A STUDY OF ELECTROMAGNETIC ABSORBERS AND CLOAKS
FOR THE REDUCTION OF ELECTROMAGNETIC SCATTERING

A Dissertation in
Electrical Engineering
by
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ABSTRACT

Electromagnetic absorbers and scattering reduction techniques have long been investigated to discover better performing configurations and exploited to reduce Radar Cross-Section, act as sensors or reduce obstruction effects, throughout the electromagnetic spectrum ranging from UHF to terahertz frequencies, and even at infrared and optical wavelengths.

This dissertation presents the research on a novel interpretation and design strategy for designing absorbers based on periodic structures and introduces an algorithm for determining the optimal material parameter for layered absorbers that are wrapped around real-world objects with structural perturbations from a planar surface, which traditional research focuses on almost exclusively. A brief history of absorbers was given and legacy configurations of absorbers were introduced in the first place. Secondly, novel Frequency Selective Surface (FSS)-based absorbers were proposed based on the interpretation of the reciprocity theorem for antenna systems. FSS-based absorbers and were incorporate into layered absorbers as composites for tailored absorption specifications. A comparison of performances was given to serve as a general rule of thumb to select optimal configuration for tailored specifications. This dissertation investigates a nascent solution to the scattering reduction problem, namely cloaking based on the physics of Transformation Optics (TO) and presents the real-world limitations of such solutions. This dissertation proposes an alternative algorithm for developing the optimal material parameter for a physical object in a real-world scenario.

These explorations show the great promise and applicability of a comprehensive tailored absorber design strategy on a case-by-case basis.
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and unconditional support I sometimes don't deserve.
Chapter 1
Overview

1.1 Dissertation Outline

This dissertation is divided into seven chapters including the introductory and the concluding ones.

In Chapter 2, legacy electromagnetic absorber designs are reviewed. Limitations of absorption performance for these legacy designs are discussed individually, which serve as motivations for this dissertation.

In Chapter 3, a novel approach to designing FSS type absorbers inspired by the reciprocity theorem is presented. Several examples are demonstrated based on the approach.

In Chapter 4, an optimization-based approach to designing multi-layer absorbers is presented. FSS type absorbers discussed in Chapter 3 are incorporated into the multi-layer absorber and the advantages of the composite structures are shown.

In Chapter 5, the principles of the Transformation Optics (TO) approach to scattering reduction are reviewed and limitations are explained. An alternative approach of Field Transformation (FT) is presented and the Generalized Scattering Matrix (GSM) interpretation of the scattering reduction problem is introduced.

In Chapter 6, an algorithm for developing the material parameters and thicknesses of an absorbing coating for an arbitrarily shaped target object is presented. Examples are given to demonstrate the technique.

Finally, Chapter 7 proposes potential directions for future work.
1.2 Notations and Symbols

The definitions and explanations of the notations and symbols used in this dissertation are included as follows:

$E =$ electric field intensity,

$H =$ magnetic field intensity,

$\eta =$ intrinsic impedance of the medium

$J =$ electric current density

$M =$ magnetic current density

$\sigma =$ conductivity of the medium

$\lambda =$ wavelength of the electromagnetic wave

$\Gamma =$ reflection coefficient at a certain interface of a system

$Z =$ intrinsic impedance of a medium, or seen from an interface

$Y =$ intrinsic admittance of a medium, or seen from an interface

$P =$ electromagnetic power

$R =$ resistance of a component or the entire system

$k =$ wave number of the electromagnetic wave

$\varepsilon =$ permittivity of the medium

$\mu =$ permeability of the medium
Chapter 2

Traditional Electromagnetic Absorbers

2.1 Introduction

This chapter follows the footsteps of the earliest attempts and major advances in absorber design. Three legacy absorbers are introduced in this chapter and their performances are evaluated analytically and numerically.

The absorption characteristics of any configuration can be evaluated by examining the normalized reflected power level, where we have defined

\[ Absorption = 1 - Reflection - Transmission \]

(2.1)

For configurations in which the absorbing screen is backed by a PEC ground plane, which blocks all transmission, we have

\[ Absorption = 1 - Reflection \]

(2.2)

Since the normalized reflected power level is most straightforward to derive analytically and examined in a measuring facility, reflection coefficients are commonly used to indicate the performance of the target absorber.
2.2 Salisbury Screen Absorber

Being one of the oldest and simplest designs for “absorbent bodies” [1], the Salisbury screen is a resonant absorber created by placing a resistive sheet on a low dielectric constant spacer in front of a metal plate.

![Diagram of Salisbury Screen Absorber](image)

Figure 2.1. Schematics for the Salisbury screen.

Figure 2.1 illustrates the geometry of the Salisbury absorber, in which an infinitesimally thin resistive sheet of conductivity $\sigma$ is placed at a distance $d$ from a metallic ground plane. Typically, a honeycomb or plastic foam is used as a dielectric spacer, so a relative dielectric constant of 1.03–1.1 is expected [2]. It’s obvious from Figure 2.2 that the electric and magnetic field in any of the two medium can be written as:

$$E_{0,1} = A_{0,1} e^{-j k_{0,1}z} + B_{0,1} e^{+j k_{0,1}z}$$  \hfill (2.3a)

$$H_{0,1} = \frac{1}{\eta_{0,1}} \left( A_{0,1} e^{-j k_{0,1}z} + B_{0,1} e^{+j k_{0,1}z} \right)$$  \hfill (2.3b)
where $\eta$ represents the intrinsic impedance of the medium, whereas $A$ and $B$ represent the amplitudes of the forward and backward propagating waves in the medium, respectively (Fig. 2.2).

To simplify this analysis, the normalized permittivity of the spacer is assumed to be that of free space (i.e., $k_0 = k_1, \eta_0 = \eta_1$). The boundary conditions at the interface are:

$$J = \sigma \cdot E_1 = \sigma \cdot E_2$$  \hspace{1cm} (2.4)

$$H_0 - H_1 = J$$  \hspace{1cm} (2.5)

which yields:

$$B_0 = -\frac{e^{-j\kappa_0 d}}{2} \left[ (\eta_0 \cdot \sigma) \cdot e^{-j\kappa_0 d} + (2 - \eta_0 \cdot \sigma) \cdot e^{+j\kappa_0 d} \right]$$  \hspace{1cm} (2.6)

For the screen to achieve maximum absorption, the backward propagating wave needs to be diminished. This requires that the magnitudes of the two exponentials in the brackets be equal in amplitude and that their phase angles be opposite. The equal amplitude requirement forces $\sigma$ to
equal to 1, or equivalently, the characteristic resistance to be $377 \Omega/sq$. In that case, (2.6) becomes:

$$B_0 = e^{-j \kappa_0 d} \cdot \cos \left( \frac{2 \pi d}{\lambda_0} \right) = 0$$

which implies

$$d = \frac{\lambda_0}{4} + \frac{\lambda_0}{2} n, \quad n = 0, 1, 2, \ldots$$

The performance for a Salisbury screen with a 7.5mm spacing is shown in Figure 2.3 for various values of sheet resistances. Note that the reflection coefficient reaches its minimum value roughly at 10 GHz ($\lambda = 30 \text{ mm}$). The best performance is achieved with a resistivity of $377 \Omega/sq$, but the performance is still a respectable -18.6 dB for a lower resistivity of $300 \Omega/sq$.

Figure 2.3. Reflection coefficients for designed Salisbury screen with different sheet resistances.
Sometimes the spacer can be replaced with dielectrics that have slightly higher values of permittivity to have stronger and sturdier structures. A more general form of the reflection coefficient with any dielectric spacer takes the form:

\[
B_0 = -\frac{e^{-jk\sigma d}}{2} \left[ \left( \frac{\eta_0}{\eta_1} - 1 + \eta_0 \cdot \sigma \right) \cdot e^{-j\kappa_1 d} + \left( \frac{\eta_0}{\eta_1} + 1 - \eta_0 \cdot \sigma \right) \cdot e^{+j\kappa_1 d} \right]
\]  

(2.9)

Similarly, the two exponentials should have equal amplitude and opposite in phase for the maximum level of absorption. This yields

\[
\frac{\eta_0}{\eta_1} - 1 + \eta_0 \cdot \sigma = \frac{\eta_0}{\eta_1} + 1 - \eta_0 \cdot \sigma
\]

(2.10)

leading to the same sheet resistance for the sheet regardless of the spacer material. Consequently, (2.9) reduces to

\[
B_0 = -e^{-jk\sigma d} \cdot \frac{\eta_0}{\eta_1} \cdot \cos \left( \frac{2\pi d}{\lambda_1} \right) = 0
\]

(2.11)

which implies

\[
d = \frac{\lambda_1}{4} + \frac{\lambda_1}{2} n, \quad n = 0, 1, 2, \ldots
\]

(2.12)

We see from (2.12) that the thickness of the Salisbury screen can be reduced by introducing denser dielectric spacers because the minimum thickness of the spacer corresponds to a quarter of the wavelength of the spacer material. Figure 2.4 shows the reflection coefficients of three screens using different lossless dielectric spacers with a designed operating frequency at 10GHz. The reduction in thickness could be beneficial for certain practical applications. However, such reduction in thickness is accompanied by a reduction in absorption bandwidth, which is modest at best.
Figure 2.4. Reflection coefficients for designed Salisbury screen with spacers of different dielectric constants and corresponding thicknesses.

