Investigations on Photoresist-based Artificial Dielectrics with Tall Embedded Metal Grids and Their Resonator Antenna Application

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Abstract— Novel photoresist-based artificial bulk dielectrics with tall embedded metal grids are investigated and their radiation characteristics are studied in detail. The embedded metal grid substantially increases (around 6 times) the effective permittivity of the low permittivity base photoresist material, and enables excitation with a simple direct microstrip feed to radiate effectively at frequencies similar to a high permittivity dielectric resonator antenna. Moreover, the grid changes the near fields inside the resonator, and introduces novel modes that are completely different than those of usual DRAs.

A set of artificial resonator antennas is designed, lithographically fabricated, and measured to demonstrate the new modes resonating at 17 GHz and 19 GHz with 2% and 6% bandwidth and 7.6 dB and 6.6 dB broadside gain, respectively, with suitable cross polarization levels of better than -20 dB. These new modes do not appear simultaneously, and can be excited separately by varying the length of the microstrip feed.

Index Terms—DRA, Meta-material, Lithography Fabrication, Miniaturized Antenna, Microfabrication.

I. INTRODUCTION

ENGINEERED artificial materials are heterogeneous materials produced by embedding inclusions in a host material to significantly change the properties [1]. These can include materials with tailored positive dielectric properties in a broad bandwidth, metamaterials with negative refractive index, and bandgap structures and frequency selective surfaces that inhibit electromagnetic propagation in certain frequency bands [2, 3]. Such materials have attracted much attention in recent years, for instance in broadband cloaking at microwave frequencies [4] and printed high refractive index terahertz materials [5]. In most cases the embedded inclusion elements are spherical or cubic particles distributed in the bulk dielectric [6], or printed strips on thin substrates that are arranged and aligned together to form a thick dielectric layer [4]. Fabrication of such assemblies can be rather difficult in a large scale due to precise alignment requirements.

Artificial materials (including metamaterials) have also attracted considerable attention in the antenna literature [7]. A cavity-backed slot antenna loaded with an artificial magnetic conductor (AMC) reflector is presented in [8]. Electromagnetic Band-Gap (EBG) structures introduced in the substrate to surround a cylindrical Dielectric Resonator Antenna (DRA) result in higher gain performance [9]. Planar artificial transmission line loading has been used to miniaturize a printed RFID antenna down to 0.33 × 0.34 [10]. A high gain silicon on-chip antenna is presented which introduces an artificial dielectric layer between the printed antenna and silicon substrate to improve the antenna gain [11]. A microstrip patch antenna on top of a substrate with metal inclusions is proposed and the effective permittivity of the dielectric is controlled by the density of inclusions [12].

In theoretical investigations of artificial dielectric materials, it has traditionally been assumed that the heterogeneous material can be replaced with a homogenous dielectric with effective permittivity and permeability [13, 14]. This effective medium method can provide valid predictions of macroscopic behaviour, however, there is some doubt on the general suitability for all geometries and types of inclusions. Embedded grids of metal inclusions, for instance, not only introduce a significant macroscopic effect on the dielectric properties of the host material, but also introduce strong local effects that vary considerably, depending on specific local grid structure. These are not predicted by the effective medium method, confirming the necessity of near field studies to understand local effects in artificial dielectric materials when used as radiators.

This paper presents engineered artificial dielectric resonators with tall embedded metal grids of various geometries and configurations. The study confirms that such artificial materials not only benefit from enhanced effective permittivity, but also have unusual near field distributions with interesting properties that could be especially beneficial when used as radiating structures. Two new radiating modes are found that are not normally attainable in homogeneous dielectric resonators. The

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radiation behaviour of the artificial resonators is predicted considering both enhanced permittivity and new boundary conditions for each mode. An example grid DRA, with a low permittivity base material with $\varepsilon_r = 2.5$, has an effective permittivity of around $\varepsilon_{eff} = 15$ resulting in a $0.28\lambda \times 0.28\lambda \times 0.07\lambda$ miniaturized antenna. The new modes are excited by simple direct microstrip feeding, providing broadside radiation patterns with 6-8 dB gain and lower than -20 dB cross-polarisation level. Samples of artificial dielectric resonators with fine featured grid structures are fabricated in thick layers using deep X-ray lithography. Measured frequency responses, radiation patterns and antenna gains confirm the theoretical results.

