Mutual Coupling Reduction Between Neighboring Strip Dipole Antennas Using Confocal Elliptical Metasurfaces

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Outline

- Introduction and Motivation
  - Mantle Cloaking

- Formulation and Theory
  - Formulation of the Scattering Problem in terms of Mathieu Functions
  - Optimum Required Reactance

- Cloaking of Elliptical Structures and Strips

- Reduction of Mutual Coupling
  - Strip Dipole Antennas
  - Strip Monopole Antennas

- Conclusions
Cloaking with a Metasurface

- **Mantle Cloaking**
  - Based on scattering cancellation
  - An ultrathin metasurface
  - Anti-phase surface currents
  - Suppression of the dominant mode

1-D and 2-D Periodic Structures

- Vertical Strips
- Mesh Grids
- Capacitive Rings
- Patch Arrays

Cloaking using Graphene

**Dielectric Cylinder**

**PEC Cylinder**


Elliptical Cloak Designs at Microwaves
Elliptical Cloak Designs at THz Frequencies

Graphene nanopatches

SiO₂

PEC

H

E

a₀

b₀

k

φ

THz

Elliptical Cloak Designs at THz Frequencies

Graphene nanopatches

SiO₂

PEC

H

E

a₀

b₀

k

φ

THz
Formulation of the Scattering Problem

\[ E^i_z = \sqrt{8\pi} \sum_{n} j^{-n} \frac{I_{pm}(q_0, u, n)}{N_{pm}(q_0, 0)} S_{pm}(q_0, v, n) S_{pm}(q_0, \varphi, n); \ u > u_0 \]

\[ E^s_z = \sqrt{8\pi} \sum_{n} j^{-n} a_{pm} H^{(1)}_{pm}(q_0, u, n) S_{pm}(q_0, v, n) S_{pm}(q_0, \varphi, n); \ u > u_0 \]

\[ H^i_v = \frac{1}{j \omega \mu \hbar} \left( -\frac{\partial}{\partial u} E_z \right) = \frac{-1}{j \omega \mu \hbar} \frac{\partial}{\partial u} E_z = \frac{1}{h} F \cos h^2 u - \cos^2 v \]

\[ H^t_v = \frac{j}{\omega \mu \hbar} \sqrt{8\pi} \sum_{n} j^{-n} \frac{J_{pm}(q_0, u, n)}{N_{pm}(q_0, 0)} S_{pm}(q_0, v, n) S_{pm}(q_0, \varphi, n) \]

\[ H^s_v = \frac{j}{\omega \mu \hbar} \sqrt{8\pi} \sum_{n} j^{-n} a_{pm} H^{(1)}_{pm}(q_0, u, n) S_{pm}(q_0, v, n) S_{pm}(q_0, \varphi, n) \]

\[ E^t_z = \sqrt{8\pi} \sum_{n} j^{-n} \left[ b_{pm} J_{pm}(q_1, u, n) + c_{pm} Y_{pm}(q_1, u, n) \right] S_{pm}(q_1, v, n) S_{pm}(q_0, \varphi, n); \ u_1 < u < u_0 \]

\[ H^t_v = \frac{j}{\omega \mu \hbar} \sqrt{8\pi} \sum_{n} j^{-n} \left[ b_{pm} J_{pm}(q_1, u, n) + c_{pm} Y^\prime_{pm}(q_1, u, n) \right] S_{pm}(q_1, v, n) S_{pm}(q_0, \varphi, n) \]
By applying boundary conditions:

1. \[ E_{t_{lu=u_1}} = 0 \Rightarrow b_{pm} J_{pm}(q_1, u_1, n) + c_{pm} Y_{pm}(q_1, u_1, n) = 0 \]