2.3 Dallenbach Layer Absorber

Another simple absorber of the resonant nature, the Dallenbach layer, comprises of a homogeneous lossy layer backed by a metallic plate, as shown in Figure 2.5 [3]. In a steady-state (as opposed to a transient) analysis, the reflection at the surface of a material is due to the impedance change seen by the wave at the interface between the two media. Therefore, if a material can be found whose impedance relative to free space equals 1 (i.e., $\mu_r = \varepsilon_r$), there will be no reflection at the interface. In this case the total absorption will depend on the loss properties of the material (dispersion profile of permittivity and permeability) and the electrical thickness.
However, materials with appropriate dielectric and magnetic properties to act as a *perfectly* matched RAM over any appreciable frequency range are difficult to find at best. So the question becomes one of optimizing the loss at a given frequency using available materials. For a single material layer backed by a conducting plate, the reflection coefficient is given by

$$\Gamma = \frac{Z_{in} - Z_0}{Z_{in} + Z_0}$$

(2.13)

where $Z_0$ represents the intrinsic impedance of free space and $Z_{in}$ represents the impedance seen by the wave at the air-absorber interface, which can be expressed as

$$Z_{in} = Z_d \frac{Z_L + j Z_d \cdot \tan(k_d d)}{Z_d + j Z_L \cdot \tan(k_d d)}$$

(2.14)

where $Z_d$ represents the intrinsic impedance of the lossy layer and $Z_L$ is the intrinsic impedance of the metallic plane, which is essentially 0, yielding the following reflection coefficient:

$$\Gamma = \frac{j Z_d \cdot \tan(k_d d) - Z_0}{j Z_d \cdot \tan(k_d d) + Z_0} = \frac{j \sqrt{\frac{\mu_d}{\varepsilon_d}} \tan\left(\frac{2 \pi d}{\lambda_0} \sqrt{\mu_d \cdot \varepsilon_d}\right) - 1}{j \sqrt{\frac{\mu_d}{\varepsilon_d}} \tan\left(\frac{2 \pi d}{\lambda_0} \sqrt{\mu_d \cdot \varepsilon_d}\right) + 1}$$
It’s evident from (2.15) that given the material used to construct the Dallenbach layer absorber, \textit{i.e.}, with the knowledge of the permittivity and permeability profile, the absorption performance relies solely on the fraction of wavelength of the layer thickness.

One thing should be noted from the derived expression. The optimal choice of the thickness for the lossy layer does not necessarily follow the straightforward quarter-wavelength rule due to the potentially dispersive nature of the permittivity and permeability profile of the lossy layer, nor does an increase in the thickness of the lossy layer guarantees a shift towards lower frequencies for the absorption peak [4].

Figure 2.6 shows such an example of a Dallenbach layer absorber with different thicknesses using materials whose electromagnetic properties are depicted in Figure 2.7.

Figure 2.6. Microwave absorption properties of the FePc–Fe₃O₄–BF/BPh/FePc–Fe₃O₄ composite laminates.
In general, the reflection coefficient of a Dallenbach layer absorber can be easily calculated either analytically by using (2.15) with full knowledge of the material profile, or by measuring in an anechoic chamber for layers of various thicknesses. This has been the most direct as well as important method available to material engineers for evaluating the performance of such absorbers [5-7].

2.4 Jaumann Absorber

One of the most critical characteristics for absorbers is the bandwidth performance. A traditional Salisbury screen with a center frequency of 10GHz will have decent absorption across the entire
X-band (Figure 2.3), but its performance will deteriorate for the remainder of the radar band—spanning from 2GHz to 18GHz and beyond even that for certain applications.

The bandwidth of a Salisbury screen can be improved by adding additional resistive sheets and spacers to form a Jaumann absorber [8]. Figure 2.8 shows an example of an eight-layer Jaumann absorber. The thicknesses and sheet resistances for each layer are shown in Table 2.1 and 2.2, respectively [9]. Only air spacers are used in this example for simplicity of the analysis.

Table 2-1. Thicknesses for all the layers of the Jaumann absorber (mm)

<table>
<thead>
<tr>
<th>$d_1$</th>
<th>$d_2$</th>
<th>$d_3$</th>
<th>$d_4$</th>
<th>$d_5$</th>
<th>$d_6$</th>
<th>$d_7$</th>
<th>$d_8$</th>
</tr>
</thead>
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<td>3.9</td>
<td>4</td>
<td>3.9</td>
<td>3.9</td>
<td>3.9</td>
<td>4</td>
<td>3.9</td>
</tr>
</tbody>
</table>

Table 2-2. Sheet resistances for all the layers of the Jaumann absorber ($\Omega$/sq)

<table>
<thead>
<tr>
<th>$R_{s1}$</th>
<th>$R_{s2}$</th>
<th>$R_{s3}$</th>
<th>$R_{s4}$</th>
<th>$R_{s5}$</th>
<th>$R_{s6}$</th>
<th>$R_{s7}$</th>
<th>$R_{s8}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>305</td>
<td>579</td>
<td>873</td>
<td>1266.5</td>
<td>1796</td>
<td>2480</td>
<td>3724</td>
<td>2067</td>
</tr>
</tbody>
</table>
Figure 2.8. Schematics for an 8-layer Jaumann absorber.

Figure 2.9 shows the simulated reflection coefficient for the eight-layer Jaumann absorber, with a 10dB absorption bandwidth of 2.5–36GHz and a 20dB absorption bandwidth of 2.95–35.2GHz. The total thickness for this Jaumann absorber is slightly larger than a quarter of the wavelength at the lowest frequency (26.8%). Alternative approaches need to be considered to further reduce the thickness of the Jaumann absorber while maintaining its ultra-wideband performance.

Figure 2.9. Reflection coefficient for an 8-layer Jaumann absorber.

2.5 Summary

In this chapter, we have briefly reviewed the very cornerstone of electromagnetic absorbers. The absorption performances for the traditional Salisbury screen, Dallenbach layer and
Jaumann absorber are evaluated and their advantages and disadvantages outlined. These legacy designs will facilitate as theoretical basis as well as crucial components for various applications, as will be demonstrated in the following chapters.
Chapter 3

Frequency Selective Surface (FSS)-based Absorber for Wideband Applications

3.1 Introduction

One major branch of absorber design in contemporary research is focused on FSS-based absorber, which comprise of an FSS screen printed above a dielectric substrate with a PEC backing. The unit cell of the FSS commonly utilizes resonant structures that correspond to the working frequency of the absorber. In common with the Salisbury screen, the FSS acts as an impedance transformer [10], which transforms the input impedance of the absorber at the air-FSS interface to that of the characteristic impedance of air, achieving perfect absorption at the design frequency. As a result, the absorption performance is extremely narrowband unless the unit cell is designed with multiple structures that correspond to resonant frequencies near the designed operating frequency, contributing to wideband absorption characteristics [11].

A different approach to designing absorbers is presented in this chapter, which was inspired by the reciprocity theorem. This approach is particularly useful when the bandwidth of the absorber is of primary design concern. A systematic design procedure can be followed to meet the desired specification in the operating frequency range.
3.2 General Reciprocity Theorem

The reciprocity theorem in electromagnetics is discussed in this section as a starting point. This applies to the use of Maxwell’s equations. Assuming that within a linear and isotropic medium, but not necessarily homogeneous, there exist two sets of sources \((J_1, M_1)\) and \((J_2, M_2)\) which radiate simultaneously inside the same medium and at the same frequency, producing fields \((E_1, H_1)\) and \((E_2, H_2)\) respectively. It can be shown \([12], [13]\) that the sources and fields satisfy

\[
-\oint_S (E_1 \times H_2 - E_2 \times H_1) \cdot ds' = \iiint_V (E_1 \cdot J_2 + H_2 \cdot M_1 - E_2 \cdot J_1 - H_1 \cdot M_2) \cdot dv'
\]

which is referred to in the literature as the Lorentz Reciprocity Theorem in integral form.

Another useful form of (3.1) is to consider that the fields \((E_1, H_1, E_2, H_2)\) and the sources \((J_1, M_1, J_2, M_2)\) are within a medium that is enclosed by a sphere of infinite radius. Assume that the sources are positioned within a finite region and that the fields are observed in the far field (ideally at infinity). Then the left side of (3.1) is equal to zero, or

\[
\oint_S (E_1 \times H_2 - E_2 \times H_1) \cdot ds' = 0
\]

under these circumstances,

\[
\iiint_V (E_1 \cdot J_2 + H_2 \cdot M_1 - E_2 \cdot J_1 - H_1 \cdot M_2) \cdot dv' = 0
\]

Equation (3.4) can also be written as

\[
\iiint_V (E_1 \cdot J_2 - H_1 \cdot M_2) \cdot dv' = \iiint_V (E_2 \cdot J_1 - H_2 \cdot M_1) \cdot dv'
\]

(3.4)
3.3 Reciprocity Theorem for Antenna Systems

There are many applications of the reciprocity theorem. To demonstrate its potential, an antenna example will be considered. Two antennas, whose input impedances are $Z_1$ and $Z_2$, are separated by a linear and isotropic (but not necessarily homogeneous) medium, as shown in Fig.3.1. One antenna (#1) is used as a transmitter and the other (#2) as a receiver. The equivalent network of each antenna is given in Fig.3.2. The internal impedance of the generator $Z_g$ is assumed to be the conjugate of the impedance of antenna #1 ($Z_g = Z_1^* = R_1 - jX_1$) while the load impedance $Z_L$ is equal to the conjugate of the impedance of antenna #2 ($Z_L = Z_2^* = R_2 - jX_2$). These assumptions are made only for convenience.

Figure.3.1. Transmitting and receiving antenna systems.

Figure.3.2. Two-antenna system with conjugate loads.
The power delivered by the generator to antenna #1 is given by (3.5)

\[ P_1 = \frac{1}{2} \text{Re}[V_1 I_1^*] = \frac{1}{2} \text{Re} \left[ \left( \frac{V_g Z_1}{Z_1 + Z_g} \right) \left( \frac{V_g^*}{Z_1 + Z_g} \right)^* \right] = \frac{|V_g|^2}{8R_1} \]

(3.5)

If the transfer admittance of the combined network consisting of the generator impedance, antennas, and load impedance is \( Y_{21} \), the current through the load is \( V_g Y_{21} \) and the power delivered to the load is

\[ P_2 = \frac{1}{2} \text{Re} \left[ Z_2 (V_g Y_{21}) (V_g Y_{21})^* \right] = \frac{1}{2} R_2 |V_g|^2 |Y_{21}|^2 \]

(3.6)

The ratio of (3.6) to (3.5) is

\[ \frac{P_2}{P_1} = 4R_1 R_2 |Y_{21}|^2 \]

(3.7)

Under conditions of reciprocity (\( Y_{12} = Y_{21} \)), the power delivered in either direction is the same.