II. DESIGN CONSIDERATIONS

A. Geometry and Design Constraints

Fig. 1 shows a typical geometry of the engineered artificial dielectric material which consists of a host dielectric body and a tall embedded metal grid. Fig. 2 shows a set of fabricated resonator antennas based on the artificial material. The artificial material consists of a main rectangular block $A \times B$ with a height of $H_d$ and relative permittivity of $\varepsilon_d$. The grid is defined in the rectangular block by a uniformly distributed arrangement of cavities, each with a particular lateral shape (in this case, an ‘I’ shape). These cavities are filled with metal from the bottom of the structure to a height of $H_s$, to form the grid. The explosive view in Fig. 1 shows details of the lateral geometries and spacing of the inclusions for an example embedded grid. The inclusions have a length of $l_x$ and width of $l_y$, their lateral thicknesses are $T_x$ and $T_y$, and the gaps between them are $G_x$ and $G_y$, in $x$ and $y$ directions respectively.

There are many physical parameters that can affect dielectric properties of the structure. Relatively large inclusions strongly interact with the electromagnetic waves at a macroscopic level. Smaller features appear more homogenous to electromagnetic waves and individually produce less complicated interactions which can simplify analysis. Generally, the largest grid feature should be substantially smaller than the guided wavelength. For the structures considered here it is assumed that

$$I_x \leq \frac{\lambda_d}{10} \Rightarrow I_x \leq \frac{c}{10f_0\varepsilon_d}$$  \hfill (1)

where $\varepsilon_d$ is the relative permittivity of the bulk dielectric and $f_0$ is the operating frequency. The condition in (1) moves the individual inclusion self-resonant frequency well away from the artificial dielectric resonator operating frequency, and therefore does not allow the embedded grid to self-resonate and increase the dielectric loss. Applying similar constraints to the height results in

$$H_l \leq \frac{c}{10f_0\varepsilon_d}$$  \hfill (2)

Equations (1) and (2) put an upper limit on the size of the individual grid inclusions.

B. Effective Characteristics

Various techniques could be considered for approximating the effective dielectric properties of materials with embedded metal inclusions. Due to the complicated geometry and arbitrary shape of the inclusions, methods that use full wave simulations of unit cells are preferred [16]. The simulated results presented in the paper are calculated with CST Microwave Studio™ using the time domain solver. Fine structural features are modelled using high mesh density (20 mesh lines per wavelength, twice that of a typical simulation).

A simple technique is demonstrated here to estimate the effective permittivity [17]. The premise is that total transmission will occur when the grid layer is sandwiched between vacuum and some particular dielectric layer with a specific permittivity ($\varepsilon_{test}$), and therefore $\varepsilon_{eff}$ of the dielectric layer with the embedded grid is assumed to be approximately equal to $\sqrt{\varepsilon_{test}}$. Fig. 3 shows the simulation results with this method, and demonstrates control of the effective permittivity of the artificial dielectric with low permittivity base material over a large range. Without the grid, the permittivity of the base material is $\varepsilon_d = 2.5$. With the grid, the effective permittivity increases both with decreasing gap, $G$, and decreasing lateral thickness, $T$, of the embedded inclusions. Permittivity is increased substantially, up to $\varepsilon_{eff} = 14.6$ for $G = T = 30 \mu m$, realizable feature sizes for high aspect ratio microfabrication. The permittivity of the base material also plays an important role.
role in determining the effective permittivity of the artificial material. For instance, changing the permittivity of the base material over a range from $\varepsilon_d = 1.5$ to $5$ increases the effective permittivity over a range from $\varepsilon_{eff} = 8.9$ to $19.4$. This magnification effect could be useful in tuning applications.

III. RESONANT FREQUENCY AND BANDWIDTH

A. The Base Artificial Resonator Antenna

This section determines the resonant frequencies of several grid embedded artificial dielectric structures with various physical parameters operating as DRAs, some of which are shown in Fig. 2. The grid DRA (GDRA) are fed with a direct microstrip feed line, which is a straightforward technique to excite medium and high permittivity DRAs [18, 19]. The 50 $\Omega$ standard feedline is assumed on a Rogers 5880 substrate ($\varepsilon_r = 2.2$) and has a width of $W = 2.2$ mm. The feedline extends a short length, $\Delta L$, beneath the DRA, which is varied in order to match the antenna. Table I summarizes various physical parameters of the dielectric block and the embedded grid features used as the baseline GDRA for the study.

