention of a type of microwave leaky-wave antenna. His current research interests include theoretical and applied electromagnetics regarding guided waves, propagation and radiation problems, mainly focused on modeling, numerical analysis, and design of antennas and passive devices from microwaves to terahertz. Specific topics involve leaky waves and leaky-wave antennas, periodic and multilayered printed structures, beam focusing devices, including complex media, metamaterials, and graphene. His research activities also concern the areas of geoelectromagnetics, bioelectromagnetics, and plasma heating.

Dr. Galli is a member of the Board of the European School of Antennas (ESoA) and of the leading scientific societies of electromagnetics. He was a recipient of various grants and prizes for his research activity, such as the Barzilai Prize for the best scientific work of under-35 researchers from the 10th National Meeting of Electromagnetics in 1994 and the Quality Presentation Recognition Award from the International Microwave Symposium by the Microwave Theory and Techniques Society of the IEEE in 1994 and in 1995. Based on a poll among about 300 attendants to the ESoA courses in a competition of more than 100 international speakers. In 2017, he has been elected as the ESoA Best Teacher. He was elected as the Italian Representative of the Board of Directors of the European Microwave Association (EuMA), the main European Society of Electromagnetics, for the 2010–2012 triennium and then re-elected for the 2013–2015 triennium. He was the General Co-Chair of the European Microwave Week, the most important conference event in the electromagnetic area at European level in 2014. Since its foundation in 2012, he has been the Coordinator of the European Courses on Microwaves, the first European Educational Institution on Microwaves. He is currently an Associate Editor of the *International Journal of Microwave and Wireless Technologies* (Cambridge University Press) and the *Institution of Engineering and Technology (IET) Microwaves, Antennas and Propagation* (Institution of Engineering and Technology).



Paolo Burghignoli (Senior Member, IEEE) was born in Rome, Italy, in February 18, 1973. He received the Laurea degree (cum laude) in electronic engineering and the Ph.D. degree in applied electromagnetics from the La Sapienza University of Rome, Rome, in 1997 and 2001, respectively.

In 1997, he joined the Electronic Engineering Department and is currently with the Department of Information Engineering, Electronics and Telecommunications, La Sapienza University of Rome. From January 2004 to July 2004, he was a Visiting Research Assistant Professor with the University of Houston, Houston, TX, USA. From November 2010 to October 2015, he was an Assistant Professor with the La Sapienza University of Rome, where he has been an Associate Professor since November 2015. In March 2017, he received the National Scientific Qualification for the role of a Full Professor of electromagnetic fields in Italian Universities. He is currently teaching courses of electromagnetic fields, advanced antenna engineering, and analytical techniques for wave phenomena for the B.Sc., M.Sc., and Ph.D. programs in the ICT area with Sapienza University. He has authored about 250 articles in international journals, books, and conference proceedings. His research interests include the analysis and design of planar antennas and arrays, leakage phenomena in uniform and periodic structures, numerical methods for integral equations and periodic structures, propagation and radiation in metamaterials, electromagnetic shielding, and graphene electromagnetics.

Dr. Burghignoli was a recipient of the "Giorgio Barzilai" Laurea Prize from 1996 to 1997 presented by the former IEEE Central and South Italy Section, the 2003 IEEE MTT-S Graduate Fellowship, and the 2005 Raj Mittra Travel Grant for junior researchers presented at the IEEE AP-S Symposium on Antennas and Propagation, Washington, DC. He is a coauthor of the "Fast Breaking Papers, October 2007" in EE and CS, about metamaterials (paper that had the highest percentage increase in citations in Essential Science Indicators). He was a Secretary of the 12th European Microwave Week (EuMW 2009) and a member of the Scientific Board and Local Organizing Committee of the 41st PhotonIcs and Electromagnetics Research Symposium (PIERS 2019). In 2016 and 2020, the IEEE Antennas and Propagation Society recognized him as an Outstanding reviewer for the IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION. He is currently an Associate Editor of the *Institution of Engineering and Technology (IET) Electronics Letters* (Institution of Engineering and Technology) and of the *International Journal of Antennas and Propagation* (Hindawi).