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Systematic Design of Printable Metasurfaces: Validation Through Reverse-offset Printed Millimeter-wave Absorbers

Xu-Chen Wang, Ana J
Jian-Fang Zhai

Abstract—In this work for realizing desired sheet This methodology allows actance (capacitance and even if the conductor prop thickness and permittivity restrictions. The derived find the physical dimension complement the required surf of the method shows ex predictions, allowing the without any optimization for designing lossy and low for absorption and wavef waves. As a representativ absorbers has been app The results demonstrate material sheet resistivity, of printable metasurfaces strongly depends on the experimental validation c waves and printed using reverse-offset techniques is presented. To the best of authors' knowledge, this is the first time when reverse-offset printing has been used to provide well-working devices for short millimeter waves.

Index Terms—Metasurfaces, absorbers, impedance control, grid impedance, conductive layer, millimeter waves, reverse-offset printing.

I. INTRODUCTION

Metasurfaces, the two dimensional counterpart of metamaterials, have been extensively studied for their powerful abilities to control scattered fields, fostering their applicability for polarization control, imaging, and reflected/refracted wavefront manipulation, among others (see review papers [1]–[3]). In particular, significant work has been done for controlling reflection and absorption by using such artificial surfaces.

Metasurfaces with negligible absorption are able to manipulate the scattered wavefronts in various ways [4]–[12]. On the other hand, lossy metasurfaces can be designed for controlling not only the scattered fields but also the absorption in the

Fig. 1. (a) Schematic representation of reflective metasurfaces. (b) Local equivalent circuit of the proposed metasurface.

metasurface. Representative examples of lossy metasurfaces are perfect absorbers where all the incident energy is dissipated in the metasurface [13]–[24], and anomalous reflectors which absorb energy that cannot be sent into the desired direction [12], [25].

A common schematic of reflective or absorptive metasurfaces is shown in Fig. 1(a). In this configuration a conductive ground plane, which ensures no penetration of the fields behind the metasurface (transmission coefficient is zero), is covered by a thin conductive layer with a separating dielectric substrate in between. If the metasurface properties vary slowly in the metasurface plane, locally at every point one can construct an equivalent circuit which models the metasurface, see Fig. 1(b). Here the equivalent surface impedance of the patterned conductive layer is modelled by the grid impedance $Z_g(x, y)$, the dielectric substrate by a transmission line section, and the ground plane by the short circuit (assuming PEC response of the ground plane). It is clear that the input impedance of the metasurface can be controlled by the grid impedance, the thickness of the substrate, and its permittivity. The grid impedance of the thin layer depends on its material and thickness and, most importantly, on how it is patterned. The design challenge is to control both the real and imaginary parts of the input impedance on a sub-wavelength scale.

In recent years, there has been considerable interest in reverse-offset printing technology due to its potential in low-cost and efficient fabrication of sub-micro structures. It is frequently reported that the sheet impedance of printed ink can be controlled by the ink type and the post-processing (sintering time and temperature) [26], [27]. However, it is very

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difficult to adjust this value to the required grid impedances of metasurfaces (the required real and imaginary parts of the grid impedance drastically change for different applications) or make it spatially dependent. Moreover, in many practical situations the ink type and the substrate material and thickness are fixed due to manufacturing restrictions or other considerations. For this reason, the only universal and reliable method for controlling the grid resistance and reactance is to pattern the conductive ink tuning the equivalent, surface-averaged grid impedance. To realize the required electromagnetic response, the ink sheet resistance, capacitance, and inductance must all take the required values. If some generic patterns are used (like arrays of square patches of Jerusalem crosses), variations of any of the dimensions lead to changes of both resistance and reactance of the layer. In this situation, the required properties can be possibly achieved only using numerical optimization. In this work, we develop a design method which allows independent control and tuning of resistance and reactance (both capacitive and inductive). As a result, we find analytical formulas for fast calculation of the pattern dimensions which realize the required complex grid impedance. For simple topologies of the patches (like squares) the method is fully analytical, and for more complex shapes the determined values of the sheet capacitance can be realized with the help of numerical simulations.

The paper is organized as follows: In view of the lack of systematic approaches for the synthesis of complex sheet impedances which meet different requirements fixed by the functionalities of the metasurfaces, in Section II we provide general design rules for independent control of the grid resistance and reactance. The method gives simple and accurate analytical formulas which allow a quick estimation of the required grid for implementing the desired impedance. Furthermore, we demonstrate that this design method can accommodate various sheet resistivities provided by manufacturing techniques. As an example of the versatility of the proposed method, in Section III we apply the method to the design of perfect absorbers. In addition, interdigital capacitors (IDCs) are introduced to effectively adjust the grid reactance instead of the straight capacitive gaps. Focusing on the designs loaded with IDC structures, in Section IV, we experimentally verify our findings.