2. \[ E^i + E^s_{lu=u_0} = E^t_{lu=u_0} \rightarrow \text{Sheet Impedance Boundary Condition} \]
   \[
   \sqrt{8\pi} \sum_n j^{-n} \left[ \frac{J_{pm}(q_0, u_0, n)}{N_{pm}(q_0, n)} + a_{pm} H^{(1)}_{pm}(q_0, u_0, n) \right] S_{pm}(q_0, v, n) S_{pm}(q_0, \varphi, n)
   \]
   \[
   = \sqrt{8\pi} \sum_n j^{-n} \left[ b_{pm} J_{pm}(q_1, u_0, n) + c_{pm} Y_{pm}(q_1, u_0, n) \right] S_{pm}(q_1, v, n) S_{pm}(q_0, \varphi, n)
   \]

3. \[ Z_s \left[ H^i + H^s_{lu=u_0} - H^t_{lu=u_0} \right] = E^t_{lu=u_0} \rightarrow \text{Sheet Impedance Boundary Condition} \]
   \[
   Z_s \left\{ \frac{j\sqrt{8\pi}}{\omega \mu h_0} \sum_n j^{-n} \left[ J'_{pm}(q_0, u_0, n) \right] \frac{N_{pm}(q_0, n)}{N_{pm}(q_0, n)} + a_{pm} H^{(1)'}_{pm}(q_0, u_0, n) \right\} S_{pm}(q_0, v, n) S_{pm}(q_0, \varphi, n) \]
   \[
   - \frac{j\sqrt{8\pi}}{\omega \mu h_0} \sum_n j^{-n} \left[ b_{pm} J'_{pm}(q_1, u_0, n) + c_{pm} Y'_{pm}(q_1, u_0, n) \right] S_{pm}(q_1, v, n) S_{pm}(q_0, \varphi, n) \]
   \[
   = \sqrt{8\pi} \sum_n j^{-n} \left[ b_{pm} J_{pm}(q_1, u_0, n) + c_{pm} Y_{pm}(q_1, u_0, n) \right] S_{pm}(q_1, v, n) S_{pm}(q_0, \varphi, n)
   \]
Matrix Equation for Coefficients

After some manipulation we come to the matrix equation below:

\[
\begin{bmatrix}
0 & J_{pm}(q_1, u_1, n) & Y_{pm}(q_1, u_1, n) \\
H_{pm}(q_0, u_0, n) & -\frac{\gamma_{pm}}{N_{pm}} J_{pm}(q_1, u_0, n) & -\frac{\gamma_{pm}}{N_{pm}} Y_{pm}(q_1, u_0, n) \\
(1) & \gamma_{pm}(J'_{pm}(q_1, u_0, n)) + \frac{\gamma_{pm}}{N_{pm}} \left( \frac{j_0 \mu F}{z_e} J_{pm}(q_1, u_0, n) \right) & \gamma_{pm}(Y'_{pm}(q_1, u_0, n)) + \frac{\gamma_{pm}}{N_{pm}} \left( \frac{j_0 \mu F}{z_e} Y_{pm}(q_1, u_0, n) \right)
\end{bmatrix}
\begin{bmatrix}
a_{pm} \\
b_{pm} \\
c_{pm}
\end{bmatrix}
= 0
\]

To find the scattered field for farfield region, we use the asymptotic form of the radial Mathieu function and we have:

\[
\vec{E}^s = \frac{\sqrt{8\pi}}{\sqrt{K \rho}} e^{j(K \rho - \frac{\pi}{4})} \sum_n j^{-2n} a_{pm} S_{pm}(q_0, v, n) S_{pm}(q_0, \varphi, n) \hat{z}
\]

Finally, the two-dimensional bistatic cross section is:

\[
\sigma_{2D} = \lim_{\rho \to \infty} 2\pi \rho \frac{|E^s|^2}{|E^i|^2}
\]

\[
\sigma_{2D} = \left[ \sum_n \sqrt{8\pi} j^{-2n} a_{pm} S_{pm}(q_0, v, n) S_{pm}(q_0, \varphi, n) \right]^2
\]
In the quasi-static limit \( q_0 = \frac{K_0^2 F^2}{4} \ll 1 \), \( q_1 = q_0 \varepsilon_r \ll 1 \), the closed-form condition for a PEC elliptical cylinder under TM-polarized illumination can be derived as:

\[
Z_{s-opt} = j \omega \mu F \cosh u_0 \frac{u_0 - u_1}{1 + \sinh 2u_0 \left( q_0 (u_0 - u_1) + q_1 \left( u_1 + \frac{1}{2} \ln q_1 \right) \right)}
\]

