SANDIA REPORT

SAND2011-6158

Unlimited Release

Printed September 2011

Thin Magnetic Conductor Substrate for Placement-Immune, Electrically-Small Antennas

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Abstract

An antenna is considered to be placement-immune when the antenna operates effectively regardless of where it is placed. By building antennas on magnetic conductor materials, the radiated fields will be positively reinforced in the desired radiation direction instead of being negatively affected by the environment. Although this idea has been discussed thoroughly in theoretical research, the difficulty in building thin magnetic conductor materials necessary for in-phase field reflections prevents this technology from becoming more widespread.

This project's purpose is to build and measure an electrically-small antenna on a new type of non-metallic, thin magnetic conductor. This problem has not been previously addressed because non-metallic, thin magnetic conductor materials have not yet been discovered.

Acknowledgment

Special thanks go to Terry Garino for his help manufacturing many ceramic cylinders for the creation of the artificial magnetic conductor substrate. Additional thanks go to Ben Hanks for his help cutting Makrolon sheets to the appropriate thicknesses and lengths to create parallel-plate waveguide calibration kits.

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Preface

This project was sponsored as a high-risk, high-reward Sandia Labs LDRD because the magnetic conductor used for the antenna's substrate has to be carefully designed to suit the antenna's frequency dependent needs. This material is not available commercially, but similar materials have been artificially created with internal metal components. The thinness of current magnetic conductor technology is limited by the presence of metal inside a substrate. By removing this metal and redesigning the magnetic conductor's resonance with particle geometry, the magnetic conductor can potentially become very thin and flexible.

By combining new research on magnetic conductor materials with electrically-small antenna design, this project can theoretically enable rapid placement of remote antennas with potential for gain improvement over existing systems.

Nomenclature

ADS Advanced Design System

AMC artificial magnetic conductor

CST Computer Simulation Technology

GHz Giga-Hertz

kHz kilo-Hertz

LDRD laboratory directed research and development

MHz Mega-Hertz

PDMS Polydimethylsiloxane

PMC perfect magnetic conductor

PPWG Parallel-plate waveguide

TRL through-reflect-line

UHF ultra-high frequency

Chapter 1

Introduction

Since 1999, Sievenpiper style high-impedance surfaces have been used to artificially create magnetic conductors at specific frequencies by making an open circuit boundary condition on the surface of a substrate for incoming electromagnetic waves [1]. This type of artificial magnetic conductor (AMC) improves upon previous AMCs by reducing the thickness of the material from a quarter wavelength to much less than a wavelength while maintaining a zero-phase reflection coefficient [2]. The Sievenpiper AMC requires via holes and capacitive patches above a ground plane to act as resonant inductor-capacitor circuits, and its thinness is limited by the interior metal required to create these resonant circuits. Other types of AMCs have been produced by embedding curved metal strips into dielectrics such that the metal strips create resonant inductor-capacitor circuits [3]. Simulations on this type of thick AMC showed that in-phase reflection from the surface occurred at a frequency where the substrate permeability was very high and the substrate permittivity was near zero. This caused the wave impedance of the substrate to be real and very large, like an open circuit.

This work proposed the creation of an AMC with in-phase field reflections without using internal electric conductors, the placement of an electrically-small antenna on this magnetic conductor, and the development of a transmit-receive system that utilizes the substrate and electrically-small antenna. By not using internal electric conductors to create the AMC, the substrate thickness can be minimized. The electrically-small antenna will demonstrate the substrate's ability to make an antenna placement immune, and the transmit-receive system combines both the antenna and the substrate while adding a third layer of system complexity to demonstrate the complete idea.

The components internal to the dielectric substrate will be designed in CST Microwave Studio® and analyzed by passing plane waves through them at various frequencies. The electric and magnetic dipole moments originate from the resonances inside the substrate building blocks. For a dielectric rod whose length is aligned with the plane wave's electric field vector, the electric dipole moment's resonant frequency corresponds to the length of the rod, whereas the resonant frequency of the magnetic dipole moment corresponds to the circumference of the rod. If the length and circumference both have dipole resonance at the same frequency, a substrate made of these rods can exhibit double-negative material properties at a fixed range of frequencies.