II. IMPEDANCE CONTROL METHOD

A. Controllable capacitive grid impedances

The first scenario under study is the implementation of capacitive grid impedances where the real and imaginary parts can take any arbitrary values which satisfy $\Re(Z_g) > 0$ and $\Im(Z_g) < 0$. Controlling the capacitive response using a dense array of metallic strips is a well-known solution [28]. An array of planar strips behaves as a capacitive grid when the electric field is perpendicular to the strip axis, as shown in Fig. 2(a). The capacitive impedance is introduced by the gaps between two adjacent strips. For lossy ink with pure resistive sheet impedance Z_s [$\Im(Z_s) = 0$], the equivalent impedance of the structure [see Fig. 2(a)] at normal incidence is denoted as Z_c

Fig. 2. (a) Unit cell of the capacitive grid (periodic strips). Yellow parts represent the conductive ink. (b) Unit cell of capacitive grid with inner strips for independent control of the resistivity.

(means capacitive grid impedance) and can be written as [22], [28]

$$Z_c = \frac{D_c}{D_c - w_c} Z_s - j \frac{\eta_{\text{eff}}}{2\alpha_c}, \quad \alpha_c = \frac{k_{\text{eff}} D_c}{\pi} \ln \left(\frac{1}{\sin \frac{\pi w_c}{2D_c}} \right), \quad (1)$$

where α_c is the grid parameter of this capacitive structure, $\eta_{\text{eff}} = \sqrt{\mu_0/\epsilon_0\epsilon_{\text{eff}}}$ is the wave impedance of the equivalent uniform host medium with the relative effective permittivity $\epsilon_{\text{eff}} = \frac{\epsilon_r + 1}{2}$, $k_{\text{eff}} = k_0\sqrt{\epsilon_{\text{eff}}}$ is the wave number in the effective host medium and $\frac{D_c}{D_c - w_c}$ is the geometric factor. It should be pointed out that (1) is accurate only when D_c is smaller than the wavelength.

In (1), the equivalent resistance is determined by the material sheet resistivity, as well as by the structure parameters D_c and w_c . For a homogeneous resistive film ($w_c = 0$) with the sheet resistivity Z_s , its equivalent grid resistance is still Z_s . However, due to the reduction of the area for the current flow, the effective resistance varies according to the geometric factor. We notice that $\frac{D_c}{D_c - w_c} > 1$ makes the resistance of the grid larger than the resistance of the continuous ink layer ($\Re(Z_c) > Z_s$). The second term of Z_c also contains a resistive part due to dielectric loss ($\epsilon_r'' < 0$), but for low-loss dielectric substrates it is relatively small compared to ohmic losses in the resistive sheet. On the other hand, the capacitive part is affected by both D_c and w_c . Smaller w_c or larger D_c result in a larger gap capacitance. Considering that both resistive and capacitive parts of Z_c are dependent on the structure parameters, it is difficult to match Z_c with the required grid impedance Z_{rg} at the working frequency.

An effective way to adjust the grid resistance of the structure in Fig. 2(a) is to introduce some slots (forming inner strips) along the current direction, as shown in Fig. 2(b). It is obvious that the grid resistance will increase if we exert more barriers for the flowing currents. However, those inner strips inevitably excite some magnetic field, which means they also exhibit inductive response. The equivalent impedance of an array of infinitely long resistive strips (excited along the strip axis) is written as [22], [28]

$$Z_i = \frac{D_i}{w_i} Z_s + j \frac{\eta_{\text{eff}}}{2} \alpha_i, \quad \alpha_i = \frac{k_{\text{eff}} D_i}{\pi} \ln \left(\frac{1}{\sin \frac{\pi w_i}{2D_i}} \right), \quad (2)$$

where Z_i means inductive grid impedance. This equation is applicable only for $D_i < \lambda$. From (2) we can see that if D_i is much smaller than the wavelength and w_i/D_i is not too small, α_i is close to zero, which means that the inductance contributed by the inner strips can be neglected.

In order to examine the accuracy of the analytical formulas, we compare the results with the retrieved grid impedance from numerical simulations. The simulation is done by Ansoft HFSS, where we apply periodic boundary conditions and Floquet ports for excitation. In the simulation, the grid pattern is positioned on a grounded substrate and analyzed in frequency domain. The substrate permittivity is assumed to be $\epsilon_r = 3.2(1 - 0.045j)$ with the thickness of $h = 125 \mu\text{m}$. We use only the reflection coefficient to extract the grid impedance. The input impedance of the structure can be expressed in terms of the simulated reflection coefficient R as

$$Z_{\text{in}} = \frac{1+R}{1-R} Z_0, \quad (3)$$

where Z_0 is the wave impedance of free space [29]. Since Z_{in} is the total impedance of the parallel connection of Z_g and the grounded substrate impedance Z_{gd} , we can extract the grid impedance using

$$Z_g = \frac{Z_{\text{in}} Z_{\text{gd}}}{Z_{\text{gd}} - Z_{\text{in}}}, \quad Z_{\text{gd}} = j Z_d \tan(k_d h), \quad (4)$$

where $k_d = \omega \sqrt{\mu_0 \epsilon_0 \epsilon_r}$ is the propagation constant in the dielectric substrate for the TEM mode (we consider the normal incidence case), and $Z_d = \sqrt{\mu_0 / \epsilon_0 \epsilon_r}$ is the characteristic impedance of supporting dielectric layer. μ_0 and ϵ_0 are the permeability and permittivity of free space, respectively [29].