And also, the closed-form condition for a dielectric elliptical cylinder under TM-polarized illumination can be derived as:

\[
Z_{s-opt} = j \omega \mu F \frac{1 - q_1 \sinh^2 u_0}{2 \sinh u_0 (q_0 - q_1)}
\]
Frequency dispersion of the surface reactance for graphene monolayer and nanopatches with respect to the optimum required is found as:

\[ D = 5.064 \, \mu m, \, g = 0.524 \, \mu m, \, \mu_c = 0.2672 \, eV \]

\[ \mu_c = 0.5718 \, eV \]
Dielectric Elliptical Cylinder at THz Frequencies

- Geometry parameters are:
  \[ \varepsilon_r = 4 \text{ (silicon dioxide)} \]
  \[ a_0 = 12.5 \, \mu\text{m} \left( \frac{\lambda_0}{8} \right) \]
  \[ b_0 = 10 \, \mu\text{m} \left( \frac{\lambda_0}{10} \right) \]

- The required reactance is found to be: \[ X_s = 260 \, \Omega \]

- The design parameters are: \[ \mu_c = 0.6158 \, \text{eV}, \ T = 300 \, \text{K}, \ \tau = 1.5 \, \text{ps} \]
Cluster of Dielectric Elliptical Cylinders

Uncloaked

Cloaked

$g = 3 \mu m$

$f = 3 \text{THz}$
Overlapping of Dielectric Elliptical Cylinders

Uncloaked

Cloaked

Uncloaked

Cloaked

$l = 140 \mu m = 1.4 \lambda$

$f = 3 \text{ THz}$
Dielectric Elliptical Cylinder at Microwave Frequencies

- Geometry parameters are:
  \[ \varepsilon_r = 10 \]
  \[ a_0 = 10 \text{ mm } (\lambda_0/10) \]
  \[ b_0 = 8.6655 \text{ mm } (\lambda_0/11.54) \]

- The required reactance is found to be: \[ X_s = 70.75 \ \Omega \]

- The design parameters are:
  \[ D = 7.339 \text{ mm } (\lambda_0/13.62) \]
  \[ w = 0.3626 \text{ mm } (\lambda_0/275.786) \]
PEC Elliptical Cylinder at THz Frequencies

- Geometry parameters are:
  - $a_1 = 10.04 \, \mu m \, (\lambda_0 / 9.96)$
  - $b_1 = 6.67 \, \mu m \, (\lambda_0 / 15)$
  - $\varepsilon_r = 4$ (silicon dioxide)
  - $a_0 = 12.5 \, \mu m \, (\lambda_0 / 8)$
  - $b_0 = 10 \, \mu m \, (\lambda_0 / 10)$

- The required reactance is found to be: $X_S = -94.73 \, \Omega$

- The design parameters are:
  - $D = 5.064 \, \mu m$, $g = 0.524 \, \mu m$, $\mu_c = 0.5718 \, eV$
PEC Elliptical Cylinder at Microwave Frequencies

- Geometry parameters are:
  \[a_1 = 10 \text{ mm} \left( \frac{\lambda_0}{10} \right), \quad b_1 = 8.865 \text{ mm} \left( \frac{\lambda_0}{11.54} \right)\]
  \[a_0 = 11.9164 \text{ mm} \left( \frac{\lambda_0}{8.39} \right), \quad b_0 = 10.8167 \text{ mm} \left( \frac{\lambda_0}{9.24} \right)\]

- The required reactance is found to be: \[X_s = -66.12 \Omega\]

- The design parameters are:
  \[D = 7.145 \text{ mm} \left( \frac{\lambda_0}{14} \right), \quad g = 0.99 \text{ mm} \left( \frac{\lambda_0}{101.01} \right)\]
2-D Metasurface Cloak for PEC at Microwaves

