

Superdirective space-to-surface wave transformations with nonlocal metasurfaces

Yongming Li^{1,2,3,*}, Xikui Ma,¹ Viktor Asadchy^{2,†} and Sergei A. Tretyakov²

¹State Key Laboratory of Electrical Insulation and Power Equipment, School of Electrical Engineering, Xi'an Jiaotong University, Xi'an 710049, China

²Department of Electronics and Nanoengineering, Aalto University, P.O. Box 15500, FI-00076 Aalto, Finland

³CSG Energy Development Research Institute Co., Ltd., Guangzhou 510530, China

 (Received 29 October 2025; revised 13 February 2026; accepted 3 March 2026; published 31 March 2026)

In this work, we demonstrate theoretically and numerically that propagating waves can be fully converted into surface waves, and vice versa, using an aperiodic metasurface composed of simple, electrically small metallic elements loaded with reactive components. The proposed approach achieves nearly 100% conversion efficiency and even enables operation in the superdirective regime. We show that these metasurfaces can be designed in a modular fashion, allowing for straightforward cascading and the realization of multifunctional responses without the need for global optimization. Specifically, employing this modular methodology, we design a metasurface that converts an arbitrary incident propagating wave into a surface wave, guides it along the surface, and reradiates the received power in any desired direction with a tailored wave-front shape. The proposed technique can be used in antenna engineering and in advanced designs of reconfigurable intelligent surfaces.

DOI: [10.1103/PhysRevApplied.25.034095](https://doi.org/10.1103/PhysRevApplied.25.034095)

I. INTRODUCTION

In the domain of electromagnetic wave manipulation, metasurfaces play an invaluable role, including versatile control of propagating [1–5] and evanescent waves [1,6–9]. One key functionality is the conversion between propagating and surface (guided) waves [8,10]. Due to the momentum mismatch between propagating waves and surface waves, the conventional realization methods, such as prism [11] and grating [12] coupling methods, suffer from inherently low conversion efficiency [13]. At microwave frequencies, this conversion is usually realized by leaky-wave antennas (LWAs) [8,14–20]. Typically, the radiated field at the aperture of an LWA does not exhibit a homogeneous in space amplitude nor a linearly varying phase profile. Hence, the conversion from the surface to propagating waves is not perfect in this scenario. Only recently, aperiodic nonlocal LWAs requiring global optimization were proposed that rigorously account for the mutual coupling between the radiating elements,

providing the opportunity for near-perfect surface-to-propagating wave-conversion efficiencies [21,22], though not the reverse.

Space-to-surface wave conversion has been shown with *periodic* metasurfaces whose phase gradients partially offset the momentum mismatch, but efficiencies were relatively low [13,23,24]. A near-ideal converter using an anisotropic impedance with quasilinearly growing surface-wave power was proposed, yet the required profile was difficult to realize and tune [8]. An omega-bianisotropic alternative removed the need for anisotropy but still required a simultaneously excited input surface wave to reach high efficiency [9]. Moreover, periodic designs are sensitive to the period and incidence angle, so any change demands redesign. A complementary approach achieved nearly 100% efficiency but was confined to electrically large metasurfaces with slowly varying modulation and relied on a locally periodic approximation [25]. Furthermore, space-to-surface wave conversion for both incident wave polarizations was demonstrated in Ref. [26]. More recently, *aperiodic* converters designed via global optimization were proposed; however, these methods assumed ideal sheet impedances that often yielded sharply varying, hard-to-fabricate surface-impedance profiles [27–30]. Their implementation requires proper consideration of mutual coupling between the physically patterned unit cells [31,32].

Once the surface wave is generated, it is crucial to guide it and at some point relaunch it back into free space

*Contact author: yongmingli@stu.xjtu.edu.cn

†Contact author: viktar.asadchy@aalto.fi

Published by the American Physical Society under the terms of the [Creative Commons Attribution 4.0 International](https://creativecommons.org/licenses/by/4.0/) license. Further distribution of this work must maintain attribution to the author(s) and the published article's title, journal citation, and DOI.

[1,33,34]. In Ref. [1], the authors proposed the hybrid metasurface concept to control the surface wave and propagating wave simultaneously, where a planar metamaterial is used to interact with surface waves and a holographic metasurface is used to modulate propagating waves. In Refs. [33,34], the authors demonstrated the conversion of plane into surface waves, their routing, and converting them back into an outgoing plane wave. However, their design suffered from low efficiency (about 10%) due to the surface-wave dissipation loss, scattering at the metasurface discontinuities, and limited conversion efficiency.

In this work, we propose aperiodic finite-sized metasurfaces that can convert between space and surface waves, guide the surface wave, and relaunch it as a plane wave in a desired direction or focus it at a specific point, all with nearly 100% efficiency. The proposed metasurfaces are modular, meaning they can be designed individually and cascaded to provide complex desired functionality without the need for global optimization. Moreover, we show that when the dissipation in the metasurface is sufficiently small, conversion efficiency can exceed the 100% limit, reaching the so-called superdirective regime, known in antenna theory for electrically small antennas and densely sparsely arrays [35–37]. The metasurface is modeled as a physically realizable array of impedance-loaded conducting strips with full mutual coupling, and the required loads can be implemented using an established synthesis procedure [38]. The values of the load impedances are determined using the analytical theory empowered by the multipopulation genetic algorithm (MPGA) optimization [39,40]. Furthermore, if empowered by tunable reactive loads such as varactor diodes, the proposed metasurfaces can maintain very high conversion efficiency, even when the angle of incidence (or the angle of radiation in the leaky-mode regime) varies *continuously*.

II. PRINCIPLE AND METHODOLOGY

We model the aperiodic metasurface as a *finite* array of thin conducting strips oriented along the x direction and located at a distance h above an *infinite* perfect electric conductor (PEC) ground plane. The configuration of the proposed structure is shown in Fig. 1. The ground plane is in the $z = 0$ plane. This setup is chosen to simplify the analysis. In practice, such a configuration can be realized by placing the strips on top of a radio-transparent Styrofoam substrate with relative permittivity $\epsilon_r \approx 1$ and thickness h . Nevertheless, the analysis can be extended to other substrates using the theoretical framework presented in Ref. [41]. The strips of width w are uniformly spaced by a distance d along the y axis. To be more specific, the very first strip is numbered 0 and placed at position $y = 0$, while the n th strip is placed at position $y_n = nd$. For the analysis, thin conducting strips can be modeled as equivalent round wires with the effective radius

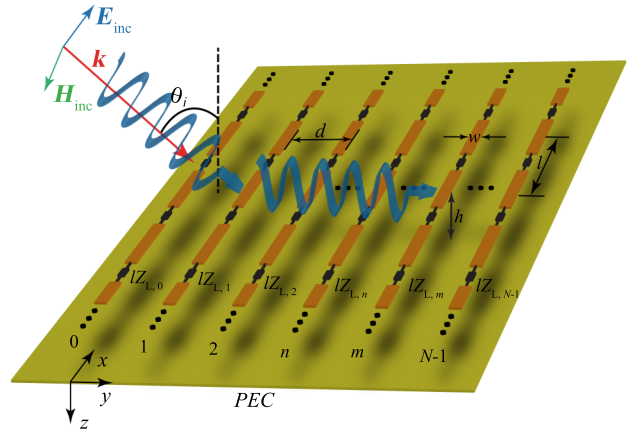


FIG. 1. A finite-sized array is composed of N metallic loaded strips over an infinite ground plane under a TE-polarized plane-wave illumination at an angle θ_i . The metasurface is designed to convert the incident plane wave into a surface wave. The strips are loaded by bulk complex-valued impedances inserted periodically with the period l along the x direction.

$r_{\text{eff}} = w/4$ [42]. The strips are periodically loaded, with a period l , by load impedances $lZ_{L,n}$, where $Z_{L,n}$ represent the load impedances per unit length of the strips (see Fig. 1). The load impedance per unit length is uniform along the x direction. It is assumed that l and w are much smaller than the wavelength in free space. Hence, the periodically loaded strips can be modeled as homogeneous thin impedance strips with the impedances per unit length $Z_{L,n}$, where $n \in \{0, 1, 2, \dots, N-1\}$. While the geometric arrangement of the strips is periodic along the y axis, the variation in load impedances along this axis renders the metasurface aperiodic in the y direction.

The metasurface is illuminated by a plane wave at an incidence angle θ_i . The incident wave is assumed to be a transverse-electric (TE)-polarized wave with the electric field given by

$$\mathbf{E}^{\text{inc}}(y, z) = E_0 e^{-jk_0(\sin \theta_i y + \cos \theta_i z)} \hat{x}, \quad (1)$$

where E_0 is the amplitude of the incident wave and k_0 represents its wave number in free space. Throughout the paper, the time dependence of the complex-valued fields is assumed to be in the form of $e^{j\omega t}$. The specular reflection of the incident wave from the ground plane has the electric field $\mathbf{E}^{\text{ref}} = -E_0 e^{-jk_0(\sin \theta_i y - \cos \theta_i z)} \hat{x}$. The superposition of the incident wave and the specular reflection of the incident wave from the ground plane is denoted as the external field, where the value at coordinate (y, z) is $E_x^{\text{ext}}(y, z) = -j2E_0 \sin(k_0 \cos \theta_i z) e^{-jk_0 \sin \theta_i y}$. In what follows, without loss of generality, the frequency and amplitude of the incident wave are chosen as $f = 10$ GHz, and $E_0 = 1$ V/m, respectively. The distance h above the ground plane is chosen $\lambda/6$ and the width of the strips is $w = \lambda/100$.