Figure 3 shows the grid impedance of these inductive strips for different values of the ratio w_i/D_i when $D_i = \lambda_0/55$ (λ_0 is the wavelength in free space). We can see that the inductive part of Z_i is always very small (below 13Ω for $w_i/D_i > 0.1$), while its resistive part is magnified by ten times (from 35 to 350 Ω) with the reduction of the strip width to $0.1D_i$. Therefore, a dense inner-strip array behaves as a resistor whose resistance can be controlled by tuning the width of the strips.

Fig. 3. Analytical and simulated grid impedance of resistive strip array in terms of w_i/D_i . $Z_s = 35 \Omega/\text{sq}$, $D_i = 20 \mu\text{m}$ and $f = 275 \text{ GHz}$.

Fig. 4. Analytical and simulated grid impedance of the structure in Fig. 2(b) in terms of w_i/D_i . $Z_s = 35 \Omega/\text{sq}$, $D_i = 22 \mu\text{m}$, $f = 275 \text{ GHz}$, $D_c = 110 \mu\text{m}$ and $w_c = 5.5 \mu\text{m}$.

The total grid impedance of the structure in Fig. 2(b) can be obtained by replacing Z_s in (1) with Z_i in (2), leading to

$$Z_g = \frac{D_c D_i}{(D_c - w_c) w_i} Z_s - j \frac{\eta_{\text{eff}}}{2} \left(\frac{1}{\alpha_c} - \frac{D_c}{D_c - w_c} \alpha_i \right). \quad (5)$$

In order to examine the accuracy of (5), analytical results and numerical simulations are compared in Fig. 4. The results are in perfect agreement except for the reactance when $w_i/D_i < 0.3$. In this region, the analytical value of the reactance behaves in an opposite way than the numerical simulations. The analytical model accounts only for the inductive reactance introduced by the thin strips, for this reason when w_i decreases, the inductance increases and reduces the total capacitive impedance. On the other hand, in the simulations, the inhomogeneous distribution of currents near the gap is also taken into account. As w_i decreases, the blank area without an ink layer becomes larger. As a result, its capacity of collecting electrons becomes weaker (capacitance decreases). But this change does not affect the result too much because the electric field in the fringe of the gap is much stronger than elsewhere. To some extent, this enhanced capacitive reactance can balance the total reactance since it cancels out the parasitic inductive reactance caused by the inner strips. It is worthy to notice that, compared to the sharp increase of the grid resistance when w_i decreases, the capacitive reactance is almost undisturbed.

In practice, the limitations of the method will be imposed for the functionality to be implemented and the precision of the printing method. The periodicity of the structure, D_c , will be defined and limited by the discretization requirements. For example, gradient metasurfaces, where smooth impedance profiles need to be implemented, will require small periods. However, in such applications as perfect absorbers where the required impedance is homogeneous, the sub-wavelength requirements are relaxed. The line width of the gaps and strips, w_c and w_i , will be limited by the accuracy of the chosen printing method.

For dual-polarization designs, square patch array is a suitable choice due to its 90° rotational symmetry, as shown in

Fig. 5(a). The analytical expression for the grid impedance of resistive patch arrays can be written as [22], [28]:

$$Z_g = \left(\frac{D_c}{D_c - w_c} \right)^2 Z_s - j \frac{\eta_{\text{eff}}}{2\alpha_c}, \quad (6)$$

where Z_s is the sheet resistivity of conductive squares, α_c is the same as for with the single-polarized case [see (1)], according to [22].

Fig. 5. (a) Unit structure orthogonally placed strips. $T_N = 6$.

strips, the equivalent circuit is controlled by narrow strips in one unit. Therefore, we introduce strip meshes which consist of two sets of orthogonally placed strips in each square patch. The cell structure is shown in Fig. 5(b). When excited by electric fields along the x direction, the currents in y -oriented strips are practically not excited and only strips along the polarization direction affect the equivalent circuit, as shown in Fig. 6(a). However, unlike the single-polarization case, the analytical model of a dual-polarized structure cannot be derived by simply replacing Z_s in (6) with the homogeneous impedance expression Z_i in (2), since the number of strips in one cell is finite. As we know from the single-polarization case, the inductive contribution of the inner strips can be neglected if it is densely arranged ($D_i \ll \lambda$). The ultra-subwavelength meshes also reduce the parasitic capacitive coupling between the unexcited strips (y -oriented). Therefore, under this condition, the strips can be viewed as N identical resistors which are parallelly connected in the equivalent circuit, as shown in Fig. 6(b). The grid impedance of the structure is approximated as

$$Z_g = \frac{D_c^2}{(D_c - w_c)w_i N} Z_s - j \frac{\eta_{\text{eff}}}{2\alpha_c}, \quad (7)$$

where N is the numbers of strips in x or y direction. The accuracy of this formula will be verified in Section III.