The resonant frequency of the GDRA can be approximated by a rectangular DRA element of the same overall size ($A \times B \times H_d$) but with a much higher permittivity, that is approximated using the effective permittivity approach described in Section II B. Therefore, the size of the radiating GDRA can be roughly approximated using known rectangular DRA design procedures, for example the dielectric waveguide model [20]. One of the simplest ways to match microstrip-fed DRAs is by changing the length of the feedline extending underneath the DRA, $\Delta L$ [19]. Fig. 4 shows the $S_{11}$ response of the baseline GDRA with permittivity of $\varepsilon_r = 2.5$ for different $\Delta L$ (0 mm, 0.15 mm, and 0.3 mm), demonstrating successful excitation and fundamental resonance at 17.0 GHz. The $S_{11}$ response of a higher permittivity normal DRA with the same size, but permittivity of $\varepsilon_r = 14.6$ is also shown in the figure as a reference. Two interesting characteristics are observed. First, the very low permittivity block with $\varepsilon_d = 2.5$, and normally extremely hard to excite, is now resonating at the frequency of a normal DRA with the same dimensions, but with a much higher permittivity. In fact to the knowledge of authors, this is the lowest permittivity DRA reported to be successfully excited at microwave frequencies, of course with the help of the embedded metal grid (low permittivity DRAs were excited with $\varepsilon_r \approx 4$ for SU8 materials [21, 22]). Second, the antenna shows a second close resonance at 19.2 GHz, which is not a usual rectangular DRA resonant mode. This behavior is further explained in the following sections. Nevertheless, both modes of the example GDRA have narrower bandwidths compared to the representative high permittivity DRA.

Therefore, at least one effect of the embedded metal grid is to increase the effective relative permittivity by increasing the electric flux density of the host material locally in the small-gap regions of adjacent metal grid elements, which is equivalent to increasing the dipole polarization in these regions. The proposed ‘I’ shaped inclusions can be viewed as adjacent rectangles, with sides that are “cut” to prevent closed current loops that might result in unwanted magnetic polarization. As a result, the embedded metal grid can significantly increase the effective permittivity of the low permittivity block, in this example by a factor of 5.8, while paramagnetic effects are minimized.

| TABLE I | VARIOUS PARAMETERS OF THE BASE GDRA |
|----------------|----------------|----------------|----------------|
| | $A$ (mm) | $B$ (mm) | $H_d$ (mm) | $\varepsilon_d$ |
| | 4.9 | 5.1 | 1.5 | 2.5 |
| Embedded grid feature dimensions | | | | |
| | $I_x$ ($\mu$m) | $I_y$ ($\mu$m) | $H_l$ (mm) | $T = G$ (Î¼m) |
| | 600 | 400 | 1.2 | 30 |
TABLE II

<table>
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<tr>
<th>Thickness,</th>
<th>Resonant Frequency (GHz)</th>
<th>Max Coupling (dB)</th>
<th>BW (%)</th>
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<td></td>
</tr>
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<td>4.2</td>
</tr>
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<td>4.2</td>
</tr>
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<td>17.5</td>
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<td>4.6</td>
</tr>
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<td>-34.9</td>
<td>5.3</td>
</tr>
<tr>
<td>0.3</td>
<td>20.5</td>
<td>-18.1</td>
<td>6.1</td>
</tr>
</tbody>
</table>

B. Effects of the Grid on Antenna Performance

Assuming that increased effective permittivity results from increased dipole polarization in the gaps between adjacent inclusions of the GDRA, the permittivity should decrease by increasing the gaps ($G_x$ and $G_y$), and also result in a higher resonant frequency. Fig. 5 shows the simulated resonant frequency of the structure when the gaps are increased in several steps from 30 μm to 300 μm, while the lateral thickness is kept constant at 30 μm. As expected, this results in the fundamental resonant frequency of 17.0 GHz shifting to 23.8 GHz. Note that all other parameters of the base antenna are kept constant, except for the $G = 150 \mu m$ and 300 μm cases, where the number of inclusions has changed to 6×9 and 5×7 in order to contain the inclusions inside the GDRA block. Likewise, an increase in the lateral thickness of the inclusions ($T_x$ and $T_y$) also tends to decrease the polarization, therefore decreasing the effective permittivity and increasing the resonant frequency. However, the effect of inclusion thickness is not as significant as the effect of inclusion gaps. Fig. 5 also shows the resonant frequency of the structure when the lateral thicknesses are increased in several steps from 30 μm to 300 μm, while the gaps are kept constant at 30 μm, resulting in the fundamental resonant frequencies of 17.0 GHz shifting to 18.4 GHz. The 30 μm minimum gap widths and lateral wall thicknesses are challenging for high aspect ratio microfabrication in 1.0-1.5 mm layers, but are achievable as discussed in Section V.