To calculate the scattered electric field from the considered metasurface, we use the approach proposed in Ref. [43] based on Ohm's law and the load impedance matrix. This approach assumes that the effective radius of the strips is negligible compared to both the wavelength ($k_0 r_{\text{eff}} \ll 1$) and the distance between the strips ($r_{\text{eff}} \ll d$), so that we can replace the field generated by each strip with that of a line current [42]. For each loaded strip, the relationship between induced line current and induced voltage on the loaded strip has to satisfy Ohm's law. The relationship between the induced line current and the external voltage can be bridged through the load impedance matrix $\bar{\bar{Z}}$ as

$$\bar{\bar{Z}} \cdot \vec{I} = \vec{U}, \quad (2)$$

where the vector of induced currents is $\vec{I} = [I_0, I_1, \dots, I_{N-1}]^T$, and the external voltage vector is $\vec{U} = [E_x^{\text{ext}}(y_0, -h), E_x^{\text{ext}}(y_1, -h), \dots, E_x^{\text{ext}}(y_{N-1}, -h)]^T$. $\bar{\bar{Z}} = \bar{\bar{Z}}_s + \bar{\bar{Z}}_L + \bar{\bar{Z}}_m$ is composed of the diagonal self-impedance matrix $\bar{\bar{Z}}_s = \text{diag}(Z_0, Z_1, \dots, Z_n, \dots, Z_{N-1})$, the diagonal load impedance matrix $\bar{\bar{Z}}_L = \text{diag}(Z_{L,0}, Z_{L,1}, \dots, Z_{L,N-1})$, and the mutual impedance matrix $\bar{\bar{Z}}_m$ with the matrix elements given by $Z_{nm} = (k_0 \eta_0 / 4) [H_0^{(2)}(k_0 d_{nm}) - H_0^{(2)}(k_0 \sqrt{d_{nm}^2 + 4h^2})]$. Here, $H_0^{(2)}$ is the Hankel function, and η_0 is the free-space impedance.

For a given set of load impedances $\bar{\bar{Z}}_L = \text{diag}(Z_{L,0}, Z_{L,1}, \dots, Z_{L,N-1})$, the corresponding induced current distribution can be obtained by solving Eq. (2). Once the induced current distribution is known, the electric field generated by the strip array in the upper half-space can be calculated analytically as

$$\mathbf{E}^{\text{strips}} = -\frac{k_0 \eta_0}{4} \sum_{m=0}^{N-1} I_m \left[H_0^{(2)} \left(k_0 \sqrt{(y - y_m)^2 + (z + h)^2} \right) - H_0^{(2)} \left(k_0 \sqrt{(y - y_m)^2 + (z - h)^2} \right) \right] \hat{x}. \quad (3)$$

The scattered electric field is calculated by subtracting the incident electric field from the total electric field. For our finite-sized array, the scattered electric field $\mathbf{E}^{\text{sca}}$ is the sum of the field generated by the strips $\mathbf{E}^{\text{strips}}$ and the specularly reflected field $\mathbf{E}^{\text{ref}}$:

$$\mathbf{E}^{\text{sca}} = \mathbf{E}^{\text{strips}} + \mathbf{E}^{\text{ref}}. \quad (4)$$

In this work, the multipopulation genetic algorithm (MPGA) is used to optimize the load impedances. Compared with our earlier work [43], where the optimization was performed using function *fmincon* in MATLAB, MPGA provides a more robust and flexible approach for handling complex, multimodal objective landscapes, which

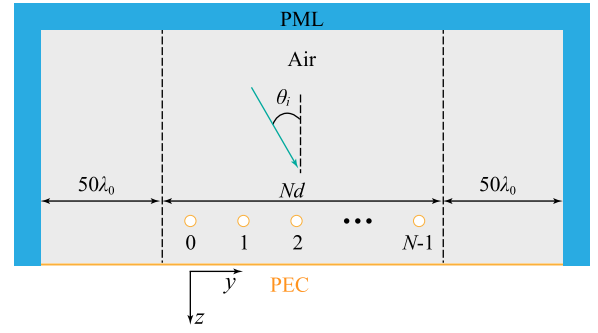


FIG. 2. COMSOL Multiphysics simulation configuration. The blue part is the PML (perfect matched layer), while the shaded area is the air background. The orange line at the bottom is PEC. The orange circles are used to simulate surface current density while connected with bulk loads under the illumination of incident wave with the incident angle θ_i .

are common in load impedance optimization problems. Unlike gradient-based methods, MPGA does not rely on initial guesses and is less likely to get trapped in local minima, making it well suited for nonconvex problems [39,40]. Moreover, MPGA inherently supports parallelism and population diversity, which enhances its global search capabilities and convergence reliability.

To keep our design feasible, we introduce the following restriction on the imaginary part of the optimized impedances in all the strips: $-9 \times 10^5 \Omega/\text{m} < \text{Im}\{Z_{L,n}\} < -500 \Omega/\text{m}$. This restriction ensures that all the optimized strips will have a capacitive response with realistic capacitance values. Once the optimization is completed, we calculate the distribution of the scattered electric field and Poynting vector above the metasurface by simulating it in COMSOL Multiphysics. The simulation domain is shown in Fig. 2. The line currents with indices from 0 to $N - 1$ are surrounded by perfectly matched layers on the three sides and a perfect electric conductor (PEC) boundary at the bottom.

III. CONVERSION FROM A PLANE WAVE TO SURFACE WAVES

To obtain a high-efficiency conversion of the incident plane wave into a surface wave, in our optimization, we assume that all the strips of the metasurface except the last one are loaded with purely reactive loads, i.e., $Z_{L,n} = jX_{L,n}$ for $n \in [0; N - 2]$, where $X_{L,n}$ denote reactance per unit length of the n th strip. The last strip is described by a complex impedance $Z_{L,N-1} = R_{L,N-1} + jX_{L,N-1}$, where $R_{L,N-1}$ is a positive-valued resistance per unit length, modeling the input resistance of the receiver. We optimize the values of the resistance $R_{L,N-1}$ and reactances $X_{L,n}$ with the objective to maximize the power delivered to the last ($N - 1$) strip. This objective will ensure that the power incident on all lossless strips will be directed along the metasurface

toward its right edge and eventually terminated in the lossy load $Z_{L,N-1}$. Such channeling of the power along the metasurface automatically results in the efficient conversion of the incident plane wave into the surface (guided) waves before the whole incident power is absorbed in the last strip as in a perfectly matched load or transmission line. Of note, in the following sections, we show that once optimized, this propagating-to-surface wave converting metasurface can be used as a module together with other metasurface modules, in particular, those that are designed to simply guide surface waves or reradiate them into free space. In this case, we can use the same optimized load impedance values and only eliminate the resistance $R_{L,N-1}$ of the last strip.

Based on the aforementioned observations, we define the conversion efficiency of propagating into surface waves as the ratio of the power absorbed in each unit length of the last strip (collected by the perfectly matched load) and the incident power $P_{\text{inc}} = (E_0^2/2\eta)N d \cos \theta_i$ illuminating the geometric area of one unit length of the metasurface:

$$\xi = \frac{P_{N-1}}{P_{\text{inc}}}. \quad (5)$$

Here, $P_{N-1} = \frac{1}{2} |I_{N-1}|^2 \text{Re}\{Z_{L,N-1}\}$ represents the dissipated power by the last strip connected with lossy loads. Thus, to optimize the metasurface for the maximum conversion efficiency, we define the objective function as

$$\mathcal{O} = \min \{-P_{N-1}\} = \min \left\{ -\frac{1}{2} |I_{N-1}|^2 \text{Re}\{Z_{L,N-1}\} \right\}. \quad (6)$$

In this section, we fix the width of the metasurface to $L = Nd = 6.5\lambda_0$ without loss of generality. First, we analyze the scenario where the incident plane wave is impinging on the metasurface at the normal direction, i.e., $\theta_i = 0^\circ$. The distance between the adjacent strips in the metasurface d strongly influences the achievable conversion efficiency. The optimized conversion efficiencies ξ for different values of spacing d are shown in Table I. When the distance between the adjacent strips is $d = \lambda/2$ (scenario of traditional reflectarrays), it is impossible to generate and optimize necessary evanescent waves (same as surface waves) [5,43]. In this case, the maximum conversion efficiency is 13.7% (see Table I). To effectively excite the optimal spectrum of evanescent waves, the density of strips should be increased. It can be noted that the conversion efficiency improves monotonically with the increase of the strip density, especially when the spacing between adjacent strips is reduced from $\lambda/2$ to $\lambda/4$, where evanescent waves are effectively controlled. When the distance is $d = \lambda/8$ or less, the strip array exhibits the so-called superdirective conversion with efficiency above 100%. It

TABLE I. The space-to-surface-wave conversion efficiency of optimized metasurfaces with different distances between the strips.

d/λ	1/2	1/4	1/6	1/8	1/10	1/12
ξ (%)	13.7	79.6	99.2	101.5	113.4	114.6

is useful to emphasize that the reported conversion efficiencies exceeding 100% do not imply power generation and are fully consistent with passivity. This outcome is a consequence of our normalization: P_{inc} is defined using the incident power illuminating the *geometric* footprint of the metasurface. In the superdirective regime, the optimized subwavelength current distribution enables the structure to extract energy from a lateral region larger than its physical extent, i.e., it exhibits an effective collection area (effective aperture or absorption cross section) exceeding the geometrical one, which is a well-known property of resonant scatterers and superdirective antennas [44–46]. As a result, the power delivered to the terminal load can exceed the incident power associated with the geometric area, even though the total extracted power is still bounded by the total power carried by the incident wave. Throughout the paper, we therefore interpret $\xi > 1$ as an indicator of this superdirective power-collection regime rather than an absolute energy efficiency. A similar superdirective regime was recently demonstrated for finite-size electromagnetic absorbers of plane waves [47–49].