B. Controllable inductive grid impedances

Here we show how to implement and control inductive grid impedances whose real and imaginary parts satisfy $\Re(Z_g) > 0$ and $\Im(Z_g) > 0$. As discussed in Section II-A, periodically arranged conductive strips behave as inductive sheets when excited by electric fields along the strip direction. According to (2), the grid inductance of a strip array can be determined by

Fig. 6. (a) Current distribution in one unit when $Z_s = 2 \Omega/\text{sq}$. (b) Equivalent circuit of the double-polarized pattern with thin meshes.

both the cell size and the strip width. Increasing the periodicity or decreasing strip width, the equivalent grid inductance increases. After tailoring the inductive part $\Im(Z_g)$, resistive grid impedance can be engineered by introducing some parallel slots in strips. These slots significantly increase the equivalent grid resistance. The unit-cell structure is shown in Fig. 7(a) with N sub-strips ($N = 3$). It is important to note that replacement of a single strip by N sub-strips so that the total width w_i is kept constant only changes the resistance of the array and has negligible effect on the already determined grid reactance. This is because the currents in strips flow along the same direction and are nearly in-phase, so that the magnetic field excited by these currents cancels out between the adjacent sub-strips (see Fig. 7(b)). The analytical formula for the grid impedance of the structure can be simply written as

$$Z_g = \frac{D_i}{Nw_r} Z_s + j \frac{\eta_{\text{eff}}}{2} \alpha_i, \quad (Nw_r \leq w_i) \quad (8)$$

where D_i is the cell size of the inductive array, α_i is given by the same expression as in (2), and w_r is the width of sub-strips. It is important to note that the period of sub-strips should be much smaller than the wavelength so that the additional inductive contribution can be neglected (surface reactance of a dense array of parallel metal strips is proportional to the array period divided by the wavelength [30].) Figure 8 presents the grid impedance of the structure in Fig. 7(a) as a function of the sub-strip width. We can see that the grid resistance increases rapidly from 30Ω to 300Ω as the width of the sub-strips reduces, while the grid reactance remains close to $+j60 \Omega$. However, we can also observe that the grid reactance slightly increases as the sub-strips become thinner. This effect can be alleviated by adding more sub-strips.

For a given operating frequency, Z_s , and sub-wavelength periodicity D_i , the width of the main strip w_i and the sub-

Fig. 7. (a) Topology of the inductive grid ($N = 3$). (b) Magnetic field in the cross section plane (x - z plane).

Fig. 8. Analytical and simulated grid impedance of the inductive grid with sub-strips (as shown in Fig. 7(a)) in terms of w_r/D_i . Here, $D_i = 180 \mu\text{m}$, $w_i = 60 \mu\text{m}$, $f = 350 \text{ GHz}$, $Z_s = 10 \Omega/\text{sq}$, $N = 3$.

strip w_r can be solved from two equations, where the real and imaginary parts of Z_g and Z_{rg} (the required grid impedance) are equated.

The above analysis is about the impedance control method for single-polarized inductive grid. In this part, we study the dual-polarization case which can be implemented by adding orthogonal strips to form a square mesh. Let us first consider the inductive mesh without sub-strips (used to control the grid resistance), as shown in the sub-figure of Fig. 9. If the ink sheet is lossless, the analytical expression of its grid reactance is the same as for the single-polarization case, $\Im(Z_g) = j \frac{\eta_{\text{eff}}}{2} \alpha_i$. However, for resistive inks, this formula for the grid reactance can be applied to dual-polarization case only for small ink resistances. It is shown in Fig. 9 that for square meshes the grid reactance is affected by the sheet resistivity of the ink. As the ink resistivity increases above approximately $50 \Omega/\text{sq}$, the grid inductance decreases for dual-polarization structures, while the single-polarization case fits very well with the analytical formula. This is due to the increasing capacitive impedance introduced by the orthogonal strips in the dual-polarization scenario. As we can see from the sub-figure of Fig. 9, for large ink resistivities ($500 \Omega/\text{sq}$ in this example) the strips orthogonal to the electric field are strongly excited, which results in capacitive coupling between these vertical strips. The existence of parasitic capacitance makes the independent

Fig. 9. Analytical and simulated grid impedance of the inductive strips in terms of the sheet resistivity of ink. Here, $D_i = 200 \mu\text{m}$, $w_i = 20 \mu\text{m}$, $f = 275 \text{ GHz}$. The red squares and blue circles represent the simulated grid impedance for single- and dual-polarizations, respectively. The sub-figure is the surface current distribution for the dual-polarization case when $Z_s = 500 \Omega/\text{sq}$.

control of grid resistance and reactance difficult in this case.

C. Limitation on applicable ink resistivity

It is important to note that the resistivity $\Re(Z_s)$ of ink layers which can be used to realize a certain grid impedance Z_{rg} cannot be arbitrarily selected and has an upper limit. In the case of lossless substrates, this upper bound can be calculated from the scenario where there are no auxiliary slots:

$$F\Re(Z_s) \leq \Re(Z_{rg}), \quad (9)$$

where F is the structural factor of the capacitive or inductive grid. F is equal to $D_c/(D_c - w_c)$, $D_c^2/(D_c - w_c)^2$, and D_i/w_i for single-polarized, dual-polarized capacitive strips, and single-polarized inductive grid, respectively. If the sheet resistivity of available inks $\Re(Z_s)$ exceeds its upper bound, introducing slots in the strips or patches becomes useless.