Taking a cue from the reciprocity theorem, we specifically choose well-matched antenna structures as unit cells, as electromagnetic wave receivers and converters.

### 3.4 Unit–Cell Design and Bandwidth Expansion

We start with the most fundamental type of antenna as the component of the unit cell. A cross-dipole antenna is introduced to account for both polarizations and the feed is replaced by a resistor patch to connect both arms of the dipole [14].

An example of such a screen is the periodic cross-strip structure consisting of conducting arms (grey in color) and a load resistor (red in color), as shown in Fig.3.3. The cross-strip structure is considered as a unit cell and the reflection coefficient for the screen is simulated for
the case where it is backed by a PEC sheet. An increase of bandwidth performance is achieved by adding an additional layer of larger cross-strips with lower resonant frequencies. The reflection coefficient for the cross-strip structures are shown in Fig. 3.4.

Figure. 3.3. (a,d) Isometric; (b,e) top; and (c,f) side view of the cross-strip model with interconnected resistors (red) and duo layer stack model, respectively (D=15, W=12, H=5.5, H’=11.5 unit: mm).
3.5 Wideband Unit-Cells and Dual-Direction Capabilities

Next, to enhance the bandwidth of the screen we replace cross-strips with bowties, as shown in Fig.3.5. In this design, conducting bowties (grey in color) are connected with load resistors (red and pink in color). We have simulated a 3-layer configuration of this structure, and Fig.3.6 shows that a wide absorption bandwidth can be achieved even in the absence of a PEC backing, which provides both dual-directional absorption capabilities and desirable absorption levels for applications in which the backing is not metallic.
Figure 3.5. Illustration of the (a) isometric; (b) top; (c) side view of the 3-layer bowtie absorber model and the (d) 1st layer; (e) 2nd layer; (f) 3rd layer (L=53.4mm, W=21mm, d=6mm).

Figure 3.6. Reflection, transmission and absorption coefficient of the three layer absorber-screen model.
3.6 Lossy FSS Analysis

Typically, of FSS-based absorber are realized by using lossy FSSs, with varying sheet resistance, or unit cells comprising of with perfectly conducting strips with resistive inserts. We start by examining the latter design, to see how we might enhance its performance.

We start with an existing design which has been proposed by Shang et al. [15], comprising of a complex dual-loop absorber illustrated in Fig.3.7. The conducting strips forming the loop can be segmented into various lengths to introduce multiple resonances. These parts can then be connected together by resistive loads to lower the Q factor of the system. The multiple resonances of the system can be adjusted to achieve a wideband performance (as shown in Fig.3.8).
Figure 3.7. Schematics of the unit cell of an EM absorber based on dual-loop periodic screens, model (a).
In this section we compare the performances of three different configurations: (a) the original design in [15] which has a resistively-loaded dual-loop (PEC); (b) a modified design with a resistively-loaded dual-loop (lossy with a sheet resistance of 100Ω/sq) as shown in Fig.3.9; (c) a complete dual-loop (lossy with a sheet resistance of 100Ω/sq) as shown in Fig.3.10.

The loop sizes as well as the load resistances are chosen to be the same in all cases, and all of the three screens are backed by a PEC ground plane with a 14.7mm air spacer, which is identical to that in [15].
The absorption characteristics of the configurations in (a) and (b) are presented in Fig.3.11 for varying sheet resistances.
The peak absorption frequency for configuration (b) can be systematically tuned by adjusting the sheet resistance of the unsegmented dual-loop, but without changing the operational bandwidth. Such a feature could be useful when designing absorbers with band-suppression requirements [16].

Fig.3.12 compares the performances of the configurations (a) and (c) when their sheet resistance is varied. We observe that the unsegmented dual-loop configuration preserves the multiple resonance characteristic of the original segmented design when the sheet resistance is chosen appropriately. The peak absorption frequency can also be tailored by changing the sheet resistance of the dual-loop, though at some expense of the overall bandwidth.
Figure 3.12. Reflected power level for model (c) with varying sheet resistance.

Two different modifications of an FSS-based EM absorber comprising of resistively loaded dual-loops have been studied. It is shown that varying the sheet resistance of the periodic screen can achieve enhanced absorption at frequencies where a high absorption level is desired, with little compromise of its bandwidth performance.
Chapter 4

Multi-Layered Absorber for Ultra Wideband Applications

4.1 Design of Multilayered Absorber

Our objective is to design wideband layered absorbers for infinite, planar, conducting ground planes. We note that this is also the basic approach to designing coating for radar targets to render them stealthy, regardless of their shapes. Furthermore, it is relatively easy to carry out this design by using optimization algorithms to determine the material properties and the layer thicknesses to reduce the reflection from the coated PEC plane below -10dB level, and over a wide frequency band; say covering radar frequencies from 2 to 18 GHz, for example.

4.1.1 Reflection Coefficient for obliquely incident stratified isotropic dielectric layers with a PEC ground plane

We start by examining the analytical expression for the reflections coefficient for an obliquely incident plane wave upon a dielectric layer with a PEC back, illustrated in Fig.4.1.
Figure 4.1. (a) TE and (b) TM wave incident obliquely on a multi-layer dielectric with a PEC ground plane.

For the case of a TE plane wave incident on a single layer, the fields inside medium 1 are of the form:

\[ \mathbf{E}_1 = a_y (A_1^+ e^{-j(k_x x + k_{z1} z)} + A_1^- e^{-j(k_x x - k_{z1} z)}) \]

\[ \mathbf{H}_1 = (-\frac{k_{z1}}{k_1} a_x + \frac{k_x}{k_1} a_z) \frac{A_1^+}{\eta_1} e^{-j(k_x x + k_{z1} z)} + (\frac{k_{z1}}{k_1} a_x + \frac{k_x}{k_1} a_z) \frac{A_1^-}{\eta_1} e^{-j(k_x x - k_{z1} z)} \]  

(4.1)

The fields inside medium 2 are of the form:

\[ \mathbf{E}_2 = a_y (A_2^+ e^{-j(k_x x + k_{z2} z)} + A_2^- e^{-j(k_x x - k_{z2} z)}) \]

\[ \mathbf{H}_2 = (-\frac{k_{z2}}{k_2} a_x + \frac{k_x}{k_2} a_z) \frac{A_2^+}{\eta_2} e^{-j(k_x x + k_{z2} z)} + (\frac{k_{z2}}{k_2} a_x + \frac{k_x}{k_2} a_z) \frac{A_2^-}{\eta_2} e^{-j(k_x x - k_{z2} z)} \]  

(4.2)

The coefficients for the fields are calculated by applying the boundary conditions at different interfaces, which are found to be:
where \( A_1^+ \) is the amplitude of the incident electric field and \( \Gamma_{TE} \) is the effective reflection coefficient given by:

\[
\Gamma_{TE} = \frac{(\eta_2 k_{x1} k_2 - \eta_1 k_{xz} k_1) - (\eta_2 k_{x1} k_2 + \eta_1 k_{xz} k_1) e^{-2jkxz d}}{(\eta_2 k_{x1} k_2 + \eta_1 k_{xz} k_1) - (\eta_2 k_{x1} k_2 - \eta_1 k_{xz} k_1) e^{-2jkxz d}}
\]

(4.4)

Similarly, for the case of a TM plane wave incident on a single layer, the fields inside medium 1 are of the form:

\[
E_1 = \left(\frac{k_{x1}}{k_1} a_x - \frac{k_x}{k_1} a_z\right) A_1^+ e^{-j(k_{xx} + k_{xz})} + \left(\frac{k_{x1}}{k_1} a_x + \frac{k_x}{k_1} a_z\right) A_1^- e^{-j(k_{xx} - k_{xz})}
\]

\[
H_1 = a_y \left(\frac{A_1^+}{\eta_1} e^{-j(k_{xx} + k_{xz})} - \frac{A_1^-}{\eta_1} e^{-j(k_{xx} - k_{xz})}\right)
\]

(4.5)

The fields inside medium 2 are of the form:

\[
E_2 = \left(\frac{k_{x2}}{k_2} a_x - \frac{k_x}{k_2} a_z\right) A_2^+ e^{-j(k_{xx} + k_{xz})} + \left(\frac{k_{x2}}{k_2} a_x + \frac{k_x}{k_2} a_z\right) A_2^- e^{-j(k_{xx} - k_{xz})}
\]

\[
H_2 = a_y \left(\frac{A_2^+}{\eta_2} e^{-j(k_{xx} + k_{xz})} - \frac{A_2^-}{\eta_2} e^{-j(k_{xx} - k_{xz})}\right)
\]

(4.6)

The coefficients for the fields are calculated by applying the boundary conditions on different interfaces, which are found to be:

\[
A_1^- = A_1^+ \Gamma_{TM}
\]
where $A_2^+$ is the amplitude of the incident electric field and $\Gamma_{TM}$ is the effective reflection coefficient given by:

$$
\Gamma_{TM} = \frac{(\eta_2 k_{z2} k_1 - \eta_1 k_{z1} k_2) - (\eta_2 k_{z2} k_1 + \eta_1 k_{z1} k_2) e^{-2 j k_{z2} d}}{(\eta_2 k_{z2} k_1 + \eta_1 k_{z1} k_2) - (\eta_2 k_{z2} k_1 - \eta_1 k_{z1} k_2) e^{-2 j k_{z2} d}}
$$

(4.8)