According to Table I, further increasing the strip density yields only diminishing returns and the conversion efficiency gradually saturates. In theory, there is no upper limit on the absorption cross section of linear passive bodies, e.g., Refs. [45,48]. However, in practice, once d becomes sufficiently small, the metasurface already has enough subwavelength degrees of freedom to approximate the near-optimal current distribution and evanescent spectrum needed for coupling into the surface-wave channel. Beyond this point, the achievable improvement is constrained mainly by the large electrical size of the aperture, for which the effective collection area cannot substantially exceed the geometric area [45,46]. Furthermore, the superdirectivity effects are limited by inevitable dissipation losses that dampen excitation of required evanescent modes [45]. In what follows, we select $d = \lambda/8$ as the optimal value, balancing conversion efficiency with a reduced number of strips to facilitate simpler and more cost-effective fabrication. For this strip spacing, the conversion efficiency calculated by Eq. (5) reaches 101.5%. The optimized load impedances of such a metasurface are depicted in Fig. 3(a). As one can see, all the impedances have capacitive responses. Their capacitances do not reach extreme values, which makes them feasible for implementation. Although the variation of the impedance values between neighboring current wires is abrupt, it does not

introduce any challenge for the implementation since our theory already takes into account the couplings between all the current strips. The scattered electric field by the optimized metasurface is plotted in Fig. 3(b). This field includes the scattering by both the strips and the ground plane. The scattered field is confined near the surface of the strip array, which means that the incident plane wave is indeed effectively converted to a surface wave prior to being absorbed in the terminating strip of the array. The incident power illuminating the metasurface is efficiently channeled along the metasurface as a guided surface mode, as seen from the Poynting vector distribution of the scattered waves shown by red arrows in Fig. 3(b). Due to the accumulation of power, the converted surface wave is not uniform across the metasurface. The intensity of the electric field becomes stronger as the wave propagates along the surface and diminishes gradually near the lossy last strip of the metasurface. Almost all the energy of the converted surface wave is eventually dissipated by the lossy loads connected to the last strip, which behaves like a perfectly matched load of a transmission line.

To illustrate the spectrum of wave numbers of the generated surface waves (evanescent modes) at the metasurface, we calculate the spatial Fourier transform of the total induced currents along the surface of the strip array $I(y) = \sum_{n=0}^{N-1} I_n \delta(y - y_n)$ given by

$$\hat{I}(k_t) = \int_{-\infty}^{+\infty} I(y) e^{-jk_t y} dy = \sum_{n=0}^{N-1} I_n e^{-jk_t y_n}, \quad (7)$$

where k_t represents the tangential component of the wave vector. The amplitude of the induced current $\hat{I}$ as a function of k_t is shown in Fig. 4. For normal incidence, it is represented by a black solid curve in the figure. The shaded region depicts the evanescent modes (modes outside the light cone). There is a single peak in the free-space-propagation region precisely at $k_t = 0$. This peak corresponds to induced space-uniform currents that generate a propagating reflected wave that destructively interferes with the wave specularly reflected by the ground plane, nearly completely eliminating it. The peak outside the light cone, close to its right edge, corresponds to the dominant surface mode with the normalized wave number $k_t/k_0 = 1.135$. It is worthwhile to note that there are other smaller-amplitude peaks in the region of evanescent modes, implying that multiple surface waves are generated along the metasurface. Nevertheless, the effect of the dominant surface mode with $k_t/k_0 = 1.135$ can be seen from Fig. 3(a) where some *effective* (quasi) periodicity in the impedance values can be observed. It is easy to show it using the momentum matching condition $k_t = k_0 \sin \theta_i + q(2\pi/d_{\text{eff}}) d_{\text{eff}}$, where q is an integer number (diffraction order) that is typically equal to 1 for the

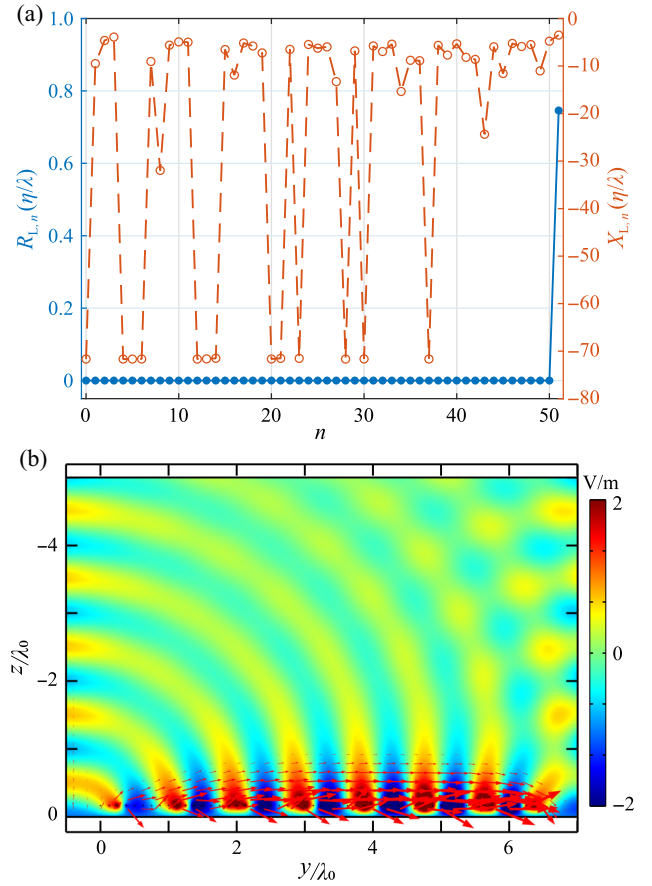


FIG. 3. (a) Real and imaginary parts of the optimized load impedance as a function of strip number n for the metasurface converting a normally incident wave ($\theta_i = 0^\circ$ and $|E^{\text{inc}}| = 1 \text{ V/m}$) to surface waves. (b) Real part of the scattered electric field $\text{Re}\{E^{\text{sca}}\} [\text{V/m}]$ above the metasurface. The red arrows indicate the Poynting vector distribution of the scattered wave.

dominant scattering mode and d_{eff} is the effective (electromagnetic) period of the metasurface. Substituting in this equation $k_t/k_0 = 1.135$, one can obtain $d_{\text{eff}} \approx 0.88\lambda_0$, that gives $d_{\text{eff}} \approx 7d$, where d is the distance between neighboring strips (we assumed $d = \lambda_0/8$). Indeed, as is seen from Fig. 3(a), the impedance pattern has an effective periodicity of every $n = 7$ strips in the left half of the metasurface.

We have also optimized the metasurfaces for converting obliquely incident waves ($\theta_i = 30^\circ$ and $\theta_i = -30^\circ$) into surface waves. For the former and latter scenarios, after the optimization, we obtained the conversion efficiencies according to Eq. (5) to be 115.4% and 110.5%, respectively. These values are even higher than for the normal illumination scenario.

IV. CONTROLLING THE WAVE-NUMBER SPECTRUM OF THE SURFACE MODES

As is seen in Fig. 4, the generated surface wave is not a single mode but a group of surface modes, in which the

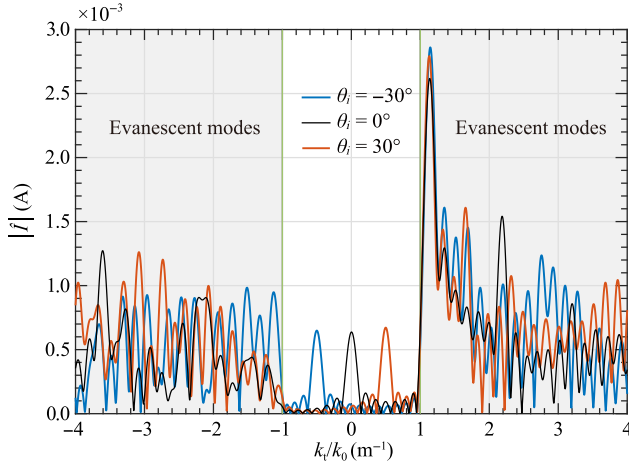


FIG. 4. The complex amplitudes of the spatial harmonics of the electric current on the metasurface ($z = -h$) for three different incident angle scenarios. The shaded area represents the evanescent modes while the propagating modes exist in the white area (inside the light cone). The vertical green solid lines are the light lines. The positive values of k_t mean that the generated wave by the strip array propagates towards the $+y$ axis, while the negative one means propagating towards the $-y$ axis.

forward and backward traveling surface waves co-exist. Indeed, although most of the power is carried by surface modes with wave number $k_t/k_0 = 1.135$, there is also strong (especially for $\theta_{\text{inc}} = -30^\circ$) power carried in the opposite direction with $k_t/k_0 = -1.135$. This superposition of the forward and backward surface modes leads to a partial standing-wave pattern at the metasurface. It may be expected that the best conversion scenario would be creation of a single surface mode that would carry the power to the matched load. However, interestingly, the optimization shows that it is optimal to allow some reflections from the load. In this optimal regime, the array behaves as a resonant object, where the reactive fields interact with the incident field at longer times, leading to higher, even superdirective conversion efficiencies. It is pivotal that in this case the resonant object is the whole array, potentially offering better bandwidth compared to more conventional superdirective arrays formed by many small resonant elements.

However, the use of resonant metasurface panels may not be desirable in practical situations. Next, we show that this resonance is effectively suppressed by introducing small real parts to the load impedances inside the first $N - 1$ strips. These nonzero real parts could be due to dissipative loss in the loads and/or due to the finite conductivity of metallic strips. In the latter scenario, the real part of the load impedance can be estimated by $\text{Re}\{Z_L\} = \rho/S_{\text{eff}}$, where ρ represents the bulk resistivity of the strips, while the effective area $S_{\text{eff}} = 2\pi r_{\text{eff}}\delta$ represents the effective skin surface area of the strip with the skip depth given by $\delta = \sqrt{2\rho/\omega\mu_0}$. For example, for copper strips with

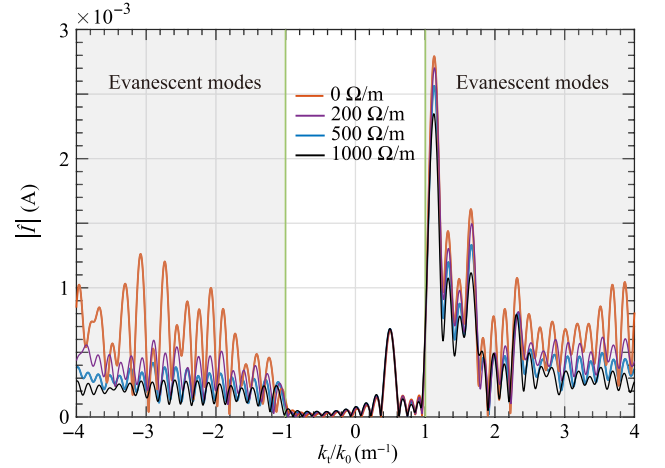


FIG. 5. The complex amplitudes of the spatial harmonics of the electric current on the metasurface ($z = -h$) for the introduced different lossy load impedances when the incident angle θ_i equals 30° . All the load impedances except the last one have the same real parts shown in the plot legend.