- Geometry parameters are:
  
  \[ a_1 = 10 \text{ mm} \left( \frac{\lambda_0}{10} \right), \quad b_1 = 8.865 \text{ mm} \left( \frac{\lambda_0}{11.54} \right) \]
  
  \[ a_0 = 11.9164 \text{ mm} \left( \frac{\lambda_0}{8.39} \right), \quad b_0 = 10.8167 \text{ mm} \left( \frac{\lambda_0}{9.24} \right) \]

- The required reactance is found to be: \( X_s = -66.12 \ \Omega \)

- The design parameters are:
  
  \[ N = 10 \quad \quad D = 7.145 \text{ mm} \left( \frac{\lambda_0}{14} \right), \quad g = 0.99 \text{ mm} \left( \frac{\lambda_0}{101.01} \right) \]
Electric Field Distribution

\[ a_1 = 10 \text{ mm (} \lambda_0 / 10 \text{)} \]
\[ b_1 = 8.865 \text{ mm (} \lambda_0 / 11.54 \text{)} \]
\[ a_0 = 11.9164 \text{ mm (} \lambda_0 / 8.39 \text{)} \]
\[ b_0 = 10.8167 \text{ mm (} \lambda_0 / 9.24 \text{)} \]

\( f = 3 \text{ GHz} \)
Strip (Degenerated Ellipse)

- A strip can be modeled as a degenerated ellipse
- Geometry parameters are:
  - $a_0 = 10.04 \, \mu m \left( \lambda_0 / 9.96 \right)$, $b_0 = 6.67 \, \mu m \left( \lambda_0 / 15 \right)$
  - $a_1 = 7.5 \, \mu m = F \left( \lambda_0 / 13.33 \right)$, $\varepsilon_r = 4$
  - $X_s = -210 \, \Omega$
  - $D = 5.29 \, \mu m$, $g = 0.6 \, \mu m$, $\mu_c = 0.9 \, eV$

![Graphs showing scattering width vs. frequency for cloaked and uncloaked cases.](image)
Electric Field Distribution
Two Strips with Overlapping Cloaks

Uncloaked

Cloaked

$f = 3 \text{ THz}$

$g = 3.7 \text{ µm}$

Uncloaked

Cloaked

$f = 3 \text{ THz}$

$l = \frac{\lambda}{3.33}$
Cloaking 2-D Metallic Strip at Microwaves

- A 2-D Metallic Strip can be considered as a degenerated ellipse.
- TM-polarized plane-wave excitation.

\[
Z_{S_{TM,Hstrips}} = -j\eta_0 c \pi \frac{1}{\omega (\varepsilon_r + 1)D \ln \csc \left( \frac{\pi g}{2D} \right)}
\]

- \( f_0 = 3 \text{ GHz} \)
- \( a_0 = 8.457 \text{ mm} (\lambda_0/11.82) \)
- \( b_0 = 3.908 \text{ mm} (\lambda_0/25.58) \)
- \( \varepsilon_c = 10 \)
- \( Z_s = -j85.15 \text{ } \Omega \)
- \( D = 8.93 \text{ mm} \)
- \( g = 0.6 \text{ mm} \)
Cloaking 2-D Metallic Strip

- $f_0 = 3$ GHz
- $a_0 = 8.457$ mm ($\lambda_0/11.82$)
- $b_0 = 3.908$ mm ($\lambda_0/25.58$)
- $\varepsilon_c = 10$
- $Z_s = -j85.15$ $\Omega$
- $D = 8.93$ mm
- $g = 0.6$ mm

\[ \phi = 90^\circ \]
Wire Dipole Antennas

Overcoming Mutual Blockage Between Neighboring Dipole Antennas Using a Low-Profile Patterned Metasurface

Alessio Monti, Student Member; IEEE, Jason Soric, Student Member; IEEE, Andrea Alì, Member; IEEE, Filiberto Bilotti, Senior Member; IEEE, Alessandro Toscano, Senior Member; IEEE, and Lucio Vegni, Life Member; IEEE

Abstract—In this letter, we investigate the possibility of using the mantle cloaking approach to reduce mutual blockage effects between two electrically close antennas. In particular, we consider the case of two dipoles resonating at different, close frequencies and separated by an electrically short distance ($\lambda_0/10$ at 3 GHz). We show that by covering the two antennas with properly patterned metasurfaces printed on realistic substrates, it is possible to make each antenna invisible to the other and preserve their individual operation as if they were isolated. This new cloaking application is confirmed by realistic full-wave numerical simulations.