The effect of metal grid height ($H_t$) on antenna performance is summarized in Table II. Decreasing $H_t$ results in an increased resonant frequency of the first mode, from 16.5 GHz for a full height 1.5 mm metal grid, to 20.5 GHz for a partial height 0.3 mm metal grid. In all cases, the height of the dielectric block $H_d = 1.5 \text{ mm}$ remains constant. Table II also shows the maximum coupling ($S_{11}$ at the resonant frequency) and the bandwidth of each example. All examples are well coupled with $\Delta L = 0.2 \text{ mm}$. The increased resonant frequency is primarily a result of reduction in the effective height of the GDRA. The bandwidth increases from 4.2% for the tall grid to 6.1% for the short grid. This is due to increasing thickness in the grid-less upper block on top of the gridded lower portion, effectively creating a two-segment rectangular DRA with a high permittivity lower segment and a low permittivity upper segment. The multi-segment technique is a known method to increase the bandwidth of DRAs [23, 24].

Increasing the bulk permittivity ($\varepsilon_d$) from 2.5 to 5 while keeping the grid constant also significantly reduces the resonant frequency of the first and second modes, from 17.0 GHz to 14.0 GHz and from 19.2 GHz to 16.6 GHz, respectively. This feature might be very useful if a tunable bulk material is used as the host dielectric, enabling the antenna to be tuned, in this case not from 2.5 to 5 but effectively magnified from 14.6 to around 20.

As described, the embedded grids introduce a second resonant mode close to the dominant mode which is excitable by a $\Delta L$ from 0 to 0.3 mm. The second mode, shows similar frequency variation characteristics as the first mode with respect to gap width and lateral dimension variation, which are omitted here for the sake of brevity. The near fields of these resonant modes are further examined to better understand the radiation properties and new boundary conditions.

IV. NEAR FIELDS AND RESONANT MODES

A. First Resonant Mode

For a microstrip-fed rectangular DRA on top of a substrate, the dominant mode is $TE_{111}$ [19], as shown in Fig. 6. In the dominant mode the electric fields are transverse and mainly circulate in the XZ plane, while the magnetic fields are directed primarily in the Y direction. Therefore the resonator acts like a magnetic dipole in the Y direction. The fields in the substrate underneath the DRA are also of the same general pattern, making the substrate effectively a part of the resonator. In many
cases, the substrate thickness is small compared to the height of the DRA and the permittivity of the substrate is lower than the permittivity of the DRA, especially for relatively low permittivity DRAs to increase the coupling. Therefore the effect of the substrate on the overall resonating fields is usually small, and can be accounted for in the design procedure as an increase in the effective thickness of the DRA.

Embedding the metal grid in the GDRA body, completely changes the boundary conditions inside the resonator, and therefore changes the near fields of any resonance. Fig. 7 shows the simulated near fields of the microstrip fed baseline GDRA of Table I ($\varepsilon_{eff} = 14.6$) for the first resonant mode, which clearly are very different than those of the normal DRA described in Fig. 6. This can be interpreted as a change in the effective boundary conditions, which introduces new radiation modes. Fig. 8 shows the magnetic fields of the first resonant mode at the middle of the artificial resonator in more detail, and compares it with the same cut plane of a normal high permittivity DRA. While the magnetic fields inside the substrate are in the $Y$ direction and similar to the usual $TE_{Y111}$ mode of rectangular DRAs, the magnetic fields inside the artificial resonators circulate in the $YZ$ plane, contrary to normal DRA where the magnetic fields in the resonator are in the $Y$ direction parallel to the fields inside the substrate. In fact, the embedded metal grid does not allow the excitation of the $TE_{Y111}$ mode inside the resonator and instead, directs the magnetic fields in the $Z$ direction, shaping a new mode that is not normally possible with a homogeneous high permittivity material. Figs. 9(a) and 9(b) illustrate the magnetic field vectors in the $YZ$ plane and the electric field vectors in the $XY$ plane at the middle of the resonator. As shown, the fields inside the resonator are similar to two back-to-back $TE_{2,5,1}$ modes of a $B \times H_d$ resonator with the thickness of $A/2$. In other words, the vector fields are similar if a $B \times 2H_d \times A/2$ resonator in its dominant mode is broken in the middle, both pieces turned 90 degrees, and connected together to shape the new resonator. Fig. 10 shows the average magnitude of all components ($x$, $y$, and $z$) of the electric fields in the $XY$ plane and the magnetic fields in the $YZ$ plane at the middle of the resonator in more detail. The electric fields are prominent between the edges of inclusions, contributing to the high effective permittivity of the overall resonator, while the magnetic fields are similar to normal DRAs in the substrate and altered by the grid inside the resonator to form the new modes as discussed.
To examine this interpretation further, Fig. 11 illustrates the simulated electromagnetic fields of the dominant mode of a normal (no grid) higher permittivity ($\varepsilon_r = 14.6$) DRA element with the same dimensions of $A \times B \times H_d$, and with two completely metalized sidewalls inserted in the XZ plane at $y = \pm B/2$. The structure resonates at 16.6 GHz, and has both resonant frequency and near field distribution similar to the first mode of the low permittivity GDRA with embedded metal grid. Therefore it can be inferred that embedding the metal grids changes the resonant electromagnetic fields inside the DRA element, in the first resonant mode from TEY$_{111}$ to resemble two back-to-back resonators with dominant TE Y$_{1,0.5,1}$ mode. However, this equivalent DRA element with metalized walls has only one resonant mode (in this case around 17 GHz), and does not produce the second resonant mode of the proposed antenna (in this example at 19.2 GHz).