$r_{\text{eff}} = w/4 = \lambda/400$, at the frequency of 10 GHz, the parasitic real parts of the impedance per unit length reaches the value of $\text{Re}\{Z_L\} = 55 \Omega/\text{m}$.

In Fig. 5, we plot the spatial harmonics of the electric current for the optimized metasurface with $\theta_i = 30^\circ$ when the resistance per unit length in the first $N - 1$ strips is varied between $0 \Omega/\text{m}$ and $1000 \Omega/\text{m}$. We can see that the backward propagating surface modes with $k_t/k_0 = -1.135$ are strongly suppressed even with very small resistance values. Expectedly, the introduction of the resistance leads to a small drop in the conversion efficiency of plane waves to surface waves (around 2% for the scenario of $200 \Omega/\text{m}$).

Furthermore, it is possible to maximize the power carried by a specific mode with some desired tangential component of the wave vector k_{des} . This can be achieved using the objective function defined as

$$\mathcal{O} = -F(k_{\text{des}})P_{N-1}, \quad (8)$$

where $F(k) = \text{ge}(|\hat{I}(k)|, |\hat{I}(k - \delta)|)\text{ge}(|\hat{I}(k)|, |\hat{I}(k + \delta)|)$ is used to control the value of k_t , δ is a small number chosen to be 1.64 in this study. Logical function $\text{ge}(x_1, x_2)$ returns 1 if independent variables x_1 and x_2 satisfy $x_1 \geq x_2$, otherwise it returns 0. Function $F(k_{\text{des}})$ returns 1 if the local maximum occurs at the position k_{des} , otherwise it returns 0. To show the effectiveness of the method, several examples with normal incidence are shown below. To illustrate the effectiveness of this manipulation, the k_{des} is selected as $1.05k_0$, $1.10k_0$, and $1.15k_0$. The calculated magnitudes of the induced current spatial harmonics are depicted in Fig. 6. The corresponding conversion efficiencies in these cases are 74.3%, 94.1%, and 97.8%, respectively. As one can see, after the optimization, the desired current spatial harmonic was strongly enhanced. At the same time, the

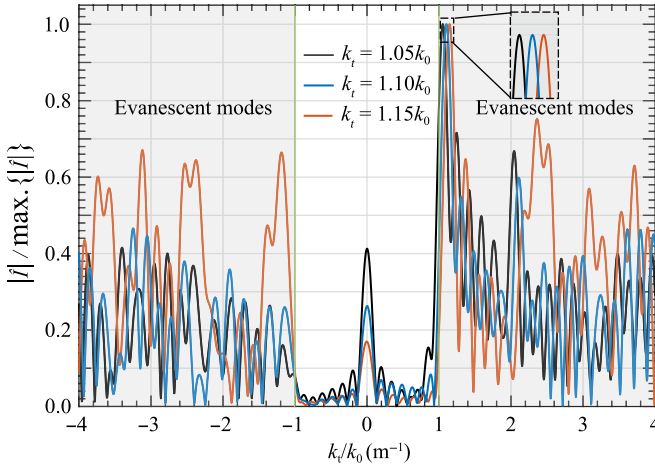


FIG. 6. The complex amplitudes of the spatial harmonics of the electric current for metasurfaces designed to generate waves with different desired tangential wave vectors k_{des} . The illumination is at normal incidence.

standing-wave pattern became more prominent, especially for $k_{\text{des}} = 1.05k_0$.

V. METASURFACES OPERATING AS TRANSMITTING LEAKY-WAVE ANTENNAS

When fed from the last strip, the designed metasurfaces are capable of launching a focused beam into free space due to the reciprocity principle [50]. Moreover, by reconfiguring the other strips, one can tune the propagation direction of this beam. In practice, reconfigurable components can be realized using components such as varactor diodes, which introduce dispersion and tunability into the load impedance [51,52].

Mechanical scanning methods of leaky-wave antennas require complex driven systems, suffering from high cost and slow scanning speed [14]. Electronic scanning methods [53–55] have limited beam-scanning range and aperture efficiency. Here, two different electrically scanning methods are illustrated for the designed metasurfaces. One is realized by tuning the reconfigurable load impedances, keeping the working frequency fixed. Another method is realized by tuning the working frequency of the feeding network with fixed load impedances.

A. Scanning method I: Tuning the working frequency of the feeding network with fixed load impedances

Here the real parts of loads (resistors) are assumed to be independent of the working frequency, while the imaginary parts (capacitors are assumed in this work) are dependent on the working frequency. For the load impedance $Z_{L,n}$, it follows $Z_{L,n} = R_{L,n} + 1/(j\omega C_{L,n})$. To create a leaky-wave antenna, we modify the optimized metasurface shown in Fig. 3(a) by equating the resistance of the last strip to zero,

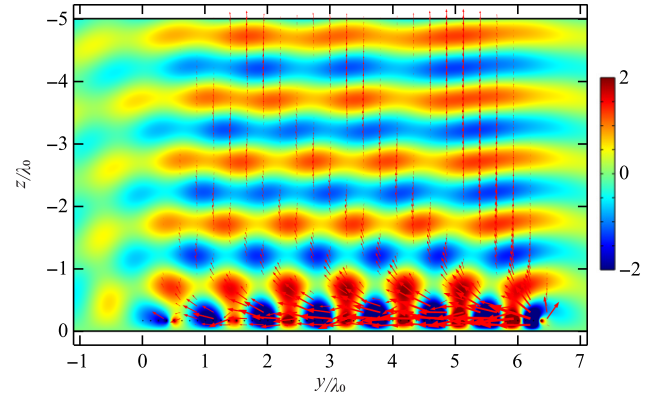


FIG. 7. The real part of the scattered electric field $\text{Re}\{E_x^{\text{sca}}\}$ [V/m] distributions over space. The red arrows represent the distribution of the Poynting vector of the scattered wave. The operating frequency is 10 GHz.

i.e., $R_{L,N-1} = 0$, assuming an ideal voltage source feeding. Now, instead of illuminating the metasurface from the normal direction, we feed it from the last $(N - 1)$ -s strip. Due to reciprocity, the metasurface generates broadside radiation in the form of a beam leaking toward the normal direction. Although this beam has a quasiplanar wave front, it is not a plane wave due to its finite waist (the metasurface is finite). The radiated electric-field distribution above the metasurface is depicted in Fig. 7 at the working frequency of 10 GHz. The wave front of the radiated wave indeed is nearly planar, exhibiting a space-uniform radiated beam by the entire metasurface area. The radiation pattern of the metasurface is shown in Fig. 8 with the black solid curve. Similarly to the conventional LWAs, the proposed structure also possesses the main beam-scanning capacity when the working frequency of the feeding network changes. As the working frequency increases from 9.5 to 10 GHz, the steered beam scans from the backward radiation across the broadside to the forward radiation (see Fig. 8 with solid curves). This broadside radiation is realized due to the perfect impedance match, and the open stop band effects are prohibited.

The metasurface can also be designed to operate as a LWA in the forward-radiation and backward-radiation regimes [56,57]. The former regime is also referred to as an anomalous (or fast-wave) radiation regime. In this scenario, the main radiation beam is directed away from the source (the last metasurface strip). The intensity of the scattered electric field and its Poynting vector show an exponential increase along the broadside direction. First, we optimize the metasurface to convert an incident plane wave from $\theta_i = 30^\circ$ into surface waves absorbed by the last strip. Next, we remove the resistance of the last strip and feed it with a wave port. The radiated electric field and its Poynting vector distribution are shown in Fig. 9 for a working frequency of 10 GHz. One can observe the beam

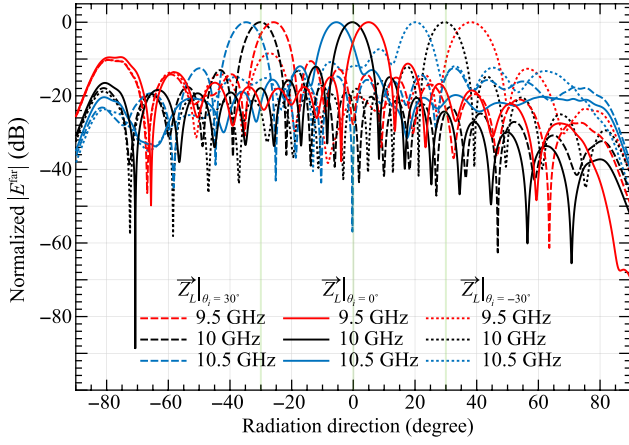


FIG. 8. Radiation patterns for different working frequencies for three different metasurfaces. Initially, the metasurfaces were designed to convert incident plane waves at angles $\theta_i = -30^\circ$, 0° , and $+30^\circ$ into a surface wave absorbed by the last strip. By reciprocity, in the transmitting regime, the metasurfaces were fed from the last strip, generating leaky waves with planar wave fronts propagating towards angles $\theta_{\text{rad}} = -30^\circ$, 0° , and $+30^\circ$, respectively. Hereafter, the radiation patterns are normalized by the maximum value of each pattern.

radiation toward angle $\theta_{\text{rad}} = 30^\circ$. When scanning the frequency of the fed signal, the radiation direction can be conveniently tuned, as shown in Fig. 8 with dashed curves.