After designing and fabricating the AMC, an electrically-small antenna will be designed in the same software and built on the substrate to demonstrate in-phase reflection for placement-immune antenna operations. The antenna will be excited with non-uniform waves that normally

feed antennas instead of the ideal plane waves used to excite the magnetic conductor substrate simulations. Examples of typical antenna feeds include coaxial transmission lines, microstrip lines, and strip lines. The fields inside these feeds will propagate down the antenna's transmission lines to excite the electrically-small antenna, which will then either confirm or reject the notion that the substrate acts as a magnetic conductor at the antenna's operating frequency by showing the in-phase or out-of-phase field reflections from the magnetic conductor. The properly operating magnetic conductor will reflect the antenna's fields with constructive interference when the antenna radiates into the magnetic conductor.

After putting a placement-immune, electrically-small antenna on a minimally thin AMC, a simple transmit and receive system will be designed in ADS and constructed to demonstrate system-level applications. This transmit-receive system will at least include an electrically-small receive antenna (which may also act as the transmit antenna), an RF amplifier, and either a circulator or an RF switch to isolate the transmit and receive signals. Since this task is auxiliary to the demonstration of the placement-immune antenna on the AMC, it may be sacrificed if necessary to provide more time and funds for the measurement of the AMC. Assuming that there is enough time and money left over after the fabrication and measurement of the AMC and the electrically small antenna, the transmit and receive system will be tested for maximum communication range in free space and on various inanimate objects such as a car, a briefcase, and a computer. Although these range measurements are unnecessary to prove the proper operation of the substrate and the antenna, they may provide helpful information for customers interested in purchasing this technology.

The largest technical challenge in this work lies in creating an AMC without using electric conductors to setup resonant inductor-capacitor circuits in the substrate. If the AMC can be created without electrical conductors embedded into it, proving the placement-immunity of an electrically-small antenna and developing a transmit-receive system for the antenna on this material should be fairly straight-forward. The functionality of the properly working magnetic conductor substrate will be proven by comparing the measured radiated fields of the electrically-small antenna on a non-magnetically conducting substrate with the radiated fields of the electrically-small antenna on the AMC substrate. This test will demonstrate in-phase reflections from the AMC's surface by showing positive gain enhancement in a single hemisphere. To demonstrate the placement-immunity of the antenna, the return loss of the antenna over the AMC will be compared when the assembly is located in free-space, near water, and near metal. This test will show that the antenna is not de-tuned by its local environment.

Chapter 2

Design

The non-metallic AMC substrate design originated as a derivative of [4], in which Jacques Loui demonstrated that spheres with moderately high dielectric constants ($\epsilon_r = 20$ to $\epsilon_r = 38$) can exhibit μ and ϵ resonances near 17 GHz when placed in cubic lattices. By choosing the sphere sizes to have magnetic dipole and electric dipole resonances near the same frequency, Loui was able to artificially make a material with negative μ and negative ϵ values at the same frequency.

In order to create an AMC, one may excite the magnetic dipole moment of a material with a special particle geometry, creating a permeability resonance that will have a large μ value over a limited bandwidth. As the μ value approaches infinity while the ϵ value remains close to unity, the wave impedance of the substrate will approach positive infinity, as shown in Eq. 2.1 [5], where Z_{Sub} is the wave impedance of the substrate, μ is the permeability of the substrate, and ϵ is the permittivity of the substrate. As the wave impedance approaches positive infinity, the reflection coefficient of the substrate approaches positive unity, as shown in Eq. 2.2 [6], where Γ_{Sub} is the reflection coefficient of the substrate, Z_{Sub} is the wave impedance of the substrate, and Z_{Air} is the wave impedance of air. When the substrate has a reflection coefficient of positive unity at a certain frequency, all fields at that frequency will reflect from the substrate's surface with no phase shift, reinforcing the radiated fields from nearby antennas that are tuned to the same resonant frequency as the substrate.

$$Z_{Sub} = \sqrt{\frac{\mu}{\epsilon}} \quad (2.1)$$

$$\Gamma_{Sub} = \frac{Z_{Sub} - Z_{Air}}{Z_{Sub} + Z_{Air}} \quad (2.2)$$

An alternative method to create in-phase field reflections that does not rely on particle geometry uses a quarter-wavelength short, which operates effectively over a limited bandwidth. If a substrate's thickness equals a quarter of its guided wavelength, then the fields entering the substrate's surface at perpendicular incidence will be reflected with the same phase from that surface. This principle is governed by the input impedance equation, shown in Eq. 2.3 [6], where Z_{in} is the input impedance looking from air into the substrate, Z_{SC} is the substrate's characteristic impedance, Z_{Short} is the load impedance of the metal backing on the substrate (approximately 0Ω), β is the

propagation constant, and l is the thickness of the substrate.