III. SYSTEMATIC DESIGN OF THIN PRINTABLE ABSORBERS

In the literature we can find a diversity of grounded electromagnetic absorbers which follow the topology studied in this work (see Fig. 1) [13]. A classical example is the Salisbury absorber [31] whose main limitation is its resonant thickness $h \sim \lambda/4$. With the development of metamaterials and metasurfaces, the studies have been focused on the design of electrically thin absorbers. In this section, as a demonstration of the applicability of the proposed method, we study its use for the design of thin printable absorbers.

A typical solution for the design of thin absorbers is to place an artificial lossy magnetic layer on a PEC ground plane. The thin artificial magnetic layer can be implemented as a high impedance surface by printing a conductive grid over the dielectric substrate. Here the total electric currents and the equivalent magnetic current (consequence of the differential currents on the ground plane and the grid) form a Huygens pair which provides the required resonant response

Fig. 10. Required grid impedance, Z_{rg} , for total absorption. Here, the substrate thickness is $125 \mu\text{m}$ and the dielectric constant is $\epsilon_r = 3.2(1 - 0.045j)$.

for producing total absorption [13]. In most of the cases, losses are considered as dielectric losses in the substrate layer.

Another alternative is a complex sheet impedance over a lossless grounded substrate. The role of the reactive part of the sheet impedance is to provide the magnetic response of the layer. On the other hand, the resistive part will provide the required losses for perfect absorption. As it was explained, the equivalent circuit of the system [see Fig. 1(b)] consists of a short-terminated transmission line with the characteristic impedance Z_d and a shunt impedance Z_g . A general design method for absorbers is to match the impedance of the grid Z_g to that of the surrounding space ensuring that the reflection coefficient, $R = (Z_{in} - Z_0)/(Z_{in} + Z_0)$, is zero. The input impedance, Z_{in} , corresponds to the parallel connection of the shorted transmission line and the grid impedance:

$$Z_{in} = \frac{jZ_g Z_d \tan(k_d h)}{Z_g + jZ_d \tan(k_d h)}, \quad (10)$$

Zero reflection (perfect absorption) can be realized when $Z_{in} = Z_0$. Under this condition, we can calculate the required grid impedance in terms of the frequency and substrate parameters:

$$Z_{rg} = \frac{Z_0 Z_d \tan(k_d h)}{jZ_0 + Z_d \tan(k_d h)}. \quad (11)$$

Figure 10 shows the required grid impedance for the design of a perfect absorber when $h = 125 \mu\text{m}$ and $\epsilon_r = 3.2(1 - 0.045j)$. The substrate is polyethylene naphthalate (PEN) which is commonly used in printing techniques. The permittivity of the substrate at millimeter-wave frequencies has been characterized using the method in [32]. We can see in Fig. 10 that the real and imaginary parts of Z_{rg} change with the frequency. The imaginary part of the grid impedance, $\Im(Z_{rg})$, takes positive and negative values, meaning that both inductive and capacitive responses need to be controlled for a systematic design method. In what follows, we will show the potential of our impedance control method by designing absorbers in two different operation frequency ranges from 0 GHz to 340 GHz and from 340 GHz to 600 GHz where the grid impedances are capacitive and inductive, respectively. In both designs, the dielectric substrate is the same, assumed to be chosen from the manufacturing considerations.

A. Perfect Absorbers Using Various Sheet Resistivity

The controllable grid resistance can be exploited in many interesting applications. In printing technology, precise control of the ink sheet impedance is always a challenging problem. Here, the structure shown in Fig. 5(b) is applied to realize perfect absorbers which can adapt to various industrially provided ink resistivities. The first term of (7) implies that for a fixed grid resistance, reduced ink area (smaller w_i) requires smaller surface resistance for realization of identical losses. Therefore, multiple choices of Z_s are possible if the inner strip width is engineered.

Here, we investigate the applicable range of sheet resistivity for perfect absorption at 275 GHz. According to (11), the objective grid impedance at this frequency is $Z_{rg} = 304.6 - j165.3 \Omega$. If the cell gap is specified as $w_c = 5 \mu\text{m}$ (near RO printing resolution), $D_c = 110 \mu\text{m}$ can be determined by the second term of (7) in order to accomplish impedance matching with $\Im(Z_{rg})$. For an arbitrary Z_s , the resistive part of Z_{rg} can be tuned by adjusting the strip width w_i . Figure 11 shows the analytically solved strip width as the available ink resistance Z_s increases. We can see that the design method can adapt to the ink resistivity in a large range (from 31 to 273 Ω/sq) only by varying the strip width from 2 to 17.5 μm . In order to examine the accuracy of (7), the simulated absorption coefficient which can be computed as $A = 1 - |R|^2$ is also displayed in Fig. 11. Simulation conditions are the same as that described in Section II. The nearly perfect absorption shows that the analytical formula (7) gives a very good estimation for the performance of the proposed structure.