In the backward-radiation (or slow-wave) regime, the main beam of radiation is directed backward, toward the source. For the metasurface designed to operate in this regime, the radiated electric field and its Poynting vector are depicted in Fig. 10. The radiated electric field has an exponential decay along the $-z$ axis. The radiation pattern of the metasurface as the working frequency increases is represented by the dotted lines in Fig. 8.

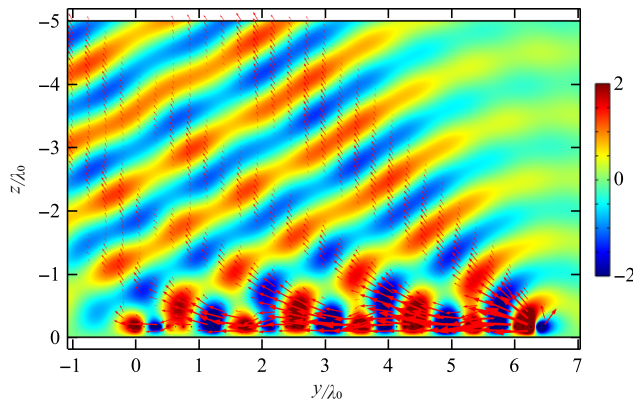


FIG. 9. The real part of the scattered electric field $\text{Re}\{E_x^{\text{sca}}\}$ [V/m] distributions over space. The red arrows represent the distribution of the Poynting vector of the scattered waves.

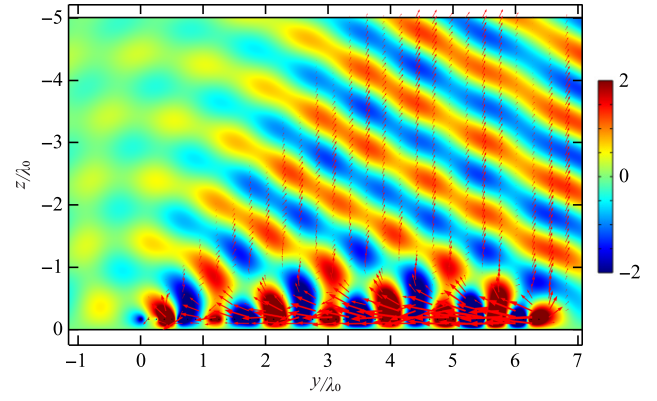


FIG. 10. The real part of the scattered electric field $\text{Re}\{E_x^{\text{sca}}\}$ [V/m] distributions over space. The red arrows represent the distribution of Poynting vectors of the scattered waves.

B. Scanning method II: Tuning the reconfigurable load impedances at a fixed frequency

The first scanning method requiring a change in working frequency typically provides narrow scanning angle ranges. When the working frequency deviates significantly from the designed frequency, the performance of the metasurface will deteriorate, especially when the expected radiation direction is very oblique. To handle this problem, the scanning method II can be used. Here, we tune the impedances of reconfigurable loads between several pre-computed sets of values while the operating frequency is fixed. The radiation patterns for the same metasurface under this scanning method are shown in Fig. 11. In this example, the radiation direction is tuned from -75° to 75° . Generally, this scanning method can cover the full angular range over the ground plane.

To fully leverage the attributes of both scanning methods, scanning methods I and II can be jointly utilized. Scanning method II provides “coarse” scanning across a broad spatial angle, making it useful for object detection, while scanning method I offers a faster scanning technique within a narrower spatial angle, ideally suited for object tracking. Theoretically, this combined approach can result in highly efficient beam scanning.

VI. MODULAR APPROACH FOR MULTIFUNCTIONAL METASURFACES

We have shown that a subwavelength strip array can effectively convert a propagating wave into surface waves. Next, we investigate how to reconfigure the array for guiding surface waves. In this study, we assume that the total number of strips is equal to $2N$. The first half of the strips is denoted as the first section, and the remaining N strips are denoted as the second section. Only the first section of the entire strip array is illuminated by propagating waves. The first section works as a surface wave launcher, while the

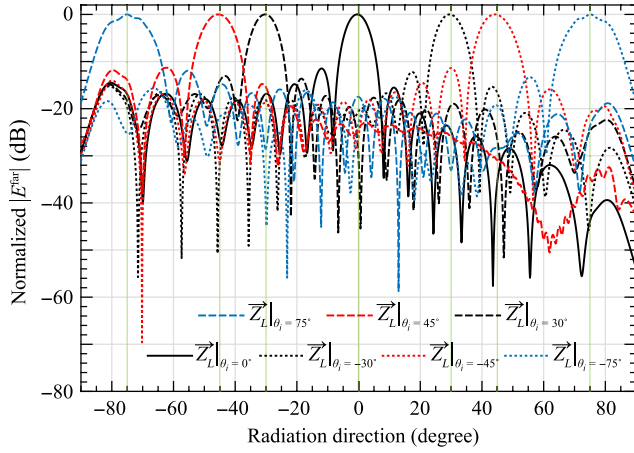


FIG. 11. Radiation patterns of the same metasurface with different sets of optimized load-impedance values. Initially, the metasurface was designed to convert an incident plane wave coming from one of the incident angles θ_i (0° , $\pm 30^\circ$, $\pm 45^\circ$, or $\pm 75^\circ$) into a surface wave absorbed by the last strip. By reciprocity, in the transmitting regime, the metasurfaces were fed from the last strip, generating leaky waves with planar wave fronts propagating towards angles $\theta_{\text{rad}} = \theta_i$. The radiation patterns are normalized by the maximum value of each pattern. The vertical green solid lines represent desired radiation directions. The operating frequency is 10 GHz.

second section works as a receiver. Here, we do not use a plane wave as an incident propagating wave anymore, as the electric field at the edge of the first section of the array significantly affects the power received by the second section. We assume that the first section is illuminated by a Gaussian beam, which should be converted into surface waves. By properly controlling the waist radius of the Gaussian beam, the field at the end edge of the first section of the strip array will be close to zero, and the effect of the reflection from the bare ground plane can be neglected. Here, for simplicity, the incident wave propagates along the z axis and is specified as [58]

$$\mathbf{E}_{\text{inc}}^g(y, z) = E_0 \sqrt{\frac{w_0}{w(z)}} e^{-\frac{y^2}{w^2(z)} - jk_0 z - jk_0 \frac{y^2}{2R(z)} + j\frac{\eta(z)}{2}} \hat{x}, \quad (9)$$

where $E_0 = 1$ V/m, $w(z) = w_0 \sqrt{1 + ((z - p_0)/z_0)^2}$ represents the spot radius for different positions along the propagation axis, the focal plane is placed on the PEC ground plane, i.e., at $p_0 = 0$, and the Rayleigh range $z_0 = k_0 w_0^2/2$, while the Gouy phase shift is given by $\eta(z) = \arctan((z - p_0)/z_0)$. The minimum input beam waist radius is set as $w_0 = Nd/6 = 13\lambda_0/12$. Here, it is assumed that the delayed fields at the edge of the axis away from the first half of the strip array, where the field amplitude falls to 0.012% of the peak value, are essentially zero.

Before delving into guiding the received surface wave, let us first assume that the strip array only has the first section (N strips in total, i.e., 52 strips). To simplify the simulation model, here the center axis of the strip array is shifted to coincide with the z axis, while the Gaussian beam is along the z axis, i.e., at normal incidence. The design process for conversion of a Gaussian beam to surface waves is similar to the above-considered plane-wave illumination. Efficient conversion of a Gaussian beam to surface waves is also achieved by globally optimizing the load-impedance distribution with the same objective function shown in Eq. (6). Equation (2) is used to obtain the induced currents.

For the case where all the strips except the last one are connected with pure reactive load impedances, the real part of the scattered field is shown in Fig. 12(a). The required load-impedance distribution (except the one in the last strip) for such conversion is shown in Fig. 13(a) with the shaded area, where the last strip is loaded by a complex impedance per unit length $2085.6 - j40280.7 \Omega/\text{m}$. The conversion efficiency calculated by Eq. (5) gives 94.6% (Matlab), while the simulation results of COMSOL Multiphysics give 94.9%. We see that in this case the whole array is resonant, as there is a superposition of forward- and backward-traveling surface waves; see the black dashed curve in Fig. 12(b).

As has been shown above, the resonance can be suppressed by introducing loss into the load-impedance densities. We also have a similar observation for Gaussian beam illumination, as depicted in Fig. 12(b). Here, we chose the real parts of all load-impedance densities (except the last one) equal $100 \Omega/\text{m}$. The optimized imaginary parts of the load impedances are depicted in the shaded area in Fig. 13(c). In this case, the optimized load impedance of the last strip is $6409.5 - j43661.0 \Omega/\text{m}$. The corresponding scattered electric-field distribution is depicted in Fig. 12(c). The efficiency found from Eq. (5) gives 82.5% (Matlab), while the simulation results of COMSOL Multiphysics give 85.2%.

