High Impedance Surface Using A Loop

With Negative Impedance Elements

by

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ABSTRACT

Antennas are required now to be compact and mobile. Traditional horizontally polarized antennas are placed in a quarter wave distance from a ground plane making the antenna system quite bulky. High impedance surfaces are proposed for an antenna ground in close proximity. A new method to achieve a high impedance surface is suggested using a metamaterial comprising an infinite periodic array of conducting loops each of which is loaded with a non-Foster element. The non-Foster element cancels the loop’s inductance resulting in a material with high effective permeability. Using this material as a spacer layer, it is possible to achieve a high impedance surface over a broad bandwidth. The proposed structure is different from Sievenpiper’s high impedance surface because it has no need for a capacitive layer. As a result, however, it does not suppress the propagation of surface wave modes.

The proposed structure is compared to another structure with frequency selective surface loaded with a non-Foster element on a simple spacer layer. In particular, the sensitivity of each structure to component tolerances is considered. The proposed structure shows a high impedance surface over broadband frequency but is much more sensitive than the frequency selective surface structure.
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1 INTRODUCTION

In the past two decades, much electric equipment has been asked to be compact and mobile operating in a broad range of frequencies. Recent studies have looked at an integrating antenna and its ground in closer proximity. Typically, antennas are placed on a perfect electric conductor (PEC) at a quarter wave distance to maximize their efficiency. For example, a dipole antenna operating at 350 MHz should be placed in a distance of about 21 cm (or about 8.44 inches) placed horizontally from a ground plane. In many applications, the antenna must be placed much closer to the ground plane. While it can be a good reflector, it is not a good ground since the phase of the incident wave is reversed according to the image theorem [1]. It cancels the phase of incident and reflected waves and shows low radiation efficiency. This constraint of the antenna system has been an obstacle to realize a low-profile and compact system.

A photonic bandgap (PBG) structure, also known as an artificial magnetic conductor (AMC), can been considered as a ground plane for low profile antennas in order to reduce the distance between an antenna and its ground by filling the spacer with an artificial medium. The height, h, can be calculated by

\[
h = \frac{c_0}{f_{\text{resonant}}} \cdot \frac{1}{4}.
\]  

(1.1)

In the past decades, the designs of periodic structures have been developed to control the propagation characteristics of electromagnetic fields [2]. They enable the development of materials that exhibit novel electromagnetic properties which
Figure 1-1: Sievenpiper’s artificial magnetic conductor and its cross section are not available in nature and can be easily optimized for the desired application. These materials are known as metamaterials [3]. In 1999, Sievenpiper and Yablonovitch first introduced a high impedance surface for the microwave and antenna domains [4]. They suggested mushroom-type surfaces which are similar to corrugated surfaces except for the fact that they exhibit two-dimensional periodicity in order to prevent the propagation of both vertically and horizontally polarized waves along its surface [5].

In the following chapters, a customized artificial magnetic conductor (AMC) for ultrahigh frequencies (UHF) is suggested that consists of a spacer, a frequency selective surface (FSS), and a ground surface as shown in Figure 1-1. The AMC can be modeled as shown in Figure 1-2 for normal wave incidence by substituting
a transmission line for the spacer and a capacitor for the FSS. Because the bandwidth is related to the permeability and the thickness of the spacer as

\[
\frac{\Delta f}{f_{\text{resonant}}} = 2\pi \mu_r \frac{h}{\lambda_0}\]

the spacer should be designed to get a desired permeability [6]. A short explanation is in Appendix 1. When the thickness is assumed to be 1.7 inches and the required frequency range is assumed to be 200 MHz to 500 MHz, for UHF application, the relative permeability should be greater than 2.7.