$$Z_{in} = Z_{SC} \frac{Z_{Short} + jZ_{SC} \tan(\beta l)}{Z_{SC} + jZ_{Short} \tan(\beta l)} \quad (2.3)$$

For a quarter-wave thick substrate, the incident fields upon the substrate see a Z_{in} equivalent to an open circuit, which effectively provides the same in-phase field reflection as having a wave impedance that approaches infinity. If the input impedance looking into the substrate approaches infinity, then substituting that value into Z_{Sub} in Eq. 2.2 shows that the reflection coefficient at the surface of the substrate would approach positive unity.

Particle Geometry Resonance Configuration

The first substrate design investigated by the authors used particle geometry to create in-phase field reflections at the surface of the substrate by tuning the magnetic dipole moments of high dielectric constant ceramics to resonate at the European ISM band (433 MHz). The unit cell of this design is shown in Fig. 2.1, in which the green cylinder is a ceramic with a dielectric constant of 3600, and the translucent surrounding material is FR-4 with a dielectric constant of 4.3. The green cylinder has a diameter of 8 mm and a height of 13 mm. This cylinder is centered inside the translucent rectangular prism, which has the dimensions 26 x 8 x 8 mm.

Figure 2.1. Unit cell of the particle geometry based substrate.

Transmission and Reflection

Using a Floquet mode simulation in CST with the cylinder in Fig. 2.1 as the unit cell reference, the simulated transmission and reflection S-parameters were observed at the surface of the substrate (Figs. 2.2-2.3). For a substrate made of a planar array of these cylinders, the substrate should ideally exhibit complete reflection at its resonant frequency with in-phase reflection also occurring at that frequency.

Figure 2.2. Simulated transmission and reflection power levels for the substrate based on particle geometry resonance.

Figure 2.3. Simulated reflection phase for the substrate based on particle geometry resonance.

For a substrate built with the structural dimensions of the cylinder in Fig. 2.1, the magnetic permeability resonance occurs at 433 MHz (as demonstrated in Fig. 2.2). The transmission approaches negative infinity dB and the reflection approaches zero dB at 433 MHz because the μ value becomes very large at that frequency, making the wave impedance of the substrate approach infinity Ω (Eq. 2.1). As shown in Fig. 2.2, the transmission power level through the substrate at 433 MHz is 30.0 dB less than the power input to the surface of the substrate, meaning that 0.1% of the input power was transmitted through the substrate. The reflected power level at the same

frequency was only 0.08 dB less than the power input to the substrate's surface, which means that 98.2% of the input energy was reflected. The remaining 1.7% of the energy not reflected or transmitted was either absorbed by the substrate or radiated out of the sides of the substrate. The reflection efficiency of this substrate over its $\pm 90^\circ$ bandwidth varies from 0.01% at 426 MHz to 98.2% at 433 MHz. At the highest frequency in the $\pm 90^\circ$ bandwidth (438 MHz), the substrate's reflection efficiency is 42.7%.

If the phase of the reflection is near 0° at the frequency where the transmission approaches negative infinity dB, then the substrate looks like a perfect magnetic conductor. Since the zero phase reflection occurs at 431.5 MHz (not 433 MHz), the substrate is not a perfect magnetic conductor. However, the in-phase reflection has a $\pm 90^\circ$ bandwidth from 426-438 MHz (as shown in Fig. 2.3), and this bandwidth covers the frequency of the substrate's resonance, which means that the substrate would provide reflections that introduce positive reinforcement into the original signal. Practically, the authors found that the phase of the reflection coefficient could be tuned by modifying the lengths of the cylinder and its unit cell box so that the in-phase reflection frequency would align with the total reflection frequency. With further tuning, it may be possible to perfectly align the zero phase reflection frequency with the magnetic resonance frequency.

Figure 2.4. Linear antenna aligned correctly over the particle geometry based substrate.