B. Second Resonant Mode

Fig. 12 shows the simulated electromagnetic near fields of the second mode of the GDRA structure excited with microstrip feed. The fields inside the resonator are consistent with the behavior observed with the first resonant mode of the artificial resonator as seen in Fig. 7. Fig. 13 shows the magnetic fields of this new mode at the center cut-plane of the resonator in more detail, and can be compared with Fig. 8. Here again the embedded metal grid does not allow the excitation of the TE$^y$ modes inside the resonator, and rather directs the magnetic fields in the Z direction, shaping another novel mode that is not normally possible with a homogeneous high permittivity substrate.
material. Figs. 14(a) and 14(b) illustrate the magnetic field vectors in the YZ plane and the electric field vectors in XY plane at the middle of the resonator. For this mode, the fields inside the resonator are also similar to two back-to-back $TE_{2,0,5,1}$ modes of a $B \times H_d$ resonator, but now with thickness less than $A/2$, reduced by the inclusion width. Essentially, the finite inclusion dimension is directing the fields to circulate in a shorter path through the grid, shifting the resonance frequency of the second mode slightly higher (around 10%) than the first mode. The vector fields are similar if a $B \times 2H_d \times (A - a)/2$ resonator in its dominant mode is broken from the middle, turned -90 degrees and put back to back at a distance, $a$, to shape the new resonator, where $a$ is approximately $I_y + 2G_y$ (460 μm in this case). To examine this interpretation, Fig. 15 illustrates the simulated electromagnetic fields of the dominant mode of a regular DRA element with the same dimensions of $A \times B \times H_d$ but with a higher permittivity of $\varepsilon_r = 14.6$ and two metallic walls at $y = \pm 250 \mu m$. The structure resonates at 19.4 GHz, and therefore has both frequency response and near field distribution similar to the second mode of the low permittivity GDRA with embedded metal grid.

To summarize, the overall effect of the embedded metal grid is twofold. Firstly, it substantially increases the effective relative permittivity of the structure by increasing electric dipole polarization through interaction with metal inclusions inside the DRA body, and therefore enabling very low permittivity materials to effectively radiate, miniaturizing the antenna. Secondly, it produces complicated new resonance modes that can be controlled by the grid and separately excited by small changes in the feedline length under the resonator. In the resonator, these new resonance modes have primary field orientations directed by the grid, and flipped 90 degrees from their usual directions, while maintaining close to usual field patterns in the substrate.