The high permeability spacer is realized by embedding electrically small artificial magnetic molecules (AMMs) in a host medium. The AMM is formed by electrically small loop antennas loaded with passive electrical circuit elements [7]. The loop can be represented by the equivalent circuit shown in Figure 1-3. The

Figure 1-3: Equivalent transmission model of Sievenpiper’s AMC layers.
magnetic polarizability of the artificial molecule can be expressed as

\[ \alpha = \frac{m}{H} = \frac{-j\omega \mu_0 a^4}{R_{\text{loss}} + j\omega L_{\text{loop}} + Z_L}. \]  

(1.3)

Then, the relative permeability of the material is

\[ \mu_{zz} = 1 + N\alpha. \]  

(1.4)

Here, it is assumed that the coupling interactions between molecules are ignored, and the host medium has the same relative permeability as that of free space. To cancel the inductance of the loop inductance, \( L_{\text{loop}} \), the passive load must be \(-j\omega L_{\text{load}}\), which is a negative inductance. When \(|L_{\text{load}}|\) is slightly higher than \(|L_{\text{loop}}|\), the relative permeability will be increased greatly.

A simple circuit simulation result is shown in Figure 1-4. When the negative inductance is controlled as desired, the relative permeability is also controlled. The next chapters will show how to take out the characteristics of the loop and to cancel its inductance in a unit cell, which is a part of a periodic structure.
Figure 1-4: A theoretical simulation result for a loop loaded with a negative inductor.
2 ARTIFICIAL MAGNETIC MOLECULES

An effective low-profile ground plane for an antenna requires certain characteristics to support the effective radiation of the antenna with reduction of undesired back lobes and reactive coupling to nearby circuits. The structure can be a high impedance photonic bandgap (PBG), also known as an artificial magnetic conductor (AMC). A designed AMC helps to realize a planar low profile antenna. For a bandwidth between +90° and -90° phases shown in Figure 2-1, the frequency bandwidth should be at least 300 MHz with a center frequency of 350 MHz. According to Equation (1.2), a spacer layer thickness of 1.7 inches must have the permeability of the spacer greater than 2.7. An artificial material is designed with an effective permeability of about 3.6 to achieve broad bandwidth.

Figure 2-1: Definition of AMC bandwidth as the range of frequency for +90° to -90° phase.
2.1 EXTRACTION OF MATERIAL PROPERTIES

To achieve the desired bandwidth in a given form factor, we decided to make a spacer with a relative permeability of 3.6. To verify the characteristics of the material, a procedure should be established to extract the material properties. The material is simulated in Ansoft HFSS (High Frequency Structure Simulator) for three-dimensional full-wave electromagnetic (EM) simulations. The material to be investigated is placed in the middle of a waveguide which has two excitation ports on both end sides as shown in Figure 2-2. The waveguide is set with one pair of opposing sides with perfect electric conductor (PEC) walls and the other pair with perfect magnetic conductor (PMC) walls to realize a transverse electromagnetic (TEM) waveguide. From image theory this is equivalent to an infinite periodic structure in the x-y plane. Note that PMC boundaries can be replaced by symmetry boundaries in the simulator.

The procedures to numerically evaluate the effective medium properties of a given material are developed. The material S-parameters are extracted from HFSS, exported, and post-processed in Mathworks MATLAB.

First, the 2 port TEM waveguide containing a material sample is simulated in HFSS and the S-parameters are extracted. Then, the data is calculated in Mathworks MATLAB. The reference plane is shifted so that the data is only for the material under test (MUT).

\[
[S] = [R][S_{HFSS}][R]
\]  
(2.1)

where
$[S_{HFSS}]$ is S-parameters from HFSS analysis,

$[R] = \text{diag}(e^{i\theta})$,

$\theta = k_0(L - d)$,

$L = \text{overall length of TEM waveguide}$, and

d = thickness of material sample.