E-Field Polarization

In order to excite the resonance of the particle geometry based substrate, only one electric field polarization may be used. The electric field vector must be aligned in parallel with the height of the

high dielectric constant cylinder, or the substrate will not provide positively reinforcing reflections. Since the high dielectric constant cylinders must see a specific electric field polarization, only linear antennas with the correct alignment angle over the substrate would see positive gain reinforcement. Additionally, if either an antenna with non-linear polarization or a mis-aligned antenna with linear polarization were placed over this type of substrate, the antenna would definitely not be placement-immune because the radiated fields would not be reflected by the substrate. An example of a linear antenna with the correct alignment over the particle geometry based substrate is shown in Fig. 2.4.

Surface Wave Suppression

Placement-immunity can be defeated by the polarization and alignment requirements for an antenna over this substrate, and it can also be defeated by poor surface wave suppression. A substrate that propagates surface waves will see field reflections at its boundaries that vary with the environment. These boundary field reflections would inevitably radiate or reflect back to the antenna and alter its performance, thereby making the antenna placement-sensitive to some degree. The simulated surface wave suppression of this substrate is shown for two orthogonal electric field vectors in Fig. 2.5. The electric field propagation direction in Fig. 2.5 is indicated by the vector k .

For the electric field polarization parallel to the height of the cylinders (E_1), a simulated wave propagating perpendicularly to the electric field through the width of the substrate (along k) sees 5.5 dB of insertion loss at 433 MHz after passing through six cylindrical units. Realistically, this means that if an antenna is placed on a substrate with a width of six cylindrical unit cells, greater

Figure 2.5. Simulated surface wave suppression through six of the particle geometry resonant unit cells for two orthogonal electric field vectors.

than or equal to 28% of the horizontally propagating energy will reach the substrate's boundary, at which it will either reflect or radiate depending on the local environment. For an electric field parallel to E_2 that is propagating in the direction of k in Fig. 2.5, the surface wave suppression at 433 MHz through six cylindrical units is only 3.4 dB, which corresponds to 46% of the horizontally propagating energy being unimpeded. Although the E_1 electric field polarization in Fig. 2.5 has good surface wave suppression at 429 MHz, only 6% of the input energy is reflected from the substrate at that frequency (Fig. 2.2), meaning that up to 94% of the input energy could be lost. Since this type of substrate has poor surface wave suppression, the authors decided that it was not the best design to pursue in fabrication and measurement.

Quarter Wave Reflection Configuration

Since the particle geometry based AMC exhibited poor surface wave suppression in simulation, an alternative AMC was designed with high surface wave suppression as its basis. Instead of using the magnetic resonance of high dielectric constant ceramic rods to reflect energy with zero phase change at the surface of a substrate, this second design uses the magnetic resonance of high dielectric constant ceramic rods to suppress surface waves inside the substrate, while maintaining zero phase change reflections through the quarter-wavelength short technique described earlier in this chapter.

Figure 2.6. Unit cell of the quarter-wave reflection based substrate.

The unit cell of this design is shown in Fig. 2.6, in which the green cylinder has a dielectric constant of 3600 and the surrounding translucent material is Plexiglas with a dielectric constant of 3.6. The green cylinder has a diameter of 9 mm and a height of 5 mm, and it is centered inside the translucent rectangular prism, which has the dimensions 9.5 x 9.5 x 5 mm. The unit cell is backed by a perfect electric conductor sheet to provide a quarter-wavelength short. Compared to the particle geometry based substrate's unit cell, the quarter wave reflection substrate's unit cell is 73% smaller by volume.

Transmission and Reflection

Since the unit cell for this design is backed by perfect electric conductor, there is no transmission through the substrate ($-\infty$ dB). Any energy not reflected by the substrate is either absorbed or

Figure 2.7. Simulated reflection power level for the quarter-wavelength based substrate.

Figure 2.8. Simulated reflection phase for the quarter-wavelength based substrate.

radiated out of the sides of the substrate. The simulated reflection power levels for a Floquet mode CST solution of this unit cell are shown in Fig. 2.7. Between 1 to 2.5 dB of input power is lost inside the substrate over the $\pm 90^\circ$ bandwidth of the substrate, which means that this substrate's reflection efficiency lies between 56-79%. The $\pm 90^\circ$ bandwidth of the substrate stretches from 428 to 438.5 MHz, and it is shown in Fig. 2.8. At 433.5 MHz the phase of the reflection is equal to zero degrees.