In general, for a TE incident multi-layer absorber, the overall reflection coefficient at the air-absorber interface is given by:

$$
\Gamma_{TE} = \frac{(\eta_2 k_{z2} k_1 - \eta_1 k_{z1} k_2) - (\eta_2 k_{z2} k_1 + \eta_1 k_{z1} k_2) e^{-2 j k_{z2} d} \Gamma_{TE,2}}{(\eta_2 k_{z2} k_1 + \eta_1 k_{z1} k_2) - (\eta_2 k_{z2} k_1 - \eta_1 k_{z1} k_2) e^{-2 j k_{z2} d} \Gamma_{TE,2}}
$$

(4.9)

where for $i = 2, 3, 4...m$,

$$
\Gamma_{TE,i} = \frac{(\eta_{i+1} k_{z,i+1} k_{i+1} - \eta_i k_{z,i+1} k_i) - (\eta_{i+1} k_{z,i+1} k_{i+1} + \eta_i k_{z,i+1} k_i) e^{-2 j k_{z,i+1} d} \Gamma_{i+1}}{(\eta_{i+1} k_{z,i+1} k_{i+1} + \eta_i k_{z,i+1} k_i) - (\eta_{i+1} k_{z,i+1} k_{i+1} - \eta_i k_{z,i+1} k_i) e^{-2 j k_{z,i+1} d} \Gamma_{i+1}}
$$

(4.10)

Similarly, the reflection coefficient for a TM wave is given by:

$$
\Gamma_{TM} = \frac{(\eta_2 k_{z2} k_1 - \eta_1 k_{z1} k_2) - (\eta_2 k_{z2} k_1 + \eta_1 k_{z1} k_2) e^{-2 j k_{z2} d} \Gamma_{TM,2}}{(\eta_2 k_{z2} k_1 + \eta_1 k_{z1} k_2) - (\eta_2 k_{z2} k_1 - \eta_1 k_{z1} k_2) e^{-2 j k_{z2} d} \Gamma_{TM,2}}
$$

(4.11)

where for $i = 2, 3, 4...m$,

$$
\Gamma_{TM,i} = \frac{(\eta_{i+1} k_{z,i+1} k_{i+1} - \eta_i k_{z,i+1} k_i) - (\eta_{i+1} k_{z,i+1} k_{i+1} + \eta_i k_{z,i+1} k_i) e^{-2 j k_{z,i+1} d} \Gamma_{i+1}}{(\eta_{i+1} k_{z,i+1} k_{i+1} + \eta_i k_{z,i+1} k_i) - (\eta_{i+1} k_{z,i+1} k_{i+1} - \eta_i k_{z,i+1} k_i) e^{-2 j k_{z,i+1} d} \Gamma_{i+1}}
$$

(4.12)
4.1.2 Optimization Process for Maximizing Frequency Bandwidth and Minimizing Thickness

The single-layered Dallenbach absorber has obvious limitations in terms of bandwidth and thickness. It is common practice to use multiple layers of absorbing materials to meet bandwidth specifications and reduce the overall thickness as much as possible. Several optimization algorithms, such as the Genetic Algorithm (GA) and Particle Swarm Optimization (PSO), have been developed to search for an optimal arrangement of the different layers of absorbers so that a continuous wideband absorption performance can be achieved. This has been vitally important since most synthetic-type real-world absorbing materials have a high level of dispersion, making a purely analytical approach to finding the optimal arrangement for the thicknesses of the absorber layers an extremely challenging task, if not impossible.

A typical cost function to be minimized, which is suitable for optimizing the layer thicknesses while balancing the absorption performance and overall thickness can be defined as follows:

\[ F = m \cdot R_{TE/TM} + (1 - m) \cdot \sum_{i=1}^{k} d_i \]

(4.13)

where \(d_i\) represents the thickness for each layer of the absorber; \(R_L\) stands for the reflection level at the air-absorber interface and is a function of the material parameters and thicknesses \(d_i\) of the absorbing layers (Fig.4.2); and \(m\) is the weight of the reflection level in the optimization, signifying the importance of the reflection level over the collective thickness of the layered absorber. Note that \(m = 1\) corresponds to the case where the total thickness is predetermined, and only the reflection coefficient is minimized. Note that the parameters of the materials available to us are dispersive, and hence the optimization to determine the thicknesses should be carried out over the entire desired frequency band.
As is well known, the RAMs (radar absorbing materials) have been around for a very long time, some for many decades, dating back to when stealth aircrafts came into vogue in the sixties, although earliest theoretical and experimental work date back to 1940s. We realize that information on some of these RAM materials is not openly available because of their “classified” or “secret” nature, understandably so because they are used in military applications to design stealth aircrafts and missiles. Nonetheless, a plethora of information about similar absorbing materials is available in the open literature, including the details of their fabrication, which have been described in [2], for instance.

Here we will use two different types of materials namely CoFe Nano-Flakes (NF) and CoFe Nano-Particles (NP), whose frequency variations are shown in Fig.4.3. We point out that these materials can be realized with relative ease, as is evident from [17, 18], where the details of their fabrication can be found.
To illustrate the fact that we can indeed achieve wideband performance in terms of reflection reduction over a wide frequency band with relative small thicknesses of 2, 4, 6 and 7 layer absorbers we refer to Fig.4.4, in which a 10dB (or better) reduction in the reflection coefficient is presented. Although not shown here, the results for the reflection coefficient reduction are also satisfactory when either the polarization, or the incident angle is varied and this is also true when both are changed simultaneously.
4.2 RCS Reduction for Test Targets

The multi-layer absorber designed for the infinite PEC plane is applied to an arbitrarily-shaped object. Initially we consider an object with a smooth surface whose radius of curvature is moderate-to-large everywhere. Fig.4.5 shows the back-scattering RCS level of three PEC structures each treated with the 2-layer and 7-layer absorber developed in Sec.4.1.2 and compared with the original PEC configurations, where the incident wave propagates along -Z direction.
Figure 4.5. Back-scattering RCS of PEC objects: (a) plate, (b) pyramid, and (c) cylinder covered with 2 and 7-layer absorbers.
4.3 FSS and Multi-Layered Composite Absorber for Performance Enhancement

4.3.1 Target Band Absorption Enhancement

We insert the cross-strip structure below a 2-layer absorber (Fig.4.6) to enhance its absorption performance (Fig.4.7). We note that we can realize a reduction in the reflection coefficient in the frequency band of 10 to 18GHz.

![Figure 4.6](image)

Figure 4.6. Illustration of the (a) isometric and (b) side view of the absorber cross-strip composite model.

![Figure 4.7](image)

Figure 4.7. Reflection coefficient of the absorber-screen composite model.
4.3.2 Thickness Reduction

In addition to the above designs for the absorbers, we have also analyzed the effect of introducing other types of periodic structures inside the layered absorbers. As an example, Fig. 4.8 and 4.9 show a broadband antenna and a modified screen implementing the antenna structure embedded within the original 2-layer absorber to help reduce the reflection coefficient. The result for the composite structure is shown in Fig. 4.10.

Figure 4.8. (a) Illustration of the broadband antenna; (b) S11 of the broadband antenna.
The screen designs discussed can also be scaled appropriately for the desired frequency band to enhance the performance of the layered absorber in a certain specified band.

### 4.4 Summary

The multi-layer absorber without FSS type inclusions yields optimal absorption performance in terms of bandwidth coverage, which can be evaluated by taking the ratio of overall thickness of the absorber and the wavelength of the lowest absorption frequency. Comparable bandwidth performance can be obtained by using only periodic screens but an increase in the overall thickness is usually required.

FSS type absorber inclusions can greatly enhance the absorption level in a desired target frequency band with some design flexibility because the scalability of periodic structures. One major tradeoff of this approach is that original wideband behavior of the multi-layer is
compromised and reduced to the target band because the standalone bandwidth of the FSS type absorber is usually smaller than that of the multi-layer absorber.
Chapter 5
Cloaking and Scattering Reduction

5.1 Introduction

The objective of this chapter is twofold. First of these is to review the basic principles of the Transformation Optics (TO) approach, also known as Transformation Electromagnetics (TEM) algorithm, which has recently surfaced as one of the most innovative techniques for designing a wide variety of electromagnetic devices including cloaks to render objects invisible. Our second goal is to present an alternative approach for designing absorptive coatings for scattering reduction, which differs from that used to design TO-based cloaks, and avoids the problems of narrow bandwidth and sensitivity to polarization and incident angle associated with the TO-based cloaks that we realize by using Metamaterials (MTMs).

The TO algorithm for designing cloaks is unique in that it provides a systematic approach, which is very innovative, as well as elegant, and is markedly different from the techniques that have been employed heretofore prior to the advent of the TO, to design RCS-reducing absorbers, for example.

Figure 5.1. (a) Physical and (b) virtual domains used in the TO algorithm.
5.2 Fundamentals of Transformation Optics

The principle upon which the TO is based has been enunciated in a number of papers, dating back to almost fifty years ago, that have examined the behavior of Maxwell’s equations in a generalized curvilinear coordinate system. To explain the basic principle of the concept, let us consider two objects belonging to physical and virtual domains, and shown in Figs. 5.1 (a) and (b), respectively. The medium parameters surrounding these objects, namely \((\varepsilon_1, \mu_1)\) and \((\varepsilon_2, \mu_2)\), are also shown in Fig. 5.1.