The parameters are then converted to ABCD parameters [8].

\begin{align*}
A &= \frac{(1 + S_{11})(1 - S_{22}) + S_{12}S_{21}}{2S_{21}}, \\
B &= Z_0 \frac{(1 + S_{11})(1 - S_{22}) - S_{12}S_{21}}{2S_{21}}, \\
C &= \frac{1}{Z_0} \frac{(1 - S_{11})(1 - S_{22}) - S_{12}S_{21}}{2S_{21}}, \\
D &= \frac{(1 - S_{11})(1 + S_{22}) + S_{12}S_{21}}{2S_{21}}.
\end{align*}

(2.2)

Then, the material’s propagation constant ($\gamma$) and characteristic impedance ($Z_c$) of equivalent transmission line are calculated as

Figure 2-2: A HFSS simulation setting with material under test in a TEM waveguide.
arccos \left( \frac{A + D}{2} \right) \) and \( Z_c = \sqrt{\frac{B}{C}} \). \hspace{1cm} (2.3)

Finally, the material properties are evaluated using

\[ \mu_r = -j \frac{\gamma Z_c}{k_0 \eta_0} \quad \text{and} \quad \varepsilon_r = -j \frac{\gamma \eta_0}{k_0 Z_c} \] \hspace{1cm} (2.4)

where \( k_0 \) is the propagation constant and \( \eta_0 \) is the wave impedance of free space.

To verify the validity of the approach, some known test materials are examined. For instance, a 4-cm slab of a lossy magneto-dielectric material with its relative permittivity \( \varepsilon_r = 2.2 - j2.2 \) and relative permeability \( \mu_r = 3.6 - j3.6 \), is placed in a 50-cm long TEM waveguide. The post-processed results are shown in Figure 2-3. Notice that both graphs closely match both the real and imaginary parts of relative permittivity and relative permeability.

### 2.2 DESIGN OF AN ARTIFICIAL MAGNETIC MATERIAL

The characteristics of any material in the TEM waveguide are extracted using the approach described in the previous section. A proposed artificial magnetic molecule (AMM) in the previous chapter consists of a loop loaded with a negative inductance, \( |L_{load}| \). The value of \( |L_{load}| \) is initially approximated using an equivalent circuit of a square loop [1]. \( L_{loop} \) in Figure 1-3 is estimated as following:

\[ L_{loop} = 2\mu_0 \frac{a}{\pi} \left[ \ln \frac{a}{b} - 0.774 \right] \] \hspace{1cm} (2.5)
where $a$ is a side length of a square loop and $b$ is a wire radius. For a loop with a side length of 1.2 inches, the other side length of 0.5 inches, and the wire width of 75 mil, $a$ is assumed to be 0.85 inches and $b$ is 37.5 mil. Then $L_{\text{loop}}$ is calculated roughly to be 40 nH. With this assumption, a loop loaded with a negative element is implemented inside a waveguide in HFSS as shown in Figure 2-4.

A negative inductance, $L_{\text{load}}$, assumed to be slightly larger than $L_{\text{loop}}$, is placed inside the loop since it is a frequency domain solver. HFSS allows inputting negative lumped element numbers even though the negative impedance is a non-Foster element. Since the loop inductance was roughly calculated, the value of $L_{\text{load}}$ is determined by a trial-and-error procedure.

Figure 2-5 shows a computed result simulated with $L_{\text{load}}$ of -30.5 nH. A desired relative permeability was achieved at a lower frequency than its resonant frequency. Since the design is targeted to have a broadband application, the resonant frequency should be moved to a higher frequency. However, the simulations showed that a higher resonant frequency has lower permeability at low frequencies. Those two relations are plotted in a coordinate shown in Figure 2-6. This relation is similar to Sneok’s limit, which is a relationship between frequency of maximum absorption and permeability at low frequency [9].
Figure 2-3: Relative permittivity (top) and relative permeability (bottom) of the sample material for the TEM waveguide in Figure 2-2.
Figure 2-4: An AMM configuration in a TEM waveguide.

Figure 2-5: Relative permeability for AMM realization with $L_{load} = -30.5$ nH.
2.3 EXTRACTION OF MATERIAL PROPERTIES

J. L. Snoek first stated Snoek’s limit, which is a rule for magnetic materials. The higher the permeability, the lower the frequency at which absorption and dispersion set in; the product of the resonant frequency and the initial permeability is constant.