Index Terms—Cloaking, dipole antennas, metasurfaces.

I. INTRODUCTION

In the last decade, there has been a worldwide effort in the design of electromagnetic covers that may strongly reduce the visibility, scattering signature, and electromagnetic in-
Here, we present the applicability of elliptically shaped metasurfaces in order to reduce the mutual coupling between two closely spaced antennas.

First, we consider two strip dipole antennas resonating at $f = 1$ GHz and $f = 5$ GHz, which are separated by a short distance of $d = \lambda/10$ (at $f = 5$ GHz). (Case I)

Second, we consider two strip dipole antennas resonating at $f = 3.02$ GHz and $f = 3.33$ GHz, which are separated by a short distance of $d = \lambda/10$ (at $f = 3$ GHz). (Case II)

To present how the mutual blockage is overcome, we consider three different scenarios of isolated, uncloaked, and cloaked for each case.
We consider Antenna I (isolated) and Antenna II (isolated) which resonate at $f= 1$ GHz and $f= 5$ GHz, respectively, with omni-directional radiation patterns as shown below. Each antenna is matched to a 75-Ω feed.
Now, the antennas are placed in close proximity to each other. The presence of Antenna II does not have much effect on Antenna I since its length is small compared to the wavelength of the resonance frequency of Antenna I, but Antenna I changes the matching characteristics and radiation pattern of Antenna I drastically.

$f = 1$ GHz

$d = 6$ mm
How to Cloak Antenna I?

Since the length of Antenna I is 2.5 times the wavelength of Antenna II, therefore, we propose to use the analytical approach for infinite length as a good approximation to find the required metasurface for this case.

- \( Z_s = -j28 \ \Omega \)
- \( a_0 = 2.2 \ \text{mm} \ (\lambda_0/27.27) \)
- \( b_0 = 0.9165 \ \text{mm} \ (\lambda_0/65.46) \)
- \( \varepsilon_r = 25 \)
- \( D = 6.515 \ \text{mm} \)
- \( g = 1.29 \ \text{mm} \)

\[ L_1 \approx 2.5 \lambda_2 \]
Neighboring Cloaked Dipole Antennas

- 3-D radiation patterns of Antenna I at 1 GHz (left) and Antenna II at 5 GHz (right) for the scenario, in which Antenna I is cloaked for the resonance frequency of Antenna II and the antennas are in close proximity.
Restoration of gain patterns at the first and second resonance frequency of Antenna I (1 GHz, 3 GHz) and resonance frequency of Antenna II

- S-Parameters (dB) vs Frequency (GHz)
- E-plane and H-plane diagrams for Antenna I and Antenna II in different states (Isolated, Uncloaked, Cloaked for Antenna II)
First of all, we consider the antenna I (Isolated Case), which resonates at \( f = 3.02 \text{ GHz} \) with an omni-directional radiation pattern as shown below. The \( S_{11} \) of the antenna along with its dimensions are shown below. The antenna is matched to a 75-\( \Omega \) feed.
Then, we consider the antenna II (Isolated Case), which resonates at $f = 3.33$ GHz with an omni-directional radiation pattern as illustrated below. The $S_{11}$ of the antenna along with its dimensions are shown below. The antenna is matched to a 75-$\Omega$ feed.
Now, the antennas are placed in close proximity to each other. As expected, the presence of each of the antennas affects the radiation pattern of the other one drastically because the near-field distribution is changed, and therefore, the input reactance is changed remarkably.
In this slide, the antenna I covered with the spacer and the metasurface is presented. To achieve a good matching at the desired resonance frequency of 3.02 GHz, we reduced the length of the antenna from $L = 45.8$ mm to $L = 41.4$ mm. The $S_{11}$ parameter, permittivity of the spacer, and also, the dimensions of the cloak structure are shown below.