A number of prominent authors, among them Pendry [19], Leonhart [20], Hao [21-23] and Werner [24], as well as several others [25-36], have presented the relationship between the medium parameters in the two domains (physical and virtual) when we transform the geometry of object #2 in the virtual domain into that of object #1 in the physical domain via coordinate transformation, under the physical constraint that the electric and magnetic fields in the two domains remain “invariant” to the transformation between the two systems. The relationship can be explicitly stated as:

\[
A = \begin{bmatrix}
\frac{\partial x_1}{\partial x_2} & \frac{\partial x_1}{\partial y_2} & \frac{\partial x_1}{\partial z_2} \\
\frac{\partial y_1}{\partial x_2} & \frac{\partial y_1}{\partial y_2} & \frac{\partial y_1}{\partial z_2} \\
\frac{\partial z_1}{\partial x_2} & \frac{\partial z_1}{\partial y_2} & \frac{\partial z_1}{\partial z_2}
\end{bmatrix}
\]

\[\text{(5.1a)}\]

\[
\bar{\varepsilon}_1 = \frac{\Lambda \cdot \bar{\varepsilon}_2 \cdot A^T}{det(A)}
\]

\[
\bar{\mu}_1 = \frac{\Lambda \cdot \bar{\mu}_2 \cdot A^T}{det(A)}
\]

\[\text{(5.1b)}\]
where $A$ represents the Jacobian matrix relating the two domains ($x, y$ and $z$ can be any arbitrary curvilinear coordinate), $\varepsilon$ and $\mu$ represent the permittivity and the permeability of the corresponding mediums, respectively. Eqn. (6.1) enables us to navigate between the two systems and relate their material parameters in a unique, systematic and rigorous way.

Pendry [19], Smith [36] and a number of other workers [20-35] have leveraged the fact that the medium parameters can be related via (6.1), in order to lay the foundations of the TO algorithm for designing cloaks, which render a target invisible when covered by using materials whose parameters are dictated by the TO. To explore the basic principles of the TO, we return to Fig.5.1 and define the following task for ourselves: Design the cloak (i.e., a cover or a coating) for the PEC object in Fig.5.1 (a) such that it is invisible to an arbitrary incident field that impinges upon it. Note that no restriction is being placed on the frequency, polarization or the angle of incidence of the illuminating field in connection with this task at this point.

To solve the problem posed above, we begin with a coordinate transformation, which morphs object #1, which is located in the physical domain, and for which we are trying to design the cloak, into object #2 residing in the virtual domain. Although this transformation is not obvious when the geometries of both the objects are totally arbitrary, it is nonetheless doable, at least theoretically. The caveat, though, is that the procedure provides us no guarantees that the materials parameters dictated by the transformation can be realized physically, and/or that they can be fabricated in practice to achieve cloak designs which satisfy the realistic specifications, such as small thickness, wide bandwidth, polarization insensitivity, etc.

The realizability issue, alluded to above, becomes even more critical when we attempt to design an “invisible” cloak, which is highly sought-after by cloak designers. We will now explain why this is the case with a simple example shown in Fig.5.2(a).

Let us assume, for the sake of convenience, that the target we wish to cloak is a cylinder. Following the TO paradigm, this problem has been extensively studied and the material
parameters for the cloak have been derived by invoking the TO, which makes the cloaked object disappear entirely (become invisible). In Figs.5.2(b) and (c) we plot the material parameters, presented in [37], that are required to make the cylinder invisible.

![Figure 5.2](image)

Figure 5.2. (a) Schematics for a cloaked PEC cylinder with \( R_2 = 2R_1 \), (b) material parameters for an all-angle, all-polarization cloak, (c) material parameters for a normal-incident, TE-polarization cloak from [37].

We observe several things from Fig.5.2. First, we see that the material parameters are anisotropic, and this in of itself can be problematic when we attempt to realize them in practice, because there is no systematic method available for synthesizing them. The second thing we observe about the required \( \mu \) and \( \varepsilon \) values is that they vary over a wide range, tending to 0 in some regions and \( \infty \) in others. Once again, this type of inhomogeneous behavior and a wide swing in the required material parameters as functions of the radial distance make it very difficult to realize them. In fact, we must resort to using Metamaterials (aka artificially-engineered materials) that are notoriously narrowband, dispersive and lossy where they attempt to realize the above type of material values. The above undesirable attributes that degrade the performance of the
cloak and render it unsuitable for most applications that involve scattering reduction for real-world targets. What exacerbates the problem even more is the fact that the thickness of the cloak is comparable to the wavelength, rather than being a small fraction of the same. As we well know, a thin coating is desired in most applications, e.g., when designing stealth targets for the radar world.

At this point we return to the TO paradigm for cloak designs and scrutinize it carefully to see if we can thresh out the root causes of the difficulties that we have just identified with the TO-based design, where upon introduction of Metamaterials for physical realization of the cloak structure. Note that polarization-insensitive Metamaterial designs for TO-based cloaks will be extremely challenging and the usual simplification is to limit the polarization to a certain linear polarization. This introduces polarization sensitivity of the cloaks. Another example is that typical Metamaterials are designed using periodic structures and the unit cell details are optimized for a single angle of incidence (usually normal to the interface). This introduces sensitivity to the angle of incidence.

To identify the problem areas with the TO-based cloak designs, we turn to an alternate derivation of the material parameters in the context of Transformation Electromagnetics. Rather than relying directly on the Jacobian of the transformation, which relates the Physical domain to the corresponding virtual domain, we turn to the integral forms of Maxwell’s equations appearing below:

\[
\oint E \cdot dl = -\frac{\partial}{\partial t} \iint \mu \cdot H \cdot dS
\]

\[
(5.2a)
\]

\[
\oint H \cdot dl = +\frac{\partial}{\partial t} \iint \epsilon \cdot E \cdot dS
\]

\[
(5.2b)
\]
Initially, we consider the relatively simple case where the two domains are simply related by scaling, say by a factor ‘γ’, as has been done in the previous TO-based cloak designs [9]. The method we propose to relate the material parameters in the two domains is very general and is applicable to the case where the two geometries have arbitrary shapes, and are not simply related to each other by a scale factor, as we will discuss later.

Figure 5.3 TO-based cloak schematics and corresponding materials in the (a) physical geometry and (b) virtual geometry.

To derive the material properties of the physical domain, from the assumed parameters in the virtual domain ($\varepsilon_0$, $\mu_0$ for free-space in this example), we turn to Fig. 5.3 and Eqn. (5.2). The next step, we choose to impose the condition that the fields ($E_1, H_1$) in the physical domain be “identical” to those in the virtual domain, i.e., ($E_2, H_2$). To facilitate the imposition of this condition, we now discretize the two domains, as shown in Fig. 5.4, by setting up a mesh to discretize the regions-2 and -3 in both domains. We take advantage of the circular symmetry of the geometries in the two domains, and of the fact that the geometry of the PEC cylinder, located in region-2 in the virtual domain is simply a scaled-down version of the one in the physical domain (region-1), and note that this transformation preserves the azimuthal symmetries of the two domains and, hence, the mesh size in the azimuthal direction remains unchanged when we
navigate between the two domains. However, we follow a different strategy in the radial direction, along which we impose the following three conditions:

(i) The number of radial cells be identical in region-2 of the domains, which spans the radial distance \( b < r < c' \) in the physical domain and \( a < r < c' \) in the virtual domain (note we have chosen \( c' \), the outer radius of region-2 to be identical in both domains);

(ii) The cell size in the outermost boundaries of region-2 is to be identical in the two-domains;

(iii) The cell sizes in region-3 in both domains be identically equal in the radial direction, not only between \( c' < r < c \), but also beyond, \( i.e., c < r < \infty \), meaning all the way to infinity.

![Cloak problem with mesh schematics: (a) physical and (b) virtual domains.](image)

There are logical reasons for imposing the above constraints, as we will now explain, before we proceed to derive the relationships between the material parameter distributions in the two domains. First, the TO must be applied to entire regions external to the PEC target—cylinder in this case—when relating the material parameters in the two domains, \( viz., b < r < \infty \) in the
physical domain and \( a < r < \infty \) in the virtual model. Specifically, we cannot truncate the regions as we transform from one domain to another, without introducing discontinuities in the fields, and thus violating the premise of TO under which we are operating. If we use condition (iii), in (5.2a and 5.2b), we immediately see that if the two fields as well as the mesh sizes would be identical in the two domains. It follows, then, that the material parameters must be exactly the same in regions-3 in the two domains. Since we have chosen the material parameters in the virtual domain in region-3 to be free-space, i.e., these parameters are \( \varepsilon_0 \) and \( \mu_0 \), then region-3 in the physical domain must correspond to free-space as well, as would the parameters of the external region \( c' < r < \infty \). This is a crucial point, and it implies that the cloak in the physical domain has but a finite thickness, spanning the region \( b < r < c' \). This is obviously necessary in order for the cloak design to be practical, since we cannot accept a cloak design whose thickness is infinite, as it would be if we did not impose the equal-mesh-size condition in the two domains in region-3 and beyond, \( i.e., \) for \( r > c' \).

To solve this problem we turn to the differential form of the Maxwell’s Equations given below:

\[
E \cdot \Delta l = -\frac{1}{\Delta t} \mu \cdot H \cdot \Delta S \\
(5.3a)
\]

\[
H \cdot \Delta l = +\frac{1}{\Delta t} \varepsilon \cdot E \cdot \Delta S \\
(5.3b)
\]