The designed spacer also exhibits the Snoek-like characteristic feature. The product of the resonance frequency and the DC (low frequency) permeability is constant as shown in Figure 2-6, which is reminiscent of Snoek’s limit [9]-[11]. A product value is desired to be higher and the relationship between the resonant frequency and the relative permeability at low frequency is plotted as in the right

![Graph showing the relationship between resonant frequency and relative permeability at low frequencies.](image)

Figure 2-6: AMM relationship between resonant frequency and relative permeability at low frequencies.
top corner in the coordinate as possible. Variations in the geometric dimensions cannot overcome this limitation. It can be considered that resonance is a result of the parasitic capacitance mode of the loop itself.

The parasitic capacitance of the loop can be compensated by a negative capacitance as seen in Figure 2-7. The loop with the cancelled parasitic capacitance exhibits a high relative permeability over a broader frequency range. An example of the achievement is shown in Figure 2-8 with $|L_{neg}|$ of 37 nH and $|C_{para}|$ of 0.32 pF. The presence of the negative capacitance causes the resonant frequency to move higher than the frequency of interest. The relative permeability exhibits close to a constant value now.

Figure 2-7: An equivalent circuit of the loop with a negative inductance and a negative capacitance to compensate the Snoek-like limitation.
CHARACTERISTICS OF ARTIFICIAL MAGNETIC MOLECULES

The artificial magnetic molecule is designed to achieve a relative permeability of about 3.6 in the UHF bandwidth from a loop loaded with negative impedance. The loop in the middle of a TEM waveguide is analyzed in the reflection testbed as shown in Figure 2-9. A waveport for exciting an incident wave is located on the opposing plane and collects the S-parameter information from the reflection coefficients. With the change of HFSS environment settings, the calculation in MATLAB is also changed for one-port reflection calculation. Then, a constant relative permeability is achieved as seen in Figure 2-10.
The AMM-AMC comprising loops loaded with negative impedances have been successfully simulated. This AMC will exhibit only TE surface wave mode suppression because it has no vias. The demonstrations show featured spacer layers with AMMs arranged only in a one-dimensional lattice. Such a structure has AMC properties that exist only for one cardinal polarization of the incident wave, which is the one whose magnetic field interacts with the loops. An extension of this structure has also been successfully simulated. It has a two-dimensional AMM lattice allowing for interaction with waves polarized in either a cardinal direction or indeed polarized in an arbitrary direction, since any arbitrary polarization is a superposition of these two polarizations for a normally incident plane wave.

Before placing two crossed substrates in the reflection testbed simulation, a split loop in a single plane is verified. In the simulation the negative impedance consisting of a negative inductor and a negative capacitor in parallel is split in half. The values of each inductor and capacitor are respectively half and twice as large as the previously simulated loop. Thus, $|L_{neg}| = 37 \text{ nH/2}$ and $|C_{neg}| = 0.32 \text{ pF } \times 2$ and they are placed as shown in Figure 2-11.

The two-dimensional AMM lattice geometry and simulation results are shown in Figure 2-11. Note that the negative impedance elements are tuned to $|L_{neg}| = 40 \frac{\text{nH}}{2}$ and $|C_{neg}| = 0.32 \text{ pF } \times 2$ to achieve the desired properties.

The simulated results show that the spacer layer exhibits more or less constant permeability over UHF frequencies.
Figure 2-9: One-port reflection testbed in a waveguide and its close-up in HFSS simulator.
To evaluate the effective permeability of the spacer layer, the reflection testbed is terminated by a PEC ground plane. To evaluate the effective permittivity, the ground plane is changed from PEC to PMC as shown in Figure 2-12. Both results agree very well at the lower frequencies and diverge at the higher frequencies. It is expected that the results of the two-port testbed simulation are more accurate.

The designed UHF AMM-AMC has a relative permittivity of 1.3+j0 and a relative permeability of 3.6+j0.