Unlike the particle geometry based substrate design, this quarter-wavelength based substrate design reflects a majority of the input energy at all frequencies (Fig. 2.7). Therefore only the zero-degree reflection phase needs to be tuned to a desired frequency to have an AMC at that frequency. Knowing that the zero-degree reflection frequency is determined by the quarter-wavelength short in the substrate, the designer can adjust the height and the effective dielectric constant of the substrate to tune the AMC to any frequency. For the particle geometry based substrate design, both the zero-degree reflection phase and the zero dB reflection power level have to be tuned to the same frequency in order to have an AMC, making the design of that substrate more time consuming.

Figure 2.9. Linear antenna over the quarter-wavelength based substrate.

E-Field Polarization

The quarter-wavelength reflecting substrate does not require a specific electric field polarization to create in-phase reflections at the surface of the substrate due to the way that the quarter-wavelength short appears as an open at the substrate's surface. Therefore any antenna polarization may be used over this type of AMC, providing great flexibility for the applications of this design. An example of a linear antenna over this substrate is shown in Fig. 2.9.

Surface Wave Suppression

Because of the magnetic resonance of the high dielectric constant ceramics, this substrate has excellent surface wave suppression for electric fields lying parallel to the substrate's surface. The surface wave suppression through five unit cells is shown in Fig. 2.10 for the fields parallel and perpendicular to the substrate's surface, and the vector k denotes the propagation direction of those fields. After passing through five unit cells, the electric fields parallel to the substrate's surface (E_1) are suppressed by 90.7 dB at 433 MHz, which corresponds to only $8.5 \cdot 10^{-8}\%$ of the input power propagating all the way through the unit cells. The fields perpendicular to the surface of the substrate (E_2) are suppressed by 1.6 dB at the same frequency, which corresponds to 69.2% transmission of the input power. Strong similarities exist between the suppression of E_2 in the particle geometry based substrate and the suppression of E_2 in the quarter-wavelength based substrate (compare E_2 in Fig. 2.5 to E_2 in Fig. 2.10).

Due to the dual-axis symmetry of the unit cell for this substrate, the substrate has symmetrical surface wave suppression performance in four directions away from the substrate's center. With high probability, the waves traveling at oblique angles would also see high attenuation as they pass through several resonant unit cells. The propagating electric field through the substrate will originate from the propagating electric field on the antenna, and therefore the electric field in the substrate will have the same polarization as the electric field on the antenna. Since the electric field on a planar antenna lies co-planar to its surface, that field also lies co-planar with E_1 in Fig. 2.10. Because of the dual-axis symmetry of the unit cell, the surface waves excited by the antenna over the substrate will be heavily suppressed in at least four directions. Because of this design's

Figure 2.10. Simulated surface wave suppression through five of the quarter-wavelength based unit cells for two orthogonal electric field vectors.

excellent simulated surface wave suppression in combination with its high reflection efficiency over its $\pm 90^\circ$ bandwidth (56-79%), this design was chosen for fabrication and measurement.

Lattice Structures

Using the high dielectric constant cylinder from the quarter-wavelength based substrate design, two different lattices were designed to house the AMC substrates. Originally the background material for the lattice structures was intended to be Plexiglas. Because Plexiglas is more difficult to cut than Makrolon, and because it exhibits nearly the same electrical properties, Makrolon was chosen for the background material of the lattices. Makrolon has a dielectric constant near 2.8 at 433 MHz.

When a sintering furnace bakes ceramics, the ceramics shrink in size. Since the unit cell cylinder has a diameter of 9 mm and the original high dielectric constant material purchased (HF-402) was known to shrink about 18% in the sintering furnace, a dye press with a diameter of 10.9 mm was needed to create the same size cylinder that was modeled in simulations. However, the dielectric constant of the HF-402 material was unknown at 433 MHz, so the authors decided that purchasing a custom dye press would not be more beneficial than using an available 12.7 mm diameter dye press, especially since a larger diameter cylinder simply lowers the in-phase reflection frequency in simulated results. After pressing and sintering the HF-402 ceramic cylinders, they have a 10.5 mm diameter. The initial spacing between unit cell cylinders from the simulations (0.5 mm) was kept, and the lattice sizes were stretched to adapt to the larger diameter cylinders.

Elastomer Option

For measurement purposes, it is helpful to have the AMC inside of a rigid lattice structure to maintain a constant measurement face, but once the AMC is characterized in the Makrolon structure it could be built in non-rigid materials to adapt the AMC to many environments. Instead of building the AMC inside of rigid Makrolon, the AMC could be built inside of Polydimethylsiloxane (PDMS), which is capable of conforming to non-planar surfaces. PDMS is inert, non-toxic, non-flammable, and it has a dielectric constant between 2.3-2.8.