Fig. 11. Calculated w_i in terms of accessible sheet resistivity (red line with square) and simulated absorption coefficient (blue circle). Here $D_c = 110 \mu\text{m}$, $w_c = 5 \mu\text{m}$, $N=6$.

The applicable sheet resistivity reaches its maximum value when there are no inner strips in the patch ($D_i = w_i = 17.5 \mu\text{m}$). In the other extreme case, if the strip width becomes thin enough, it is possible to use very low Z_s to obtain high grid resistance. However, the minimum applicable Z_s is restricted by manufacturing technologies due to the resolution limit of printing. One effective method to further reduce the applicable Z_s is decreasing the number of strips in the patch, while keeping w_i at its minimum value.

B. Perfect Absorbers at Different Frequencies

For perfect absorber operating at other frequencies, accurate control of both grid resistance and reactance is necessary in order to realize the required frequency-dependent Z_{rg} . In mass production of printed absorbers, the use of inks with some specified sheet resistivity (most stable and achievable in manufacturing) is expected. Here, using the grid structure depicted in Fig. 5(b), we present an absorber design method for different frequencies with a fixed ink resistivity.

According to (6), the capacitive grid impedance can be tuned by either unit gap width or the periodicity. As we know, it is not reasonable to shrink the cell gap immoderately for lower capacitive reactances due to the limited fabrication resolution. An intuitive way for solving this problem is to enlarge the array period. The sub-figure of Fig. 12 shows the calculated D_c from (7) for reactance matching when D_c/w_c is assumed to be 20. The calculated results are accurate if the grid periodicity is smaller than or compared to the substrate thickness. As the periodicity further increases ($D_c > h/0.3$), the impact of evanescent waves reflected from the ground plane must be considered [33]. The evanescent field triggers vertical electric field between the patch and the ground plane, introducing additional capacitance. A more accurate expression with correction terms for higher-order modes can be found in [34].

In order to match the grid resistance to $\Re(Z_{rg})$, we should also select a suitable w_i . The optimal strip width can be determined by equation (7) after Z_s is specified. Figure 12 shows the simulated and analytical results for the absorption coefficient for absorbers designed at three different frequencies. The sheet resistivity of the ink is assumed to be $Z_s = 35 \Omega/\text{sq}$. It is evident that the analytical model is very accurate for absorber design in a wide frequency range.

The equivalent circuit of the proposed perfect absorber can be interpreted as a parallel RLC resonant circuit, where R should be equal to the free-space impedance for perfect absorption. The bandwidth of a parallel resonant circuit is proportional to $\omega_0 L/R$ [29], where $L = \mu_0 \tan(k_0 \sqrt{\epsilon_r} h)/k_0 \sqrt{\epsilon_r}$ (for very thin substrates, $L \approx \mu_0 h$). Therefore, the bandwidth of the proposed absorber is determined by the operational frequency and the substrate thickness, as well as the operational frequency (the substrate permittivity is weakly affecting for small h). Figure 12 shows that for a fixed substrate the absorption bandwidth decreases when operating at a lower frequency. For further improving the bandwidth of this structures, one can consider the typical patterns which exhibit both capacitive and inductive behavior (e.g. loops or crosses) to generate multiple resonances in a wider band [22].

As seen in Fig. 10, the required grid impedance is inductive when the perfect absorption frequency is between 340 GHz to 600 GHz. Here, we use the inductive strip array (single-polarization) shown in Fig. 7(a) to implement perfect absorbers in this frequency range. According to (8), the grid reactance and resistance can be controlled by adjusting w_i and w_r , respectively. The sub-figure in Fig. 13 shows the solved w_i and w_r from 350 GHz to 450 GHz, assuming $D_i = 300 \mu\text{m}$ and $N = 3$. In order to verify the accuracy of these analytical so-

Fig. 12. The absorption coefficient calculated by HFSS and analytical model for absorbers with centre frequency at f_1 , f_2 and f_3 . The structure parameters are $D_{c,1} = 317 \mu\text{m}$, $w_{i,1} = 43 \mu\text{m}$; $D_{c,2} = 142 \mu\text{m}$, $w_{i,2} = 7 \mu\text{m}$; $D_{c,3} = 116 \mu\text{m}$, $w_{i,3} = 2 \mu\text{m}$. The number of strips is $N = 6$. The sub-figure shows the calculated D_c/λ_0 in terms of the operating frequency, according to (7).

lutions, absorption coefficient calculated from simulation and analytical formulas are also presented in Fig. 13, for perfect absorbers designed at 350 GHz and 450 GHz. Obviously, we can see that the analytical grid impedance in (8) gives good estimation of the performance of absorbers.

Fig. 13. Absorption coefficient obtained from analytical model and simulation for perfect absorption at $f = f_4$ and $f = f_5$. $w_{i,4} = 140 \mu\text{m}$, $w_{r,4} = 6.3 \mu\text{m}$, $w_{i,5} = 55 \mu\text{m}$, $w_{r,5} = 14.5 \mu\text{m}$, $Z_s = 25 \Omega/\text{sq}$. Sub-figure: Calculated w_i and w_r from 350 GHz to 450 GHz, according to (8).