Figure 2-10: Relative permeability of the AMM simulated on the one-port reflection testbed.
Figure 2-11: Two half loops with the respective $|L_{neg}|/2$ and $|C_{neg}| \times 2$. 
Figure 2-12: Oppositely polarized loop and resulting material characteristics.
Figure 2-13: Spacer with two-dimensional lattice and its relative permeability.
Figure 2-14: Two dimensional AMM lattice on PMC testbed.
3 FREQUENCY SELECTIVE SURFACES

The design of a spacer layer for an artificial magnetic conductor (AMC) has been discussed in the previous chapter. To provide anti-resonance at the frequency of interest, a capacitive layer is usually added on a spacer layer as a Sievenpiper’s model. Therefore, the structure will emulate high-impedance condition, experienced by an antenna placed one quarter wavelength above a PEC.

The design of a capacitive layer is based on the concept of a frequency selective surface (FSS). In some sense, this terminology is appropriate in the context of AMCs since the capacitive layer design will also influence the surface-wave suppression properties of an AMC.

In this chapter, we design a FSS to obtain a desired reflection coefficient phase behavior in an AMC, which is incorporated with the previously designed spacer layer. Then, we analyze the surface wave properties of the resulting AMC using the effective media model implemented in MATLAB and the structural simulation in HFSS.

3.1 DESIGN OF FREQUENCY SELECTIVE SURFACES

The AMC is modeled as a two layer bi-uniaxial model as shown in Figure A-1. The spacer is represented as $Y_1$, the FSS as $Y_2$, and the radiation space as $Y_3$. The reflection coefficient, $\Gamma$, is derived for both polarizations of the incident wave, but usually a normal angle of incidence is assumed, $\theta_{inc} = 0^\circ$. For example, the
reflection coefficient for the TM mode is shown in Equation (3.1) based on the definitions of Equation (A.6)-(A.8).

\[
\Gamma = \frac{Y_3 - Y_M}{Y_3 + Y_M}
\]  

(3.1)

where

\[
Y_M = Y_2 \frac{Y_L + Y_2 \tan(\gamma_2 d_2)}{Y_2 + Y_L \tan(\gamma_2 d_2)} \quad \text{and} \quad 
Y_L = Y_1 \coth(\gamma_1 \cos \theta_1 d_1).
\]

The magnitude and the phase of the reflection coefficient for a normal incidence are shown in Figure 3-1. Notice that each admittance, \(Y\), is calculated as in Appendix A. With the obtained relative permittivity and relative

![Figure 3-1: Reflection coefficient phases of an AMM show the bandwidth frequency of 360 MHz.](image.png)
permeability of the spacer from the previous chapter, an initial design can be obtained using the effective media model in MATLAB. Based on this MATLAB model, the geometry was simulated in HFSS as seen in Figure 3-2.

It was needed to adjust the geometry several times in order to obtain the desired reflection coefficient behavior in the HFSS simulations. We attribute this need for iteration to inaccuracies in the calculation of the FSS layer capacitance in the MATLAB model.

Figure 3-2: FSS implementation on a spacer layer.
3.2 INVESTIGATION OF DISPERSION DIAGRAM

The AMC design discussed above exhibits only TE surface wave mode suppression. To suppress TM surface waves, we need to add properly positioned vertical pins (vias) to the design. Using MATLAB and the dimensions that correspond to Figure 3-1 results, the dispersion diagram is plotted in Figure 3-3. Since the TE bandedge is already fixed with the FSS dimension, only the TM bandedge can be adjusted. In addition, the AMC should be appropriately designed in order to suppress all surface waves as shown in Figure 3-3.

![Dispersion Diagram for the designed AMC, Pvia=1.3times Pfss](image)

Figure 3-3: FSS implemented on the spacer showing the bandgap from 280 MHz to 500 MHz.
As discussed in Chapter 2, a high impedance surface is achieved over a broad bandwidth without a FSS layer and with a FSS layer. Because their sensitivities to non-Foster elements are different, S-parameter’s sensitivities to the changes of negative lumped elements are investigated.