$$D = 3.4034 \text{ mm}$$
$$w = 0.35 \text{ mm}$$
$$a = 2.2 \text{ mm}$$
$$b = 0.9165 \text{ mm}$$
$$\varepsilon_r = 6.15$$

$L = 41.4$ mm

Rogers RO3006 (Lossy)
In this slide, the antenna II covered with the spacer and the cloak design is presented. To achieve a good matching at the desired resonance frequency of 3.02 GHz, again, we reduced the length of the antenna from $L = 41.5$ mm to $L = 38.8$ mm. The $S_{11}$ parameter, permittivity of the spacer, and also, the dimensions of the cloak structure are shown below.

$D = 3.4034$ mm  
$w = 0.3$ mm  
$a = 2.2$ mm  
$b = 0.9165$ mm  
$\varepsilon_r = 9.8$

$L = 38.8$ mm

RCS

Rogers TMM 10i (Lossy)
Neighboring Cloaked Dipole Antennas

- The reflection coefficients at the input port of the Antenna I and the Antenna II in the cloaked case (the antennas are in close proximity to each other) are shown here. As can be seen, the impedance matching of the antennas are good near the resonant frequency of each isolated strip dipole antenna.

- Radiation patterns are:
Neighboring Cloaked Dipole Antennas

- **E-Plane**
  - $f = 2.9441 \text{ GHz}$
  - $f = 3.3515 \text{ GHz}$

- **H-Plane**
  - $f = 2.9441 \text{ GHz}$
  - $f = 3.3515 \text{ GHz}$
First of all, we consider the Antenna I (Isolated) and the Antenna II (Isolated), which resonate at $f = 3.02$ GHz and $f = 3.33$ GHz, respectively, with omni-directional radiation patterns as shown below. The antennas are matched to a $37.5\,\Omega$ feed.
Now, the antennas are placed in close proximity to each other (d = 10 mm). As expected, the presence of each of the antennas affects the radiation pattern of the other one drastically because the near-field distribution is changed, and therefore, the input reactance is changed remarkably.
The reflection coefficients at the input port of the Antenna I and the Antenna II in the cloaked case (the antennas are in close proximity to each other) are shown here. As can be seen, the impedance matching of the antennas are good near the resonance frequency of each isolated strip monopole antenna.
Design Parameters for Monopole Antennas

Antenna I:
- \( L_1 = 20.7 \text{ mm} \)
- \( \Delta = 0.1 \text{ mm} \)
- \( W_1 = 4 \text{ mm} \)

Antenna II:
- \( L_2 = 19.4 \text{ mm} \)
- \( \Delta = 0.1 \text{ mm} \)
- \( W_2 = 4 \text{ mm} \)

\( \varepsilon_r = 6.15 \) for Antenna I
\( \varepsilon_r = 9.8 \) for Antenna II
Electric Field Distribution

Antenna I

<table>
<thead>
<tr>
<th>Isolated</th>
<th>Uncloaked</th>
<th>Cloaked</th>
</tr>
</thead>
</table>

Antenna II

<table>
<thead>
<tr>
<th>Isolated</th>
<th>Uncloaked</th>
<th>Cloaked</th>
</tr>
</thead>
</table>
Future Work

- Extension of the idea to different planar antennas!!!
Conclusions

- In this presentation, we proposed the idea of utilizing elliptically shaped cloak metasurfaces (which are used to cloak metallic strips) to reduce the mutual coupling between neighboring strip monopole and dipole antennas.

- In Case I, the strip monopole/dipole antennas have been designed to resonate at $f = 1$ GHz and $f = 5$ GHz and separated by a short distance of $d = \lambda/10$ (at $f = 5$ GHz).

- In Case II, the strip monopole/dipole antennas have been designed to resonate at $f = 3.02$ GHz and $f = 3.33$ GHz and separated by a short distance of $d = \lambda/10$ (at $f = 3$ GHz).

- Each antenna’s radiation properties are restored in a way that the radiation patterns are nearly similar to the isolated case.