VII. GUIDING THE SURFACE WAVE AND RERADIATING IT AS PROPAGATING WAVES

In practical applications, the surface wave often needs to be guided. To show a possibility to reconfigure arrays for this function, we consider strip arrays formed by two cascaded sections, where each section has N strips. Specifically, the first section is the considered above converter of a Gaussian beam to surface waves (a receiving leaky-wave antenna and a surface-wave launcher), while the second section operates as a surface-wave waveguide and/or propagating-wave launcher. Here, for simplicity, the central axis of the first section of the strip array coincides with the z axis. Of note, only the first section is illuminated by a Gaussian beam, that is, the second section receives

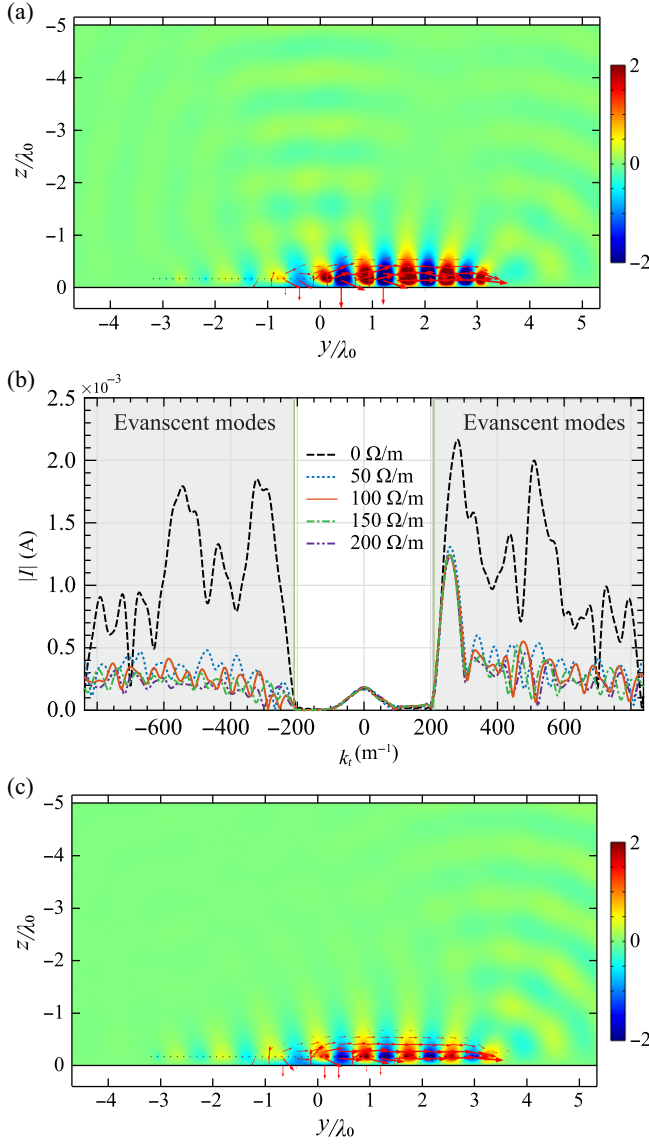


FIG. 12. Real part of the scattered electric field (with units of V/m) for lossless receiving array (a) and for an array with the load impedances having the real parts 100 Ω/m (c). Panel (b) shows the induced current amplitude as a function of the tangential component of the wave vector for load impedances with different values of resistance per unit length.

the signal predominantly from the first section and not from the incident wave. The required load impedances for conversion from the Gaussian beam to surface waves are already known, as mentioned in the previous section. The required load impedances of the second section are found by a global optimization.

A. Surface-wave waveguide

When the second section of the strip array works as a surface-wave waveguide, it should collect as much power as possible and carry it from left to right. Hence, the

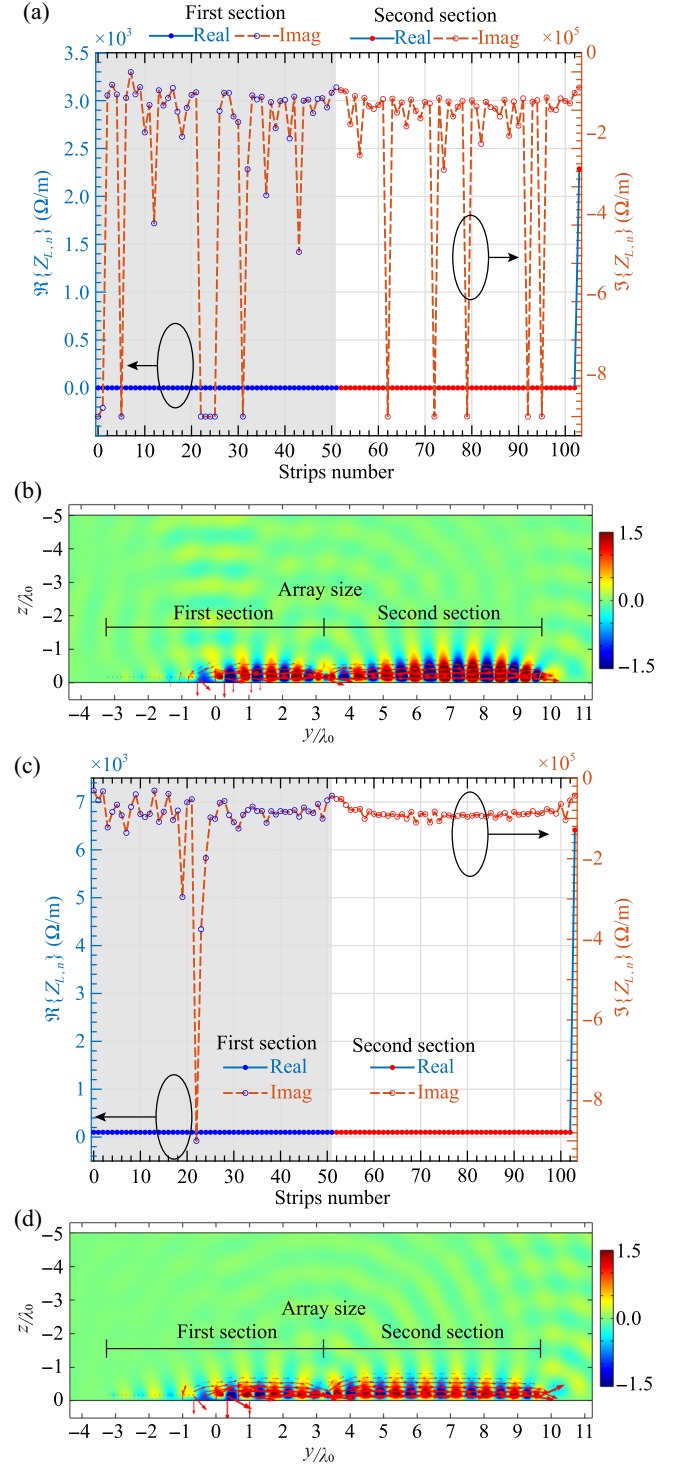


FIG. 13. (a),(b) The load impedances and the scattered electric-field distribution (with units of V/m) for lossless surface waveguide. (c),(d) The same for the real parts of the load impedances equal 100 Ω/m .

objective function is also given by Eq. (6). Now we have $2N$ strips and only the load impedances of the second section need to be optimized, as the first section is already

known. This is totally different from Ref. [29], where the optimization is done for all unit cells.

First, we consider the scenario where all the strips except the last one are loaded with lossless impedances. The first section is the same as in Fig. 12(a) with the only difference that we set the real part of the load impedance of the N th strip to zero. By optimizing the load impedances of the second section (only the last $2N$ th strip is loaded with a complex load impedance), we find the required load impedance distribution depicted in Fig. 13(a), while the scattered electric field is depicted in Fig. 13(b). The total conversion efficiency is as high as 93.1% (COMSOL), while the Matlab model gives 91.9%. We observe that this high total conversion efficiency corresponds to a resonant regime with a superposition of forward- and backward-traveling surface waves.

To dampen the resonance, lossy load impedances can be introduced. As is shown in Fig. 12(b), the backward-traveling surface waves are effectively suppressed by adding resistance per unit length equal to $100 \Omega/\text{m}$. We add the real parts equal to $100 \Omega/\text{m}$ to all previously lossless strips and optimize the load impedances of the second section of the strip array. In other words, the variables to be optimized are the imaginary parts of the load impedances of the whole second section and the real part of the load impedance of the last ($2N$ th) strip. The determined load-impedance distribution is depicted in Fig. 13(c), while the real part of the scattered electric field is shown in Fig. 13(d). The conversion efficiency is 70.4% (COMSOL model) and 68.4% (Matlab model). The fluctuations of the required load impedance of the second section in Fig. 13(c) are much smaller compared to those in Fig. 13(a), and the surface wave moves smoothly to the right direction.

B. Anomalous reflectors

Apart from the function of guiding surface waves, the second section can also work as a launcher of a propagating wave traveling in a desired direction. This way the whole device realizes the functionality of an anomalous reflector, where a wave incident on the first section is effectively reflected to an arbitrary direction when launched from the second section [59]. Regarding the function of anomalous reflection, the objective function is defined as

$$\mathcal{O} = -\max. \{ |E^{\text{far}}(\theta_r, \vec{I}(\vec{Z}_L))| \}, \quad (10)$$

where θ_r represents the desired radiation direction.

The scattered electric field of two examples that correspond to the desired radiation angles equal 75° (forward radiation) and -75° (backward radiation) are shown in Figs. 14(a) and 14(b), respectively. Here, the impedance loads of the first section of strips correspond to the result in Fig. 12(a) after removing the real part of the N th strip load impedance. The variables to be optimized are the purely imaginary-valued load impedances of the second section.

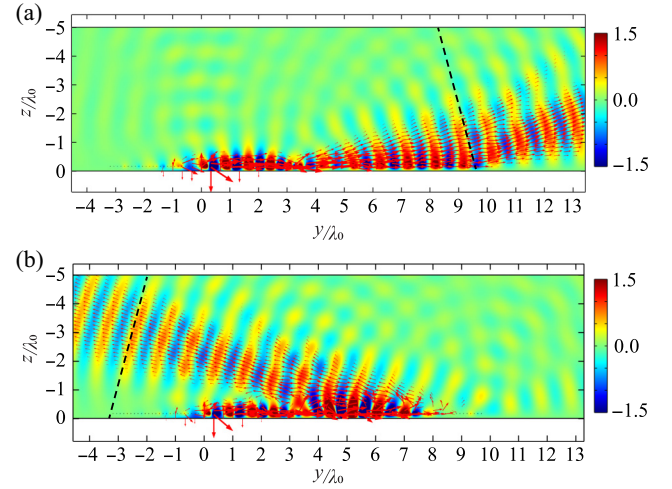


FIG. 14. The real part of the scattered electric field $\text{Re}\{E_x^{\text{sca}}\}$ [V/m] distributions over space for the desired radiation angle equal (a) 75° and (b) -75° .