For a spacer layer only structure as shown in Figure 2-9, the negative inductance is changed to see how much it affects its S-parameter. With \( |L_{neg}| \) of -44 nH, the magnitude and angle of \( S_{11} \) are shown in Figure 4-1. It shows the center frequency of 300 MHz and its bandwidth of 200 MHz.

First, a spacer loaded with a negative inductance is investigated without the Snoek-limit consideration. The angle of \( S_{11} \) at the frequency of 300 MHz, the initial center frequency, is plotted as shown in Figure 4-2. As the inductance is changed by -0.5 % to 1.3 %, the angle of \( S_{11} \) experiences changes of +30 degrees to -30 degrees. The rate of the change is expressed by

\[
\frac{\Delta \angle \Gamma}{\Delta L_{neg}} \approx -80 \sim -65 \text{ deg nH} \quad (4.1)
\]

Since the loop shows the Snoek-like limitation, the spacer should have negative capacitance to cancel the loop parasitic capacitance. When the non-Foster capacitance is included, the sensitivity rate is sharply increased by

\[
\frac{\Delta \angle \Gamma}{\Delta L_{neg}} \approx -220 \text{ deg nH} \quad (4.2)
\]

as shown in Figure 4-3. The sensitivity of the structure is also compared with a sensitivity of the structure with a FSS layer.
Figure 4-1: S-parameter magnitude and angle simulated a spacer only model.
Figure 4-2: Change of angle of S-parameter by change of inductance for a spacer-only structure with |$L_{neg}$|.

Figure 4-3: Change of angle of S-parameter by change of inductance for a spacer-only structure with |$L_{neg}$| and |$C_{para}$|.
The property of a simple spacer layer is first extracted and a FSS structure with negative elements is combined as shown in Figure 4-4. Then, the spacer capacitance is cancelled out using non-Foster elements on the FSS as shown in Figure 4-5. The S-parameters are shown in Figure 4-6.

Now, the sensitivity of the angle of $S_{11}$ is investigated with respect to change of negative inductance. The relationship is plotted in Figure 4-7.
Figure 4-5: A FSS pads combined with non-Foster elements and its equivalent circuit.
Figure 4-6: Magnitude and Angle of S-parameter from FSS structure controlled by non-Foster elements on its layer.
The change of the angle of S11 at the center frequency of 300 MHz is observed and its relationship is shown as

\[
\frac{\Delta \angle \Gamma}{\Delta L_{neg}} \approx -7.8 \sim -7.6 \text{ deg/nH.}
\]  \hspace{1cm} (4.3)

It shows that the angle of S11 varies from 30 deg to +30 deg as \( |L_{neg}| \) changes by -5% ~ 7%. That shows the FSS layer with non-Foster elements will be less sensitive than the spacer-only layer with the loop loaded non-Foster elements on it.
CONCLUSION AND FUTURE WORKS

An AMC structure implementation for a high impedance surface is interesting and challenging. An AMC is suggested with loops loaded with non-Foster elements on them. The structure has a spacer layer only without a FSS layer, unlike Sievenpiper’s structure. The spacer only structure is much simpler even though it does not suppress the propagation of surface wave mode and it is sensitive with regard to non-Foster elements.

The structure is also compared to a FSS layer only model. Regarding the sensitivity, a FSS layer is expected to reduce the sensitivity. The structure can be implemented to combine the spacer layer and the FSS layer together to make it less sensitive.

When a negative impedance converter with non-Foster elements is available for the frequency of interest, it will provide the $L_{neg}$ and $C_{neg}$ for the AMM loops or FSS layers. This will enable a low-profile UHF antenna less than $\lambda/4$ with a high impedance ground surface.
REFERENCE


APPENDIX A
IMPEDANCE CALCULATION OF A BI-UNIAXIAL STRUCTURE
FOR TM MODE
The impedance of a two-layer bi-uniaxial structure is calculated for the TM mode. Each layer can be modeled as a transmission line. The entire configuration is shorted at the one end as shown in Figure A-1.