C. The Effects of Different Grids with Different Number of Inclusions

Changing the number of inclusions can change the resonant modes of the GDRA, and some of these effects are summarized in Table III. In all cases, the size and permittivity of the low permittivity GDRA body, and also the substrate and the feedline parameters are kept constant to solely investigate the effects of the number and pattern of inclusions. The GDRA with the $7 \times 10$ grid is considered as the benchmark. For each set, the $\Delta L$ is optimized between 0 to 0.3 mm to achieve the highest coupling to the antenna for both modes. The resonance frequency and bandwidth of both modes, the ratio of the first and second resonance, and the ratio of the resonance frequency of both modes to the benchmark GDRA are shown in Table III. The first part of Table III shows the results of different grids when the lengthwise number of inclusions in the $x$ direction is changed from 7 to 4. This decrease in the number of inclusions increases the resonance frequency of both modes up to around 30%. However, the grid is still producing the second mode resonance, and the change is approximately the same for the first and second mode. Moreover, the ratio of the resonance frequency of both modes remains the same at around 1.12 (the second mode is resonating at approximately 12% higher frequency than the first mode for all grids). It is worth mentioning that the optimized $\Delta L$ is constant for all these cases at 0 mm and 0.3 mm for the first and second mode, respectively. By decreasing the number of inclusions, the obtained...
bandwidth for both modes is increased, from 3.7% to 5.3% for
the first mode and from 2.3% to 5.1% for the second mode. It
can be concluded that the decrease in the number of inclusions
in the $x$ direction, effectively decreases the effective
permittivity of the resonator, and therefore increases the
bandwidth and the resonant frequency of both modes.

The second part of the Table III shows the results of the
different grids when the widthwise number of inclusions in the
$y$ direction is decreased from 10 down to 4. It should be
mentioned that for all samples in this part, the grid is moved
laterally into the center of the DRA in the $y$ direction. The
effects in this case are totally different compared to the effects
of varying the number of inclusions in the $x$ direction, and can
be divided into two sets. First, from 10 inclusions down to 6,
and then from 6 down to 4. In the first set an interesting
phenomenon is observed. The first resonant frequency remains
rather constant, only increasing marginally up to 4%, while the
second resonance increases more as expected, up 16%. In other
words, the ratio of the first and second resonance frequencies
were constant at around 1.12 for the first set of grid
variations, are now increased from 1.14 to 1.26. In other words,
the frequencies are getting further apart. Moreover, the
bandwidth of the second mode decreases to only 1.1%.
Therefore reducing the number of the inclusions in the $x$
direction it can be excited effectively and the resonant modes vanish.

<table>
<thead>
<tr>
<th>Grid $N_x \times N_y$</th>
<th>First Resonance Frequency $f_{r,1}$ (GHz)</th>
<th>First Resonance Bandwidth (%)</th>
<th>Second Resonance Frequency $f_{r,2}$ (GHz)</th>
<th>Second Resonance Bandwidth (%)</th>
<th>$f_{r,2}/f_{r,1}$</th>
<th>$f_{r,1}$ Relative to 7×10 GDRA</th>
<th>$f_{r,2}$ Relative to 7×10 GDRA</th>
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<td>-</td>
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</table>

* $N_x$ and $N_y$ are the number of inclusions in $x$ and $y$ directions, respectively.

Finally, it should be stressed that in all these cases, the size
of the low permittivity GDRA body is kept constant at 5.1 mm × 4.9 mm. When the smaller grids are used, one might think that
the size of the GDRA can be reduced accordingly. However, it
should be remembered that the grid only facilitates the
excitation of special modes inside a very low permittivity DRA,
and by changing the DRA body, the resonant frequency will
change accordingly. For the samples here, for example, the
GDRA size can be reduced by approximately 10% while still
exciting these new modes.

V. LITHOGRAPHY FABRICATION OF THE GRID DRAS

Besides the constraints described in Section 2, i.e. (1) and (2),
fabrication techniques also put certain limitations on the
physical parameters of a realizable artificial material. The bulk
block material is likely a polymer with low relative permittivity
(i.e. $2 < \varepsilon_d < 4$). Moreover, most fabrication methods are
limited in realizable vertical to lateral aspect ratio features,
usually between 10 and 100 depending on the technique. In the
proposed structure, the smallest lateral feature of the embedded
metal grid is $T = \min(T_x, T_y)$. The aspect ratio of the structure
is $AR = H_d/T$ and therefore

$$T \geq H_d/MAR$$

where $MAR$ is the maximum realizable aspect ratio. The same
limitation is also applied to the gaps, $G$.