where \( \Delta l, \Delta S, \) and \( \Delta t \) represent the perimeter of a cell, surface area of a cell and time step in FDTD, respectively. Note that we are considering only the \( E_z \) and \( H_\phi \) components for this 2D geometry, namely a cylinder, for which \( z \) is along the axis of the cylinder, and \( \varepsilon \) and \( \mu \) in (5.3) are the appropriate elements of the \( \bar{\varepsilon} \) and \( \bar{\mu} \) tensors.
Let us now turn to the regions-2 in the two domains, namely physical and virtual. Recall that we have imposed the condition that the cell size at the outermost boundary of this region be identical in the two domains, which guarantees that the transition of the material parameters would be smooth as we transition from region-2 to region-3. (Recall region-3 and beyond is free-space in both domains.) So, all that remains for us to do now is to determine the material parameters of the cloak region in the physical domain, which spans from \( b < r < c' \), by invoking the condition that the fields \( (E_1, H_1) \) of the physical domain be identical to the fields \( (E_2, H_2) \) in the virtual domain (which we have chosen to be free space), as well as condition (i) on the number of cells in the two domains associated with region-2, namely that this number be identical in the two domains. At this point, we impose an additional condition, without loss of generality, that in the radial direction the cell sizes in region-2 in the physical domain be all equal, as we go from \( r = b \) to \( r = c' \). We also choose the cell size in the radial direction to be \( \lambda/20 \), though there is no hard and fast rule that says that we must adhere to this last condition, which is dictated more by the numerical discretization of the integral forms of Maxwell’s equations (5.3), than by anything else. At this point, we note that since the dimensions of region-2 in the physical and virtual domains are different, we typically choose the radius \( a \) such that \( a \ll b \), in order to ensure that the scattering from the small cylinder in the virtual domain would be small—in fact vanishingly small in the ideal case—in order to render it invisible as \( a \to 0 \) in the limit. Note that we must choose a non-uniform mesh in the virtual domain, so that we can simultaneously satisfy condition (iii), as well as the constraint on the number of cells in the radial direction in the two domains, namely that they be equal. Though we have some flexibility in terms of the variation of the cell size in the radial direction in the virtual domain, we choose this variation in \( \Delta r \) to be smooth, and monotonically increasing in terms of the cell size as we go from \( r = a \) to \( r = c' \), so that the summation of all the \( \Delta r \)'s equal \( (c' - a) \). An example of such a mesh is shown in Fig.5.5.
Having defined the meshes in the two domains, we finally turn to the task of determining the material parameters of the cloak in the physical domain. Recall that we wish to impose the criterion that the two sets of fields, namely \((E_1, H_1)\) and \((E_2, H_2)\) in the two domains, respectively, be identical. Eqn. (5.2) tells us that the \((\varepsilon_i, \mu_i)\) values in the physical domain, must be \(\varepsilon_0 / \gamma_n\) and \(\mu_0 / \gamma_n\), where \(\gamma_n\) is the ratio of the areas of the \(n^{th}\) cell in the physical and virtual domains, respectively. It is evident that, under these conditions, the \(\varepsilon_i\) and \(\mu_i\) must start out at \(\varepsilon_0 / \gamma_1\) and \(\mu_0 / \gamma_1\), where \(\gamma_1\) is the ratio of the dimensions of the first cell (at \(r = b\)) in the physical domain to that of the dimension of its counterpart (at \(r = a\)) in the virtual domain. Also, we recall from our previous discussion, that \(\varepsilon\) and \(\mu\) values are identical in both domains when we reach \(r = c'\), and that they are both just \((\varepsilon_0\) and \(\mu_0\)), i.e., material parameters of free space.
We now make two important observations. First, the cloak we have designed by following the procedure just described, which is based on the TO algorithm—though implemented differently than via the use of the Jacobian—varies inhomogeneously in the radial direction, albeit smoothly. Second, the $\varepsilon$ and $\mu$ values in the physical domain are larger, by a factor of $1/\gamma_1$, than their free-space counterparts. This factor is responsible for the root cause of difficulty encountered when designing TO-based cloaks, since it calls for Metamaterials to fulfill the requirements on the material parameters if we insist that the cloak must render the target totally “invisible”. This is because we can fulfill that condition if and only if $a \to 0$, in the strictest sense, and this implies that $\gamma_1$ must also follow suit and tend to zero as well.

We should mention that although we chose the simple geometry of the cylinder to identify the fundamental difficulties with the implementation of the TO paradigm, the problem with the realization of material parameters persists regardless of the geometry of the target, as long as we insist that it becomes invisible, which in turn requires that the scale factor (equivalent to $\gamma$) tends to 0.

Before we discuss our strategy for overcoming this fundamental roadblock, we examine another fundamental limitation posed by the TO paradigm when we attempt to reduce the thickness of the cloak $t \ ( = c' - b )$, to realistic values, e.g., a small fraction of the wavelength. We note that the material values ($\varepsilon_0$ and $\mu_0$) near the outer boundary of the cloak, located at $r = c'$, do not change, regardless of whether $r$ is large or small, and, for that matter neither do the behaviors of ($\varepsilon$ and $\mu$) of the cloak in the neighborhood of $r = b$, where the above parameters $\to \infty$. While we can attempt to partially mitigate this problem by imposing a cap on the values of these parameters, we cannot change the fact that the relative $\varepsilon_n$ and $\mu_n$ must reduce from very large values at the surface of the cylinder to unity within a relatively short distance; and this, in turn, again poses realizability problems in practice. For this reason, practical realizability of thin invisibility cloaks have not met with too much success in the past, and it is unlikely that the
thickness issue will be resolved anytime soon if we continue to impose the above invisibility criterion on the TO-based cloak designs. Fig.5.6 shows the deterioration of the performance of the cloak (see Fig.5.6(c) when we design a thin three-layer cloak by using discretized values of the cloak parameters, which vary continuously from $r = b$ to $r = c'$). It is evident that the wavefront of the field becomes considerably distorted when we compare the performance of the thin cloak with the ideal one the results for which are included in Fig.5.6 (b) for comparison.

Figure.5.6 Field distribution for A PEC cylinder in (a) an ideal TO cloak, (b) an ideal, thin TO cloak, (c) a 3-layered TO cloak in which the medium parameters at each layer correspond to those of the ideal thin cloak.

### 5.3 Field Transformation in the context of Generalized Scattering Matrix Approach

When designing an electromagnetic device, such as a lens antenna, we typically begin with a given source, such as a feed horn, and specify that the field radiated by the horn be transformed into a planar phase front, for instance. As is well known, this task is easily accomplished by placing the feed horn at the focal point of a conventional convex or plano-convex lens, and the procedure for designing such a lens based on the ray optics approach is well established in the literature; in fact, it is a just a textbook case.

The situation is different when we stipulate that the shape of the lens be flat instead of convex. To address this problem, the Transformation Optics (TO) offers an elegant recipe based on transforming the geometry of the original convex lens into a planar one. The TO algorithm
then shows us how to derive the requisite material parameters of the planar lens by using well-established relationships between the \( \varepsilon \) and \( \mu \) values of the original convex lens and its surrounding medium, and those of its planar counterpart. These relationships involve the Jacobians of the geometry transformation and are relatively straightforward to find, even for arbitrary geometries being transformed from the real space to virtual space. The caveat, though, is that the \( \varepsilon \) and \( \mu \) values are in general anisotropic and may be difficult if not virtually impossible to realize in practice. It is not uncommon, therefore, to set the \( \mu \) values equal to \( \mu_0 \), \( i.e., \) that of free space, to ignore the \( \varepsilon \) values less than unity, and to only work with isotropic dielectrics, albeit at the risk of compromising the performance of the lens in comparison to that of the original TO-design prior to introducing the modifications. What is equally important to realize is that there is no clear roadmap provided by the TO algorithm that tells us how we can improve the performance of the modified design, should we need to do so.

Given this background, we pose the following question for ourselves: Can we modify the problem statement that forms the basis of the TO algorithm to circumvent the problems alluded to above, without compromising the performance relative to that of the convex lens, in way such that we can still use realizable materials found in nature, and without having to resort to MTMs? We will now present an approach based on the Generalized Scattering Matrix (GSM) method [38], which indeed offers a way to address the problem at hand, as we have just enunciated above.

To introduce the GSM approach in the context of Field Transformation method we refer the reader to Fig.5.7, where we have defined the input and output ports to correspond to interfaces that bound an electromagnetic device. The field distribution in the input port, which is illuminated by the source located at the left of the port, can be expressed in terms of a set of coefficients \( \mathbf{a}_i \) (vector) associated with the basis functions employed to represent this “incident” field in the absence of the device when there are no reflections. Next, we insert the electromagnetic device, whose Scattering Matrix we desire to describe, inside the region bracketed by the input and
output ports. We define a set of coefficients $b_i^1$, again associated with the same basis functions as we used to define $a_i^1$, to represent the outgoing fields scattered by the electromagnetic device, i.e., the “reflected” fields that originate from the device and propagate back towards the source. We can similarly define a set of coefficients $c_i^1$, associated with the field distribution in the output port, through which these fields propagate in the free-space region to the right of this port, and are termed the “transmitted” fields. Our next step is to place the illuminating source to the right of the output port, which we have previously defined when the source was at the left, and reverse the roles of the input and output ports to correspond to the new source location. The incident, reflected, and transmitted fields are now characterized by a new set of coefficients $a_i^2$, $b_i^2$, and $c_i^2$, where $c_i^2$ fields now propagate to the left of the device, whereas the $b_i^2$ fields do the opposite, i.e., propagate to the right.

![Diagram](image.png)

Figure 5.7 Generalized Scattering Matrix (GSM) approach in the context of the Field Transformation (FT) method.

We are now ready to define the scattering matrix $[S]$, via (5.4) below, which we will use to characterize the device, as follows:

$$b = [S]a$$
or explicitly,

$$\mathcal{S} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}$$

where \( \mathbf{b} = [b_1, b_2] \) and \( \mathbf{a} = [a_1, a_2] \) represent the weights of the outgoing and incoming field representations at the input and output ports, respectively. Eqn. (5.2) provides us a convenient way to characterize an electromagnetic device in terms of its response to a plane wave, regardless of whether the illuminating source is incident from the left or the right of the device.

We will now explain how we can specify the desired characteristics of a device in terms of its S-parameter description by turning to the problem of reducing the level of scattering from radar targets—a problem that has been extensively researched into in the context of the TO. We will cast the problem in the language of Scattering Matrices, to help us understand why the TO approach leads us to untenable situations and/or to solutions which call for MTMs that are lossy, dispersive, polarization-dependent and have narrow bandwidths in terms of frequency of operation. We will also show how the FT approach mitigates the problems alluded to above by restating the design objectives and modifying them slightly from those associated with the TO-based designs.