The efficiency of the anomalous reflection is evaluated by integrating the Poynting vector of the scattered field along the dashed line shown in Figs. 14(a) and 14(b), and then dividing it by the input power. The anomalous reflection efficiency for 75° (forward radiation) equals 94.6%, while for -75° (backward radiation) 87.9%.

The normalized radiation patterns for the case of lossless arrays for different radiation directions are depicted in Fig. 15. By changing the load impedances, the beam can be scanned over the *full* upper space.

C. Focusing lens

The second section of the array can act also as a focusing lens for the scattered field. For this functionality, the

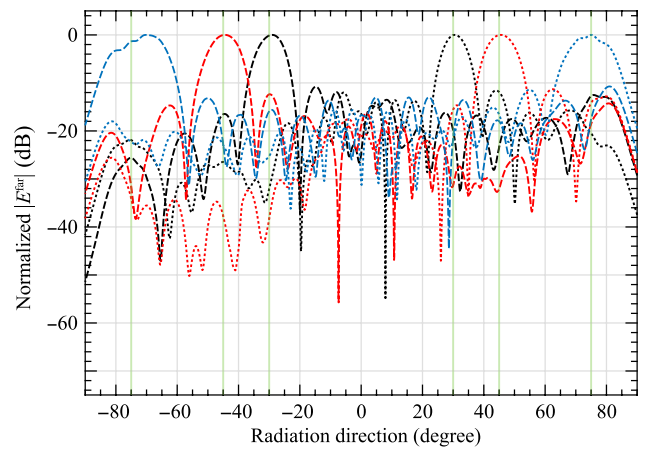


FIG. 15. Normalized far-field radiation patterns for different radiation directions. The green vertical lines indicate the desired radiation directions.

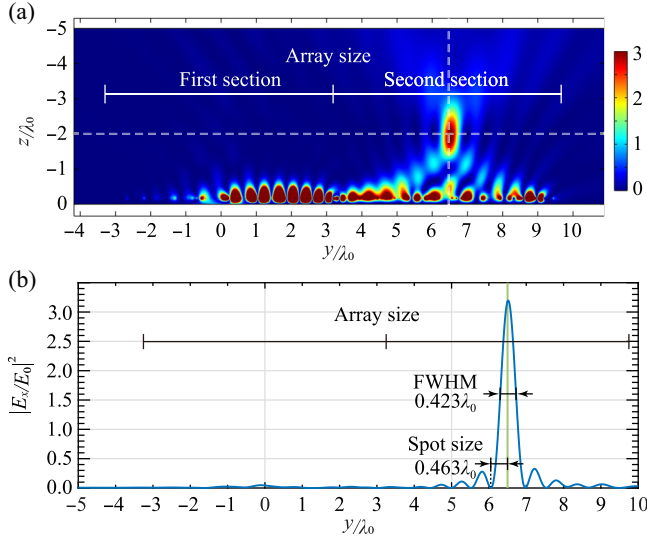


FIG. 16. Intensity of the scattered electric-field distribution $|E_x^{\text{sca}}(\hat{y}_0, \hat{z}_0)|$ [V/m]² (a) over the space and (b) along the focal plane [see the horizontal white dashed line in subfigure (a)].

objective function is defined as

$$\mathcal{O} = -\max. \{ |E_x^{\text{sca}}(\hat{y}_0, \hat{z}_0)| \}, \quad (11)$$

where the coordinate $(\hat{y}_0, \hat{z}_0)$ represents the position of the desired focal point. Here, the strips of the first section are loaded by impedances that correspond to the result in Fig. 12(a) shown in the shaded area. The variables to be optimized are the purely imaginary-valued load impedances of the second section. To show a near-field focusing possibility, the focal point is chosen at $(6.5\lambda_0, -2\lambda_0)$. The intensity of the scattered electric field is shown in Fig. 16(a). The position of the focal point is placed at the central axis [see the vertical white dashed line in Fig. 16(a)] of the second section of the strip array, and the focal plane [see the horizontal white dashed line in Fig. 16(a)] is away from the PEC ground plane by $2\lambda_0$. It is worthwhile to note that the position of the focal point can be controlled by tuning the load impedances.

The reflection efficiency is calculated by integrating the Poynting vector along the horizontal white dashed line as is shown in Fig. 16(a) and dividing by the input power, which gives 96.9%. The spot size is measured from the central maximum to the first minima in the diffraction pattern, which gives $0.463\lambda_0$. The FWHM gives $0.423\lambda_0$. The focusing efficiency is defined as the flux of the total Poynting vector over the spot size along the focal plane normalized by the input power, which gives 83.1%.

VIII. CONCLUSION

To conclude, an effective method for converting propagating waves into surface waves and vice versa has been

demonstrated, which can be used for designing propagating and surface wave converters and bidirectional scanning LWAs. The superdirective conversion is realized by using a subwavelength strip array, and the required load impedances for various conversions are found by globally optimizing the load impedances. The demonstrated high and superdirective conversion efficiency is often accompanied by a resonance of the whole array. This resonance can be suppressed by introducing loss into load impedances, if the goal is to excite a single surface wave traveling in one direction.

The designed LWAs allow two different scanning methods. One of them is realized by tuning the operational frequency, and the other by tuning the load impedances of the unit cells. From the theoretical point of view, the considered arrays have the ability to scan into all directions on the broadside of the strip array. The investigation on guiding surface waves is made for illuminations by a Gaussian beam, by cascading two sections of strips array with the same structure, that is, by fixing the load impedances of the first section and globally optimizing the second section. The first section works as a surface-wave launcher, while the second section can play the role of a surface-wave waveguide, an anomalous reflector, or a focusing lens.

Our findings promise to be a valuable asset for many applications in the field of antenna and microwave engineering. Of note, the developed modular approach is not limited to only two sections at the same plane. There can be more sections, for example, a guiding surface between the receiving and reradiating ones, and the sections can be in different planes, allowing guiding over corners and other functionalities.

ACKNOWLEDGMENTS

The authors thank Dr. Nemanja Stefan Perović, Professor Xuchen Wang, Dr. Francisco Cuesta, and Dr. Xiaojie Zhu for helpful discussions. V.A. acknowledges the Finnish Foundation for Technology Promotion. V.A. and S.T. acknowledge Research Council of Finland within the RCF-DoD Future Information Architecture for IoT initiative under Grant No. 365679.

DATA AVAILABILITY

The data that support the findings of this article are not publicly available because they contain commercially sensitive information. The data are available from the authors upon reasonable request.

- [1] X. Wan, Y. B. Li, B. G. Cai, and T. J. Cui, Simultaneous controls of surface waves and propagating waves by metasurfaces, *Appl. Phys. Lett.* **105**, 121603 (2014).