Deep X-ray lithography (XRL) is used here to demonstrate
precise fabrication of prototype GDRA structures. This method
produces among the highest aspect ratios available in
microfabrication, with excellent structure quality in resists up
to a few millimeters thick [25, 26]. An $MAR = 50$ is fairly
conservative with deep XRL, therefore $Min (T) = H_d/50$ has
been assumed for the prototypes. Finest fabricated structures,
therefore, have heights up to $H_d = 1.5$ mm dielectric thickness,
while the metal widths and gaps are as small as 30 μm, also with
excellent lateral precision and smooth vertical sidewalls. Other
methods, such as deep UV lithography with very sensitive
resists [27] might also be applicable for lower MAR.
The fabrication process is illustrated in Fig. 16. A 2.7 μm thick circular titanium membrane supporting 30 μm thick gold absorbers representing the grid layouts is patterned. Poly-Methyl-Methacrylate (PMMA) sheet is fly-cut down to 1.5 mm thickness, and glued to a silicon wafer. The silicon wafer is coated with a 2.5 µm thick oxidized titanium layer to facilitate later release of the GDRAs and act as an electroplating seed layer. The mask is then used to irradiate PMMA to fabricate the polymer body portion of the antennas. After exposure to X-rays for several hours, the irradiated parts of the PMMA are removed in developer, to expose vertical cavities for metallization of the grid inclusions. A nickel sulfamate electrolyte is used to electroform the metal inclusions up to 1.2 mm. More details on high aspect ratio electroforming may be found in [28]. Finally, the GDRAs are released by dissolving the silicon wafer in KOH for a few hours.

Fig. 17 shows the Scanning Electron Microscope (SEM) micrograph of the metallic parts of the fabricated structure in a 3D view. Smooth and vertical side walls, precise features, and complete metallization of individual grid elements are apparent. Figs. 18 and 19 show details of individual elements and gaps along with measurements. Better than 2% dimensional accuracy is obtained for the inclusion elements, while the lateral features and gaps vary by 5 to 10%. It should be remembered that these are 30 micrometer lateral cavities realized within a PMMA block of >1 mm thickness, representing an aspect ratio of around 40 while maintaining sidewall verticality of better than 89 degrees. Additionally, the cavities are electroplated in nickel, up to thicknesses on the order of 1 mm. A photograph of a set of fabricated GDRAs can be seen in Fig. 2.

Detailed simulations show that GDRAs are very scalable and can theoretically operate between 2 GHz and 70 GHz, and beyond. Therefore the major limiting factor for their operation is the fabrication technique. The maximum thickness of the polymer block attainable with deep XRL is 2-3 mm, with a practical minimum of ~50 μm features at these thicknesses. It can be therefore assumed that GDRAs operating from ~10 GHz well into the 60 GHz range and beyond can be fabricated by deep XRL. Other fabrication techniques might also be used to fabricate GDRAs with less demanding aspect ratios. For instance UV lithography may be used to fabricate thinner antennas that can potentially work at higher mm-wave ranges.
VI. REFLECTION COEFFICIENT AND PATTERN MEASUREMENTS

In order to experimentally study the GDRAs, two samples of the base antenna design are selected and their impedance characteristics and radiation properties measured. The samples are fed using a 50 Ω, 2.2 mm wide microstrip line printed on a 0.78 mm thick, low-permittivity microwave substrate ($\varepsilon_r = 2.2$; $\tan \delta = 0.0009$).

The $S_{11}$ frequency response of the antennas is measured using a universal test fixture (Anritsu 3680V) without additional connectors. In order to de-embed the loading effect of the test fixture and cables on the input impedance of the antenna, a standard calibration was performed over 1-40 GHz. The $S_{11}$ measurement setup, the feeding, and the GDRA antenna is shown in Fig. 20. The antenna is fixed on the microstrip line using silicone adhesive and the reflection coefficient of the antenna is measured, while varying $\Delta L$ to obtain a sharp resonance. Fig. 21 shows the results compared to simulations. The measured first resonance occurs at 16.9 GHz and the second at 19.1 GHz. The measured results agree well with the simulations (resonances at 17.0 GHz and 19.2 GHz, respectively) and verify the concept. The bandwidths of the first and the second mode are approximately 2% and 6%, respectively. The measured bandwidth of the first mode is slightly lower, while the bandwidth of the second mode is larger than the simulations.

Figs. 22(a) and 22(b) show the radiation patterns of the antenna at 17 and 19 GHz in (a) E and (b) H planes. The maximum gains are 7.6 dB and 6.6 dB, respectively. In the E-plane, the pattern of the first mode has a small dip at the broadside that is probably due to the ground plane effect. Overall, both patterns are slightly skewed (around 10-30 degrees) in the direction opposite to feeding, which is a known
phenomenon for asymmetric direct-microstrip-fed low and medium permittivity DRAs [19]. On the other hand, the patterns in the H planes have high symmetry as seen in Fig. 22(b) resulting in an excellent broadside pattern with better than 20 dB cross polarization level.