Let us consider an arbitrarily shaped radar target placed in the region between the input and output ports as shown in Fig.5.8. Next, let us suppose that our objective is to reduce the level of scattering from the target, both in the forward and backward directions to the extent that the target becomes totally invisible to the incident field, say from an interrogating radar. We could cast this objective in the language of the Scattering matrices, by specifying that S11 be identically zero at the input port and S12 be such that the field distribution at the output port is identical to the incident field, as though the scatterer was totally transparent or invisible. This is precisely the premise upon which the TO algorithm is based, and the TO shows how to achieve this goal by
filling the center region, which contains the target, with materials that cloak the target to render it invisible, just as though it were absent. While such a goal is highly attractive, it is also very unrealistic, insofar as the physical realization of the cloak is concerned, for several reasons: first, the materials called for by the TO algorithm are highly anisotropic, even for simple targets such as cylinders; second, the requisite material values span a wide range, as shown in Figs.5.2 for a cylindrical target; third, these materials can only be realized by artificially synthesizing them, i.e., by using metamaterials that are inherently lossy, dispersive and narrowband (bandwidth is typically on the order of a few percent of the center frequency); fourth, the design is dependent on the incident angle as well as the polarization of incident wave—which makes it very impractical for real-world applications; fifth, the thickness of the cloak is much too large for real-world applications at microwaves; sixth, the TO algorithm does not provide a clue for systematically applying it to realistic targets that can be highly inhomogeneous—not made of just PEC but composite materials as well. (A literature research has failed to reveal the existence of such cloaks though we conjecture that there must have been many failed attempts to address this problem by workers in this field.)
Figure 5.8 Scattering from an arbitrarily shaped target described in the context of Generalized Scattering Matrix (GSM) method in the physical domain (a) and virtual domain (b).

To circumvent these problems that we encounter when attempting to use the TO to meet the ideal but unrealistic goal of making the target altogether invisible, we turn to the FT approach and modify our stated objectives, again in the context of S-parameters. In contrast to the TO, this
time we ask that $S_{11}$ be small--in terms of magnitude only--but not 0, as we demanded in the case of TO. Furthermore, we do not impose any restrictions on $S_{12}$, as we did in the case of the TO when we stipulated that the scattered fields at the output port be identically zero, so that the total field there be just the incident field. While we concede that the performance of the FT-based cloak or the blanket won’t be as ideal as the performance of the TO-based design would have been if we could realize it, we certainly stand to gain considerably when we follow this strategy since we can now obtain realizable solutions for the cloak, that are very wideband, covering the entire frequency range of 2-18 GHz, for instance, if we so desire. Furthermore, the cloak (or blanket) can now be very thin (only a few millimeters) and it would work for arbitrary incident angles and polarizations as well (an FT treated radar target is shown in Fig.5.9). In short, we can say that this strategy of following the FT-based algorithm to design the cloak enables us to circumvent all the problems we encounter when employing the TO-based design strategy instead. Of course, we compromise the performance of the cloak in the forward scattering direction when we use the FT-based strategy, though that is not a problem for either monostatic or bistatic radars that are only concerned with the backscattered fields.
Figure 5.9 Scattering from an FT-treated arbitrarily shaped target described in the context of Generalized Scattering Matrix (GSM) method.
Chapter 6
Scattering Reduction for Real-World Targets

Our objective is to design wideband layered absorbers for infinite, planar, conducting ground planes. We note that this is also the basic approach to designing coating for radar targets to render them stealthy, regardless of their shapes. Furthermore, it is relatively easy to carry out this design by using optimization algorithms to determine the material properties and the layer thicknesses to reduce the reflection from the coated PEC plane below -10dB level, and over a wide frequency band; say covering radar frequencies from 2 to 18 GHz, for example.

We now move to the second step in our design procedure, which is to adapt the blanket designed for the infinite PEC plane to an arbitrarily shaped object. Initially we consider an object with a smooth surface whose radius of curvature is moderate-to-large everywhere. We will generalize the procedure in the third step, using the principles of the TO when the above assumption regarding the smoothness of the object is not valid, as for instance when the object has sharp edges or bumps, as a general target would in practice.

When the object has a relatively smooth geometry, we initially wrap the multilayer absorbing blanket, which we have designed earlier for the planar surface around the PEC target whose scattering cross-section we are attempting to reduce, and test the effectiveness of the blanket for the new object. For a wide variety of targets we have examined, a number of which are shown in Fig.6.1, we have found that the blanket does reduce the monostatic as well as the bistatic radar cross-section in the “reflection” region near the surface of the object for different angles of incidence and polarizations of the incoming wave. The results for a two-layer absorber are shown in Figs.6.2 and 6.3 for a rectangular cylinder of finite length, which we have studied as a test case.
Figure 6.1 Back-scattering RCS of PEC objects: (a) plate, (b) pyramid, and (c) cylinder covered with 2 and 7-layer absorbers.
Figure 6.2 Phase behavior of the E-field near the rectangular PEC cylinder, which is wrapped around by an absorber blanket, normally incident on the cylinder.
Figure 6.3 Phase behavior of the E-field near the rectangular PEC cylinder, which is wrapped around by an absorber blanket, for an obliquely incident plane wave.

A simple test, which is typically applied to cloak designs, is to examine the wavefront of the total (incident+scattered) field, and see how the level of distortion of the wavefront decreases when the scatterer is covered by the layered absorbing coating. We present the plots of these wavefronts of the total fields for normal and oblique incidence cases for both polarizations in Figs. 6.2 and 6.3, respectively.

We observe that the object, which is a finite cylinder of height 18 cm, generates distorted wavefronts owing to the contribution of the scattered field from the object, even when covered by a two-layer blanket, designed for the infinite planar PEC object, for the nominal frequency range of 4.6-18 GHz, with a nominal reflection coefficient of -10 dB or less. However, we also note from Figs. 6.2 and 6.3 that the distortion in the phase front is relatively small once we go above
the low-end of the design frequency range, viz., 4.6 GHz for the planar geometry, confirming that
the planar design performs reasonably well even though we are dealing with a rectangular
cylinder now. We hasten to point out that the results presented in Fig. 6.2 are not for a coating
which has been optimized for the object at hand, and we expect some compromise in the
performance of the coating. However, we can improve this performance by optimizing the
parameters of the two-layer design, specifically the relative thicknesses of the two-layer, even as
we maintain the total thickness intact. We expect the changes to be relatively minor, however,
except for the corner regions and, hence, the optimization process should be realistic as well as
numerically feasible. The above remarks are also applicable to the oblique incidence case, for
which some sample results are presented in Fig. 6.3.

Figure 6.4 Target detection in conventional radar scenarios: mono-static and bi-static scheme
[39].
It is important to point out that the strategy for designing the absorptive coating, presented herein, is very different from that employed for ideal traditional TO cloak, since the latter is designed to render the (object+cloak) composite to have a zero scattering cross-section in all directions, whereas the blanket design introduced here seeks to reduce both backscattering and bi-static scattering scenarios but only in the reflection region, and not in the forward-scattering direction. We hasten to point out, however, that this type of performance is perfectly well suited for modern radar systems, symbolically depicted in Fig.6.4 above, where only the scattering in the reflection region is of concern.

For the final step, we consider the problem of absorber design when a shape perturbation is introduced in an object. Let us say that our modified target is the same rectangular cylinder we just considered above, except for a bump on the top surface. The extra corners introduced by the perturbation, be they smooth or sharp, would obviously introduce additional distortions in the planar phase front, and potentially increase the scattering level. Our objective here is to restore the field behavior so that it is close to that of the original object that we had prior to the introduction of the perturbation.
Figure 6.5 (a) Original 2-layer absorber wrapped around a rectangular cylinder with shape perturbation (left); (b) 2-layer absorber with TO-modified material properties around regions with the shape perturbation (left); (c) 2-layer absorber with TO modified material properties for all regions (left) and 2-layer absorber wrapped around a rectangular cylinder (right).

6.1 Field Transformation Algorithm, Material Modification

We now outline the procedure for the blanket design for the new object, shown on the left in Fig. 6.5, i.e., a (1), which is in the physical domain, and is a modified version of the one shown in the right side of the same figure; i.e., Fig. 6.5.(2), which corresponds to the virtual domain. Note that unlike the cylinder example we discussed earlier, the medium in the virtual domain, surrounding the object, is no longer free-space, as was the case shown in Fig. 6.5. Note also that the dimensions of the objects in the two domains are comparable, and are totally different from the legacy TO-design case, in which the scale factor between the dimensions of the object in the physical and virtual domains tends to infinity to render the target invisible.

To find the parameters of the cloak for the modified geometry in Fig. 6.5, we revisit the integral forms of Maxwell’s Equations, presented earlier in (5.2a) and (5.2b), to relate the material parameters associated with the two systems shown in the figure. Fig. 6.5.(1) shows the physical system with locally modified medium parameters for the perturbed object, while Fig. 6.5.(2) depicts the virtual system with the original medium parameters covering the unperturbed object. The field distribution near the perturbed object would obviously be different from that of its unperturbed counterpart, since the perturbation introduces additional scattering to the incoming wave. We link the change in the field with the modification in the geometry, and then compensate it by changing the medium parameters, which are illustrated by a slight change of colors from the ones representing the original planar design in Fig. 6.5.(1.a). Given the specific profile of the perturbation and the simulated field distributions, the only unknowns, namely the medium parameters, can be derived as follows:
\[ \mu_1 = \frac{\oint E_1 \cdot dl_1}{-\frac{\partial}{\partial t} \iint H_1 \cdot dS_1} = -\frac{\partial}{\partial t} \iint H_1 \cdot dS_1 = \frac{\oint E_2 \cdot dl_2}{-\frac{\partial}{\partial t} \iint H_2 \cdot dS_2} \left( -\frac{\partial}{\partial t} \iint H_2 \cdot dS_1 \right) \left( -\frac{\partial}{\partial t} \iint H_2 \cdot dS_2 \right) \]

\[ \mu_1 = \frac{\oint E_2 \cdot dl_1}{\oint E_2 \cdot dl_2} \cdot \left( -\frac{\partial}{\partial t} \iint H_2 \cdot dS_2 \right) \cdot \mu_2 \]

\[ \epsilon_1 = \frac{\oint H_2 \cdot dl_1}{\oint H_2 \cdot dl_2} \cdot \left( +\frac{\partial}{\partial t} \iint E_2 \cdot dS_2 \right) \cdot \epsilon_2 \]

(6.1a)

(6.1b)

where we have used scalar quantities for simplicity in Eqn.(6.1) as though the geometry under consideration is two-dimensional, which it is in the present example, and we would need to replace the field quantities with vectors and the material parameters with tensors for the general 3D case.