C. Perfect Absorbers with Interdigital Capacitors

In the previous section, we have shown that increasing the lattice constant one can enlarge the capacitance when the spacing between the unit cells reaches its minimum allowed value. Here, we present another alternative that can effectively increase the equivalent capacitance of the pattern even if the lattice periodicity is unchanged. This method allows for the reduction of the unit-cell size and can be used in the implementation of gradient metasurfaces where small periodicity is necessary to constitute a smooth impedance profile.

In this method, the interdigital capacitor (IDC) is loaded in one cell as a capacitive component instead of a straight slot, as shown in Fig. 14(a). The IDC-shaped component extends the slot length and increases the equivalent grid capacitance. The value of this capacitance can be adjusted by the spacing between fingers w_f , the number of fingers N_f (in one dimension) and the finger length l_f . Figure 14(b) shows the capacitive part of the grid impedance for different finger lengths. It is evident that the capacitive impedance significantly decreases increasing l_f from 0 to 60 μm . Figure 15 displays the absorption coefficient of the absorber when tuning the finger length l_f . The frequency of perfect absorption is shifted from 125 GHz to 75 GHz while keeping the cell periodicity unchanged. Particularly, at 75 GHz, the unit-cell size is $\lambda_0/15$ which is much smaller than that without an IDC load ($\lambda_0/4$ calculated from (7)). Note that the tuning capacity of IDC is weakened as the increment of finger length (the reduction of capacitive grid impedance becomes less sensitive to the increase of l_f). In this case, one can increase N_f as well as

Fig. 14. (a) Configuration of IDC-loaded design (numbers of resistive strips is $N = 5$). (b) Grid reactance in terms of finger length l_f . Here, $D_c = 250 \mu\text{m}$, $w_f = 5 \mu\text{m}$, $l_e = 28 \mu\text{m}$ and $N_f = 18$. These results are extracted from simulations. The material used in simulation is PEC.

IV. EXPERIMENTAL VALIDATION

In the previous sections, we analytically and numerically verified our proposed design method for perfect absorption with different ink properties, and also with the same ink resistivity but at different operating frequencies. Moreover,

Fig. 15. Simulated reflection coefficient of the perfect absorbers designed at 75 GHz, 95 GHz and 125 GHz. Here, $D_c = 250 \mu\text{m}$, $w_f = 5 \mu\text{m}$, $N_f = 18$, $N = 5$ and $Z_s = 5 \Omega/\text{sq}$.

the employment of interdigital capacitors largely broadens the tuning freedom of the grid reactance, although the design method for this structure is not purely analytical. In this section, focusing on the structures with IDCs [see Fig. 14(a)], we will experimentally demonstrate important advantages of the proposed method.

A. Reverse-offset Printing Process

Electromagnetic metasurfaces with adequate line resolution even for millimeter wavelengths can be fabricated using traditional photolithographic technologies, which, however, are slow and make the product expensive. Therefore, other printing technologies such as ink-jet, screen, flexography, gravure, offset and reverse-offset printing have been or are now actively being studied for manufacturing printable electronics [26], [35]–[38]. The printing technologies mentioned above utilize conductive nanoparticle inks and can provide high throughput rates especially in a rotary form on flexible substrates. However, only reverse-offset printing seems to have the potential to provide sub-micron line resolution, needed for printing, e.g., millimeter-wave antennas or absorbers.

The samples are fabricated by reverse-offset printing technology with dimensions of $2 \text{ cm} \times 2 \text{ cm}$. Figure 16 presents the experimental reverse-offset printing equipment. The printing process can be described as follows: As the first step, a silicone coated polydimethylsiloxane (PDMS) blanket is covered with ink. In the second step the stamp cylinder, which can be made with high resolution (sub-micron structures), removes ink from the PDMS blanket. The stamp takes all the ink in the contact area. In the third step the PDMS blanket is taken to a contact with the substrate and the remaining ink is transferred to the substrate. Then follows sintering of the conductor pattern in an oven at 180°C . The ink is made up of silver nanoparticles (AgNP) and the ink film is printed on a polyethylene naphthalate (PEN) substrate. The printed and sintered samples are adhered at a layer of copper tape used as a ground plane. For more details of the printing process, see [39].

substrate. Then follows sintering of the of the conductor pattern in an oven at 180 degree centigrade.

The ink is made up of silver nanoparticles (AgNP) and the ink film is printed on a polyethylene naphthalate (PEN) substrate. The printed conductor pattern is the sintered in an oven at 180 degrees centigrade. The printed and sintered samples are adhered at a layer of copper tape used as a ground plane.

Fig. 15. Experiment

The schematic of the Fig. 16(a), which is an extender (from 90° to 180°).

Fig. 17(a), v frequency extender, and with the focal length of the antenna, and with the focal length of the antenna, respectively. collimated light into a narrow waist diameter reasonably small.