VII. COMPARISON WITH ORDINARY DRAS

As discussed in Section III-A, the frequency response of a GDRA is similar to that of a higher permittivity normal DRA, and simulation results indicate that this is generally independent of the frequency of operation, at least at microwave and lower mm-waves. Fig. 23 demonstrates the generality and scalability of the GDRA concept and presence of the two fundamental resonant modes at different operating frequencies, as the sizes of the base GDRA block and inclusion features are scaled by certain factors. The $S_{11}$ response and first mode resonant frequencies of the antennas are illustrated in Fig. 23 using a logarithmic scale. The response is similar while the resonant frequency is scaled with roughly the same factor, from 2 to 70 GHz. The second mode at slightly higher frequencies is evident as a small dip in all responses. For the last two samples (GDRA×1/2 and GDRA×1/4), thinner substrates are used to suppress excitation of higher order modes in the 50 ohm microstrip feed. This scalability is strongly similar to that of normal DRAs.

The bandwidths of the modes (2%-6% according to Table III) are smaller than equivalent DRAs (typically ~10%), but can be controlled by adjusting the number of metal inclusions and their dimensions. The radiation patterns and realized gain of the GDRA antennas are also very similar to that of equivalent DRAs. For instance, the simulated realized gain of the antennas described in Fig. 4 are 7.7 dB for the GDRA (with $\Delta L = 0$ mm) and 7.3 dB for the normal DRA (with $ed = 14.6$), and their patterns are both symmetrical in H planes and slightly skewed in E planes, as discussed.

Finally, the radiation efficiency of a typical medium permittivity and low loss DRA (e.g. Rogers RT/duriod 6010, $\varepsilon_r = 10; \tan \delta = 0.002$) is as high as 95% at Ka band and below [29]. To realistically calculate the radiation efficiency of a lossy GDRA to compare with normal DRAs, it is assumed that the PMMA block has a dielectric loss of $\tan \delta = 0.007$ at 17-19 GHz [30]. It is also assumed that the electroplated nickel grown inside the polymer metal has smooth vertical sidewalls, with a conductivity of $\sigma = 1.266 \times 10^7$, while a 10 $\mu$m thick rough layer with mean roughness of 1 $\mu$m, resulting in a reduced conductivity of $\sigma = 5.236 \times 10^6$, is put on top of all the inclusions. The assumed nickel conductivities are obtained from measurements of a similar process [31], and are utilized to calculate the reduced conductivity of the rough nickel using the CST surface roughness calculator. It is also supposed that the permeability of nickel is $\mu \approx 1$ at higher than 10 GHz [32]. The simulated radiation efficiency of this lossy base GDRA at the first mode (17 GHz) is 88%, which is comparable to normal DRAs with similar dielectric loss. At the second mode (19.2 GHz) the efficiency decreases to 83% due to slightly higher metallic and dielectric losses. The efficiency of the base GDRA increases to 91% and 86% for the first and second modes, respectively, if copper is used instead of nickel for electroplating. The group is currently optimizing necessary processes to fabricate high efficiency GDRAs with copper at mm-wave frequencies.

VIII. CONCLUSIONS

Novel photoresist based high-permittivity artificial dielectrics realized lithographically by embedding thick metal grid structures were theoretically and experimentally investigated as radiating dielectric resonator antenna (DRA) elements. The experimental results show that the proposed grid DRA antennas (GDRAs) can be efficiently excited with a simple direct microstrip feed and resonate at frequencies similar to those of much higher permittivity DRAs, as a result of the grid of tall metal inclusions with small lateral features embedded inside the bulk photoresist material. It has been found that the grid also significantly modifies the resonator near fields in comparison with ordinary DRAs with no grid, and introduces two new modes that are completely different than those of high permittivity rectangular DRAs, but still radiate quite effectively with broadside patterns and 6-8 dB gain. For the measured GDRAs, these modes resonate at 16.9 GHz and 19.1 GHz, and can be controlled by altering the embedded grid. Comparison between the GDRAs and ordinary DRAs in terms of frequency response and scalability, radiation characteristics, and efficiency concludes the study.

In contrast to high permittivity materials, metals and low permittivity materials such as polymer resists are standard materials in many fabrication techniques including standard lithography processes. Therefore enhancement of the effective permittivity of these low permittivity materials by embedding metal grids not only provides new performance characteristics, but also gives designers a practical method to control local electrical properties of the structure, and tailor the near fields of the artificial material.

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