The field behaviors for the perturbed object with locally modified material parameters can be seen from Figs.6.6 and 6.7. Fig.6.6 shows that the amplitude of the scattered E-field is reduced, and that the phase front of the total E-field is approximately restored as well in the reflection region. Fig.6.7 compares the amplitudes of the scattered electric fields for five different scenarios listed in the figure.
Figure 6.6 Phase behavior of scattered E-field for (a) perturbed object wrapped by blanket with original medium parameter; (b) perturbed object wrapped by blanket with locally modified medium parameter.

Figure 6.7 Comparison of the amplitudes of electric fields scattered by different objects.

We should clarify the fact that although we are referring to this geometry as a slab, what we are really dealing with is a wide rectangular cylinder, with a small thickness.
We note that the modified slab does introduce additional scattering, and that the absorber does help reduce the same. We also note that the modified absorber improves the performance over the initial one, but only slightly, which shows that the planar version of the cloak is not all that inferior to the one modified for this type of geometry. Additional optimization of the modified cloak is expected to improve the performance even further, if so desired.

Figure 6.8 Possible domain decomposition of two aircrafts: F-16 Falcon fighter jet (left) and Predator Drone UAV (right) for scattering-reduction treatment.

For an arbitrary target, we can first decompose the geometry of the target in a manner illustrated in Fig. 6.8 into several large “blocks” that closely resemble previously investigated objects, and wrap each part of the target with an absorber, designed by using the approach based on shape perturbation of a related smooth object. We then follow the methodology we have described above to determine the material parameters of the coating. It should be evident that this is a far more realistic approach than transforming the geometries of these complex objects into an infinitesimally small-size target, as called for by the TO algorithm for cloak designs, and following the TO recipe corresponding to such a geometry transformation, which is bound to lead to unrealistic and impractical material parameters and/or structural elements.
6.2 Field Transformation Algorithm, Thickness Modification

Modifying the permittivity and permeability parameters of the lossy materials is not the only approach one could take to reducing the scattering from objects that have curved profiles, especially when laboratory resources to fine-tune the material parameters in accordance with (6.1a) and (6.1b) are not available.

Changing the composition of the absorber can be quite effective for enhancing the absorbing performance of the coating. Fig.6.9 shows the RCS results of a PEC cylinder of radius 3cm, illuminated by a plane wave. The bi-static RCS evaluated at 10GHz, which corresponds to a wavelength of 3cm. Various thickness compositions are analyzed to determine the one most suited for this particular cylinder, while the total thickness of the absorber is kept unchanged at 3mm.

Fig.6.10 shows the result for another example of a PEC sphere of radius 30cm illuminated by a plane wave. The bi-static RCS is again calculated at 10GHz. Various thickness compositions are analyzed to determine the one that provides an optimal performance. Once again, the total thickness of the absorber is kept unchanged at 3mm.

We note from the figures that the optimal choice for a planar absorber is not necessarily the best one for a curved object. A moderate performance enhancement in terms of absorption can be achieved by simply changing the thickness of the absorber, as we might expect.
Figure 6.9 Comparison of bi-static RCS for a long PEC cylinder coated with absorbers with various thickness compositions, for horizontally (up) and vertically (down) polarized illumination.
Figure 6.10 Comparison of bi-static RCS for a PEC sphere coated with absorber with various thickness compositions.

It should be pointed out that layered absorbing materials, which comprise of more than two types of lossy components, and/or more than two layers, are likely to be better suited for the task of coating objects with a curved profile. Fig. 6.11 shows the example of a five-layer absorber designed for a planar surface using the optimization techniques discussed previously. Wideband absorption performance is achieved from 2 to 16.9GHz with a total thickness of 4.26mm (as opposed to a frequency range of 4.6 to 18GHz and a total thickness of 3mm) [40]. The increased number of layers provides increased degrees of freedom, which could potentially achieve the required level of performance without having to resort to synthesis of new material.
Optimization techniques can still be efficiently applied to design absorptive coatings for objects with cylindrical or spherical shapes, since the fields scattered from these canonical shapes can be analytically expressed as functions of the material parameters and compositions of the coatings. When the shapes of the objects are no longer canonical, as in the case in real-world scenarios, applying optimization technique may be a time-consuming task. This is because the forward problem, namely solving for the electric field distribution through numerical simulation—since analytical calculation is not an option here—needs to be carried out for each and every object, for which we are designing the absorber. It is typically computer-intensive, and
costly both in terms of CPU time and memory. Thus the use of powerful optimization algorithms is highly recommended for designing these absorbers.

Fig. 6.12 show the results for a PEC ellipsoid whose major and minor axes are 15cm, 3cm and 3cm respectively and which is illuminated by a plane wave with horizontal and vertical polarizations. The bi-static RCS is calculated at 10GHz, which corresponds to a wavelength of 3cm. Two absorber compositions are analyzed to investigate the performances of these two configurations. The first one of these uses the original absorber design for the planar structure, while the second corresponds to the optimal thickness composition we presented in Figs. 6.9 and 6.10. The total thicknesses of the two configurations are kept unchanged and it’s 3mm in both cases.
Figure 6.12 Comparison of bi-static RCS for a PEC ellipsoid coated with absorber of various thickness compositions, under horizontally (up) and vertically (down) polarized illumination.

It is evident from Fig.6.12 that the modified composition exhibits a slightly enhanced performance for both polarizations. However, we should point out that there is no guarantee that
an improvement can always be achieved over that of the original planar design, and that an optimal design for the absorber coating can indeed be derived, for a given object by simply varying the layer thicknesses using a trial-and-error approach. An alternative technique is to modify the material parameters, as illustrated above, which provides a more straightforward solution, especially for geometries shaped like the ellipsoids.

6.3 Practical Applications

Reduction of RFI, another important application for the absorber-wrapping is to mitigate the problem of antenna blockage in a shared-platform environment, as shown in Fig.6.13, in which the introduction of the aggressor antenna can raise the far-end sidelobe levels of the parabolic dish significantly. To mitigate this effect, we can wrap the monopole by using a multilayer absorber optimized for a planar geometry, as we have done in the past, or we can optimize the layer thicknesses and material parameter for the circular cylinder geometry. Alternatively, we can use a conducting saucer-like structure wrapped around the monopole, as proposed in [41-43]. Fig.6.14 shows the model used to construct the alternative cloak with the monopole as the cloaked target. Fig.6.15 shows the far field patterns of a parabolic dish antenna—the victim antenna—when a monopole, which is the aggressor antenna, is placed in the vicinity of the dish. The patterns of the dish/monopole composite are also included in the figure for two treatment plans applied to the aggressor antenna. We note that the introduction of the absorber treatment reduces the sidelobe levels of the dish antenna, as compared to the case for the far-end antenna combination without the treatment.

However, we also note that an elaborate cloak design, shown in Fig.6.14, isn’t really needed, and same level of performance can be achieved with a thin absorber type of coating.
Figure 6.13 Model schematics for the blockage problem: (left) side view and (right) isometric view of the victim antenna (parabolic dish) and the aggressor antenna (monopole) sharing the same platform.

Figure 6.14 Model schematics for the alternative cloak.
Figure 6.15 Radiation patterns for the antenna blockage problem: (blue) dish antenna, (green) dish antenna and the monopole antenna in the vicinity and (red) dish antenna with absorber-treated monopole antenna.

Recent advances in material engineering and research in graphene-based absorbing materials also broadens the choice of the materials used for the absorber designs discussed in this paper. Graphene-based absorbing materials generate sufficient level of magnetic loss without the need to introduce magnetic elements such as Co, Fe and Ni. These new materials also have the advantage of being lightweight, and thin, that are highly desired attributes for airborne applications. There have been reported cases of achieving below -70dB reflection level shielding.
with 2.09mm of graphene nanoplatelet-synthesized material [43], as well as below -20dB reflection level over the entire 4.5-18GHz with 8mm of multilayered carbon nanotube fabric [44].

The performance of the absorbers can also be enhanced by introducing FSS-type structures within the layers, which add virtually no extra thickness to the composite structure. The FSS structures can be tailored to enhance the absorption in target frequency bands [16] or to broaden the working bandwidth [44] of the initial layered absorber.
Chapter 7

Suggestions for Future Works

Several directions can be explored in the future that could further complement the works in this dissertation.

First, all the FSS type absorbers investigated in this dissertation have been strictly periodic. Aperiodicity can be introduced in the periodic elements to further enhance the bandwidth performance of the absorber screen. Statistical analysis can be applied to an absorber screen with a large enough number of elements so that the absorption level at specific frequency bands can be tailored (to an extend) and a systematic approach can be developed to determine the relationship between the absorbing performance and the size as well as the number of a single element.

Secondly, the algorithm developed in this dissertation for determining the modified material parameter for objects with local shape/structural perturbations works well when the shape perturbation does not have a small radius of curvature, \textit{i.e.}, no sharp corners. It is likely that the calculated material parameter will vary greatly in the vicinity of a sharp corner. This means these results might not fall into the range of achievable material parameters due to the constraints of the synthesizing process. A systematic approach of realizing artificial material with desired material parameter, \textit{e.g.} Dial-a Dielectric (DaD) approach, would greatly increase the flexibility of the algorithm.

The wideband absorbers discussed in this dissertation have been designed to have continuous absorption performance. It would be desirable for antennas on some airborne vehicles to be functional in their operational frequency and to have low scattering levels outside the
working frequency window. Band-notch periodic structures can be investigated to meet the requirements for this application.
Bibliography

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