First, one antenna is adjusted to the highly reflective of the second mirror. As a reference, the reflection coefficient of the copper tape ($|S_{11,r}|$) is measured to determine incident power on the sample. The measured $|S_{11,r}|$ should be better than -4.3 dB (90-140 GHz). The reflection coefficient of the system with the sample is denoted as $|S_{11,s}|$. The reflectivity is

$$|S_{11,m}| = \frac{|S_{11,s}|}{|S_{11,r}|}.$$

Here, time gating is used to avoid the signal diffracted edge of the mirrors or other stray reflections.

In this experiment, all samples are designed based on the unit pattern in Fig. 14(a), with the same structural parameters (D_c , w_f , N_f and N) described in the caption of Fig. 15. The designed frequency of sample S1 and S2 is 95 GHz, with the ink resistivity $Z_{s,1} = 2 \Omega/\text{sq}$ ($l_f = 28 \mu\text{m}$, $w_i = 4 \mu\text{m}$) and $Z_{s,2} = 14 \Omega/\text{sq}$ ($l_f = 28 \mu\text{m}$, $w_i = 35 \mu\text{m}$), respectively. The sheet resistivity is controlled near the prescribed values by heat treatment. Characterization of ink resistivity is based on the method reported in [27]. Figure 18 shows the measurement results for samples S1 and S2. We can see that for sample S2, the measured and simulated results agree well. However, obviously, sample S1 deviated from its simulated curve with a resonance shift. This is caused by the non-uniform resistivity at different areas of the sample. That is to say, the ink resistivities in the area for material characterization and the patterned area are different due to the inhomogeneity of sintering environment or the ink thickness. On the other hand,

Fig. 17. (a) schematic of experiment setup. (b) Micro-photograph of the fabricated absorber (sample S4). (c) Photograph of experiment setup.

since the grid resistance is proportional to $\frac{D_i}{w_i} Z_s$, the difference between the measured and actual values of the ink resistivity, ΔZ_s , will be amplified by $\frac{D_i}{w_i}$ on the total grid resistance, which results in impedance mismatch. Here, the strip width of sample S1 is very thin ($\frac{D_i}{w_i} = 13.25$), so that its absorption response is very sensitive to the uneven ink resistivity. As an evident comparison, sample S2 with $\frac{D_i}{w_i} = 1.5$ is much less sensitive to ink resistivity variations. The above explanation tells us that more accurate control of ink resistivity should be achieved if the inner strip of the structure is thin.

Absorbers realized at different frequencies with the same ink resistivity have also been verified. Perfect absorption frequency of samples S3 and S4 are selected at 95 GHz

and 105 GHz, respectively. The ink resistivity of samples S3 and S4 is designed to be $Z_{s,3} = Z_{s,4} = 5 \Omega/\text{sq}$ and experimentally achieved around this value by heat treatment. The grid resistance and reactance matching are obtained by adjusting w_i and the finger length l_f , respectively. As shown in Fig. 19, the measured results agree very well with the simulated ones and perfect absorption is observed at the designed frequencies.

Fig. 18. Measured and simulated results for absorbers designed at 95GHz with different ink resistivity, $l_f = 28 \mu\text{m}$. For sample S1: $Z_s = 2 \Omega/\text{sq}$, $w_i = 4 \mu\text{m}$; For sample S2: $Z_s = 14 \Omega/\text{sq}$, $w_i = 35 \mu\text{m}$.

Fig. 19. Measured and simulated results for absorbers with ink resistivity $Z_{s,3} = Z_{s,4} = 5 \Omega/\text{sq}$, operating at 95 GHz and 105 GHz. For sample S3: $w_i = 12 \mu\text{m}$, $l_f = 28 \mu\text{m}$; For sample S4: $w_i = 10 \mu\text{m}$, $l_f = 20 \mu\text{m}$.

V. CONCLUSION

In this paper, we proposed and developed a grid impedance control method which allows one to independently design and realize required grid resistance and reactance of metasurfaces. Simple analytical formulas are derived and verified for capacitive and inductive grid impedances. The general design methodology is applicable even if we do not have explicit and enough accurate analytical formulas for the sheet reactance of arrays of complex-shaped patches (like the interdigital capacitors). In this case we can control the real part using the theoretical formulation and, independently, the imaginary part

using numerical results. In any case, the general considerations provide a good physical insight into means for independent control over all surface parameters.

The proposed technique provides a flexible and fast design method for printable metasurfaces. In particular, we have demonstrated its applicability to the design of thin perfect absorbers. It is demonstrated that multiple choices of conductor resistivity are possible for the printing process, allowing realization of the thought metasurface performance with an industrially prescribed printed conductor resistivity. Furthermore, the IDCs-loaded structures are numerically demonstrated to tune the grid reactance while keeping the unit size relatively small. The designed absorbers (loaded with IDCs) working around 100 GHz have been printed using reverse-offset printing method with sub-micron line resolution. Measurement results obtained with a quasi-optical setup show excellent agreement with the simulation results. The authors believe that this is the first time when reverse-offset printing has been used to provide well-working devices for short millimeter waves.

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