- [2] Y. Ra'di and A. Alù, Reconfigurable metagratings, *ACS Photonics* **5**, 1779 (2018).
- [3] D.-H. Kwon, Lossless scalar metasurfaces for anomalous reflection based on efficient surface field optimization, *IEEE Antennas Wirel. Propag. Lett.* **17**, 1149 (2018).
- [4] L. Zhu, R. Dai, J. Yi, X. Chen, Z. Tan, M. Lin, and S. N. Burokur, Bifunctional microwave metagrating used as reflectarray for diffraction pattern manipulation, *IEEE Antennas Wirel. Propag. Lett.* **22**, 2155 (2023).
- [5] F. Liu, D.-H. Kwon, and S. Tretyakov, Reflectarrays and metasurface reflectors as diffraction gratings: A tutorial, *IEEE Antennas Propag. Mag.* **65**, 21 (2023).
- [6] A. Díaz-Rubio, V. S. Asadchy, A. Elsakka, and S. A. Tretyakov, From the generalized reflection law to the realization of perfect anomalous reflectors, *Sci. Adv.* **3**, e1602714 (2017).
- [7] X. Wang, A. Díaz-Rubio, V. S. Asadchy, G. Ptitsyn, A. A. Generalov, J. Ala-Laurinaho, and S. A. Tretyakov, Extreme asymmetry in metasurfaces via evanescent fields engineering: Angular-asymmetric absorption, *Phys. Rev. Lett.* **121**, 256802 (2018).
- [8] S. N. Tsvetkova, D.-H. Kwon, A. Díaz-Rubio, and S. A. Tretyakov, Near-perfect conversion of a propagating plane wave into a surface wave using metasurfaces, *Phys. Rev. B* **97**, 115447 (2018).
- [9] V. Popov, A. Díaz-Rubio, V. Asadchy, S. Tsvetkova, F. Boust, S. Tretyakov, and S. N. Burokur, Omega-bianisotropic metasurface for converting a propagating wave into a surface wave, *Phys. Rev. B* **100**, 125103 (2019).
- [10] D.-H. Kwon and S. A. Tretyakov, Arbitrary beam control using passive lossless metasurfaces enabled by orthogonally polarized custom surface waves, *Phys. Rev. B* **97**, 035439 (2018).
- [11] A. Otto, Excitation of nonradiative surface plasma waves in silver by method of frustrated total reflection, *Z. Phys.* **216**, 398 (1968).
- [12] M. Neviere, R. Petit, and M. Cadilhac, About the theory of optical grating coupler-waveguide systems, *Opt. Commun.* **8**, 113 (1973).
- [13] W. Sun, Q. He, S. Sun, and L. Zhou, High-efficiency surface plasmon meta-couplers: Concept and microwave-regime realizations, *Light Sci. Appl.* **5**, e16003 (2016).
- [14] D. R. Jackson and A. A. Oliner, in *Modern Antenna Handbook*, edited by C. A. Balanis (John Wiley & Sons, Inc., Hoboken, NJ, USA, 2008), Chap. 7, pp. 325–367.
- [15] S. N. Tsvetkova, S. Maci, and S. A. Tretyakov, Exact solution for conversion of surface waves to space waves by periodic impenetrable metasurfaces, *IEEE Trans. Antennas Propag.* **67**, 3200 (2019).
- [16] S. N. Tsvetkova, E. Martini, S. A. Tretyakov, and S. Maci, Perfect conversion of a TM surface wave into a TM leaky wave by an isotropic periodic metasurface printed on a grounded dielectric slab, *IEEE Trans. Antennas Propag.* **68**, 6145 (2020).
- [17] E. Abdo-Sánchez, M. Chen, A. Epstein, and G. V. Eleftheriades, A leaky-wave antenna with controlled radiation using a bianisotropic Huygens' metasurface, *IEEE Trans. Antennas Propag.* **67**, 108 (2019).
- [18] H. Lee and D.-H. Kwon, Large and efficient unidirectional plane-wave–surface-wave metasurface couplers based on modulated reactance surfaces, *Phys. Rev. B* **103**, 165142 (2021).
- [19] M. Bodehou, K. A. Khalifeh, S. N. Jha, and C. Craeye, Direct numerical inversion methods for the design of surface wave-based metasurface antennas: Fundamentals, realizations, and perspectives, *IEEE Antennas Propag. Mag.* **64**, 24 (2022).
- [20] J. Budhu, L. Szymanski, and A. Grbic, Design of planar and conformal, passive, lossless metasurfaces that beamform, *IEEE J. Microwaves* **2**, 401 (2022).
- [21] G. Xu, G. V. Eleftheriades, and S. V. Hum, Wide-angle beam-steering and adaptive impedance matching with reconfigurable nonlocal leaky-wave antenna, *IEEE Open J. Antennas Propag.* **3**, 1141 (2022).
- [22] G. Xu, V. G. Ataloglou, S. V. Hum, and G. V. Eleftheriades, Extreme beam-forming with impedance metasurfaces featuring embedded sources and auxiliary surface wave optimization, *IEEE Access* **10**, 28670 (2022).
- [23] S. Sun, Q. He, S. Xiao, Q. Xu, X. Li, and L. Zhou, Gradient-index meta-surfaces as a bridge linking propagating waves and surface waves, *Nat. Mater.* **11**, 426 (2012).
- [24] N. Mohammadi Estakhri and A. Alù, Wave-front transformation with gradient metasurfaces, *Phys. Rev. X* **6**, 041008 (2016).
- [25] T. Arshed, S. Maci, and E. Martini, Direct synthesis of metasurfaces for efficient space- to surface-wave conversion and beamforming, *IEEE Trans. Antennas Propag.* **72**, 7793 (2024).
- [26] T. Cai, Y. Zhong, D. Liu, H. Huang, D. Wang, Y. Yang, H. Chen, and X. Lin, Observation of polarization-maintaining near-field directionality, *Prog. Electromag. Res.* **181**, 35 (2024).
- [27] D.-H. Kwon and S. A. Tretyakov, Arbitrary beam control using passive lossless metasurfaces enabled by orthogonally polarized custom surface waves, *Phys. Rev. B* **97**, 035439 (2018).
- [28] D.-H. Kwon, Modulated reactance surfaces for leaky-wave radiation based on complete aperture field synthesis, *IEEE Trans. Antennas Propag.* **68**, 5463 (2020).
- [29] J. Budhu, Near-perfect space-wave to surface-wave coupler enabled conformal space wave transporting metasurfaces, *IEEE Trans. Antennas Propag.* **72**, 2540 (2024).
- [30] J. Budhu, N. Ventresca, and A. Grbic, Unit cell design for aperiodic metasurfaces, *IEEE Trans. Antennas Propag.* **71**, 7387 (2023).
- [31] J. Budhu and A. Grbic, Perfectly reflecting metasurface reflectarrays: Mutual coupling modeling between unique elements through homogenization, *IEEE Trans. Antennas Propag.* **69**, 122 (2021).
- [32] M. Kim, Cavity-excited impedance surfaces for full-space beam steering via combination of integral equations and baffles, *Phys. Rev. Appl.* **23**, 024011 (2025).
- [33] K. Achouri and C. Caloz, in *Proc. 10th Int. Congr. Adv. Electromagn. Mater. Microw. Opt. (METAMATERIALS)* (Chania, Greece, 2016), pp. 13–15.
- [34] K. Achouri and C. Caloz, Space-wave routing via surface waves using a metasurface system, *Sci. Rep.* **8**, 7549 (2018).
- [35] A. I. Uzkov, in *Comptes Rendus (Doklady) de l'Academie Des Sciences de l'URSS* (Academy of Sciences of the Soviet Union Saint Petersburg, 1946), Vol. 53, pp. 35–38.

- [36] R. Harrington, Reactively controlled directive arrays, *IEEE Trans. Antennas Propag.* **26**, 390 (1978).
- [37] A. D. Yaghjian, T. H. O'Donnell, E. E. Altshuler, and S. R. Best, Electrically small supergain end-fire arrays, *Radio Sci.* **43**, 1 (2008).
- [38] X. Wang, A. Díaz-Rubio, A. Sneek, A. Alastalo, T. Mäkelä, J. Ala-Laurinaho, J. Zheng, A. V. Räisänen, and S. A. Tretyakov, Systematic design of printable metasurfaces: Validation through reverse-offset printed millimeter-wave absorbers, *IEEE Trans. Antennas Propag.* **66**, 1340 (2018).
- [39] S. Katoch, S. S. Chauhan, and V. Kumar, A review on genetic algorithm: past, present, and future, *Multimed. Tools Appl.* **80**, 8091 (2021).
- [40] A. Slowik and H. Kwasnicka, Evolutionary algorithms and their applications to engineering problems, *Neural Comput. Appl.* **32**, 12363 (2020).
- [41] N. Engheta, C. H. Papas, and C. Elachi, Interface extinction and subsurface peaking of the radiation pattern of a line source, *Appl. Phys. B* **26**, 231 (1981).
- [42] S. Tretyakov, *Analytical Modeling in Applied Electromagnetics* (Artech House, Norwood, MA, USA, 2003).
- [43] Y. Li, X. Ma, X. Wang, G. Ptitsyn, M. Movahediqomi, and S. A. Tretyakov, All-angle scanning perfect anomalous reflection by using passive aperiodic gratings, *IEEE Trans. Antennas Propag.* **72**, 877 (2024).
- [44] S. Tretyakov, Maximizing absorption and scattering by dipole particles, *Plasmonics* **9**, 935 (2014).
- [45] S. A. Schelkunoff, A mathematical theory of linear arrays, *Bell Syst. Tech. J.* **22**, 80 (1943).
- [46] R. F. Harrington, Effect of antenna size on gain, bandwidth, and efficiency, *J. Res. Natl. Bur. Stand.* **64D**, 1 (1960), paper 64D1-31.
- [47] Y. Li, X. Ma, X. Wang, and S. Tretyakov, Going beyond perfect absorption: Superdirective absorbers, *Phys. Rev. Appl.* **21**, 054060 (2024).
- [48] C. A. Valagiannopoulos, J. Vehmas, C. R. Simovski, S. A. Tretyakov, and S. I. Maslovski, Electromagnetic energy sink, *Phys. Rev. B* **92**, 245402 (2015).
- [49] S. I. Maslovski, H. R. L. Ferreira, I. O. Medvedev, and N. G. B. Brás, Superabsorbing metamaterial wormhole: Physical modeling and wave interaction effects, *Phys. Rev. B* **98**, 245143 (2018).
- [50] V. S. Asadchy, M. S. Mirmoosa, A. Díaz-Rubio, S. Fan, and S. A. Tretyakov, Tutorial on electromagnetic nonreciprocity and its origins, *Proc. IEEE* **108**, 1684 (2020).
- [51] V. Popov, B. Ratni, S. N. Burokur, and F. Boust, Non-local reconfigurable sparse metasurface: Efficient near-field and far-field wavefront manipulations, *Adv. Opt. Mater.* **9**, 2001316 (2021).
- [52] V. G. Ataloglou and G. V. Eleftheriades, A reconfigurable intelligent surface with surface-wave assisted beamforming capabilities, *IEEE Trans. Antennas Propag.* **73**, 5346 (2025).
- [53] S. Lim, C. Caloz, and T. Itoh, Electronically scanned composite right/left handed microstrip leaky-wave antenna, *IEEE Microw. Wirel. Compon. Lett.* **14**, 277 (2004).
- [54] D. Jiang, X. Li, Z. Fu, P. Ran, G. Wang, Z. Zheng, T. Zhang, and W.-Q. Wang, Liquid crystal-based wideband reconfigurable leaky wave X-band antenna, *IEEE Access* **7**, 127320 (2019).
- [55] O. H. Raheem Al Soad, C. Zhou, J. Fu, and Q. Wu, in *2021 Cross Strait Radio Science and Wireless Technology Conference (CSRSWTC)* (IEEE, Piscataway, NJ, USA, 2021), pp. 123–125.
- [56] D. R. Jackson and A. A. Oliner, in *Modern Antenna Handbook* (John Wiley & Sons, Inc, Hoboken, NJ, USA, 2008), Chap. 7, pp. 325–367.
- [57] F. Monticone and A. Alù, Leaky-wave theory, techniques, and applications: From microwaves to visible frequencies, *Proc. IEEE* **103**, 793 (2015).
- [58] *Wave Optics Module User's Guide* (COMSOL Multiphysics® v. 6.2, COMSOL AB, Stockholm, Sweden, 2023), pp. 153–155.
- [59] K. Achouri and C. Caloz, Space-wave routing via surface waves using a metasurface system, *Sci. Rep.* **8**, 7549 (2018).