![Figure A-1: Equivalent transmission line representation of two layer bi-uniaxial structure.](image)

For the bi-uniaxial structure, the material characteristics of both layers are applied to Equation (A.6) - (A.8).

The second layer is corresponds to the capacitance instead of a FSS as shown in Figure 1-3. The admittance of this layer is designated by $Y_2$.

\[
\varepsilon_{2t} = \varepsilon_{2x} = \varepsilon_{2y} = \frac{Y_2}{j\omega\varepsilon_0 t}, \varepsilon_{2n} = \varepsilon_{2z} = 1. \tag{A.1}
\]

\[
\mu_{2t} = \mu_{2x} = \mu_{2y} = 1, \varepsilon_{2n} = \varepsilon_{2z} = \frac{Z_2}{j\omega\mu_0 t}. \tag{A.2}
\]

The first layer is a periodic structure where a unit cell has a via. It is a spacer region which has the medium properties derived in Section 2.4. The material properties are calculated as a Brown’s rodlike medium shown in [12].
\[ \varepsilon_{1t} = \varepsilon_{1x} = \varepsilon_{1y} = \varepsilon_{r, \text{spacer}} \frac{1 - \alpha}{1 + \alpha'}, \quad (A.3) \]

\[ \varepsilon_{1n} = \varepsilon_{1z} = \varepsilon_{r, \text{spacer}} - \frac{\varepsilon^2}{\omega^2 \mu_{r, \text{spacer}} \frac{4\pi}{a^2} (\alpha - l\ln \alpha - 1)}, \quad (A.4) \]

\[ \mu_{1x} = \mu_{1y} = \frac{\varepsilon_{r, \text{spacer}}}{\varepsilon_{2n}} \mu_{r, \text{spacer}}, \text{ and } \mu_{1z} = (1 - \alpha) \mu_{r, \text{spacer}}. \quad (A.5) \]

where

\[ \alpha = \frac{\pi r_{\text{via}}^2}{a^2}, \]

\[ r_{\text{via}} = \text{radius of via.} \]

And \( a = \text{length of side.} \)

From these material properties, each of admittance can be calculated. Here, TM wave propagation is assumed, and the incident wave impinges at an angle of \( \theta \) with respect to the normal direction.
\[ Y_1 = j\omega \frac{\varepsilon_0 \varepsilon_{1t}}{\gamma_1 \cos \theta_1}, \] \hspace{1cm} (A.6)

\[ Y_2 = j\omega \frac{\varepsilon_0 \varepsilon_{2t}}{\gamma_2 \cos \theta_2}, \] \hspace{1cm} (A.7)

\[ Y_3 = j\omega \frac{\varepsilon_0}{\gamma_0 \cos \theta_{\text{inc}}}. \] \hspace{1cm} (A.8)

Where

\[ \theta_{\text{inc}} = \text{incident angle}, \]

\[ \varepsilon_t, \varepsilon_n = \text{transverse and normal directional permittivity}, \]

\[ \mu_t, \mu_n = \text{transverse and normal directional permeability}, \]

\[ \theta_1 = \cot^{-1}\left( \frac{\varepsilon_{1t} \mu_{1t}}{\sin^2 \theta_{\text{inc}} - \varepsilon_{1t} \varepsilon_{1n}} \right), \]

\[ \theta_2 = \cot^{-1}\left( \frac{\varepsilon_{2t} \mu_{2t}}{\sin^2 \theta_{\text{inc}} - \varepsilon_{2t} \varepsilon_{2n}} \right), \]

\[ Y_1 = \sqrt{-k_0^2 \frac{\varepsilon_{1t} \mu_{1t} \varepsilon_{1n}}{\varepsilon_{1t} \sin^2 \theta_1 + \varepsilon_{1n} \cos^2 \theta_1}}, \]

\[ Y_2 = \sqrt{-k_0^2 \frac{\varepsilon_{2t} \mu_{2t} \varepsilon_{2n}}{\varepsilon_{2t} \sin^2 \theta_2 + \varepsilon_{2n} \cos^2 \theta_2}}. \]