Square Lattice

The first lattice design extends the original unit cell into a 10 x 10 square array, which is a similar unit cell expansion to the Floquet mode simulations performed in CST that were presented previously. The 100 cell lattice has the dimensions 121 x 121 x 6 mm. This size is wide enough to hold a 433 MHz quarter-wavelength monopole antenna above a Rogers 4003 substrate ($\epsilon_r = 3.55$). At 433 MHz, a quarter-wavelength in Rogers 4003 material is equal to 92 mm. It is possible that the measured in-phase reflection frequency of the substrate may be lower than the expected 433 MHz if the dielectric constant of the ceramic cylinders is higher than expected. In this case, it may

Figure 2.11. Fabricated 100 cell Makrolon square lattice.

Figure 2.12. Fabricated 101 cell Makrolon hexagonal lattice with HF-402 ceramic cylinders, copper ground, and Makrolon covers included.

be necessary to use meandering techniques to fit a lower frequency antenna onto the substrate. By using meandering techniques, it is possible to reduce the length of an antenna in one dimension

by increasing its electrical length in other dimensions. The fabricated 100 cell Makrolon lattice is shown in Fig. 2.11.

Hexagonal Lattice

The second lattice design rotates the unit cell around a hexagon and expands that hexagonal structure into a nearly square structure that holds 101 cylinders. By expanding the unit cell into a hexagonal array, an additional axis of symmetry is established, which provides excellent surface wave suppression in at least six directions from the substrate's center. The hexagonal array also gives the ceramic cylinders tighter coupling that will increase the effective dielectric constant of the substrate, hypothetically decreasing the substrate's in-phase reflection frequency. A copper sheet grounds the ceramic cylinders, and two 0.030" Makrolon sheets act as covers to prevent the ceramic cylinders from falling out of the Makrolon lattice. The 101 cell hexagonal lattice has the same dimensions as the 100 cell square lattice (121 x 121 x 6 mm), and it is shown in Fig. 2.12.

High Dielectric Constant Ceramics

Since high dielectric constant ceramics ($\epsilon_r \approx 3600$) are normally used to manufacture capacitors, the material manufacturers producing these capacitors do not usually have frequency dependent dielectric constant data for their materials. Instead, these manufacturers provide the dielectric constants for their materials only at the source frequency of their capacitance meters (usually 1 kHz or 1 MHz). Because of the lack of information available on high dielectric constant materials' properties at UHF frequencies, the authors have to purchase the materials and measure their dielectric constants using two-port S-parameter techniques to be certain that the materials have the desired dielectric constants at UHF frequencies.

MRA Labs HF-402

The authors first investigated the HF-402 high dielectric constant material created by MRA Laboratories, Inc. This material has a manufacturer stated dielectric constant of 3900 ± 300 at 1 kHz, is composed of mostly Barium Titanate, and is RoHS compliant. MRA Labs ships this material as a powder in a minimum quantity of 2 kg. For proper powder binding, MRA Labs recommends soaking the powder in a solution of 30% polyvinyl alcohol and 70% water prior to pressing the powder. After the powder has been pressed, it should be baked and sintered according to the temperature profiles provided by MRA Labs.

After soaking the HF-402 powder in the binder solution, the powder was ground into finer particles to prepare it for pressing. For each of the 101 ceramic cylinders that were created, three grams of the fine powder (with binder included) were weighed on a balance and then inserted into the 12.7 mm diameter dye press. The weighed powder was then pressed at 1000 psi and removed

from the dye press for baking and sintering. When the cylinders were removed from the sintering furnace, they had 10.5 mm diameters and 6.2 mm heights ± 0.1 mm in both dimensions.

DiLabs BL, BJ and BN

After examining the HF-402 material from MRA Laboratories, the authors investigated three materials created by Dielectric Laboratories (DiLabs). These materials were appealing to the authors because they have high dielectric constants measured by the manufacturer at 1 MHz, as opposed to the 1 kHz dielectric constant measurements performed by MRA Labs. The BL, BJ, and BN materials have stated dielectric constants of 2000, 3300, and 4500 (respectively) with a ± 10 to 15% tolerance on those values. Since DiLabs measures the dielectric constant of their materials at a frequency a factor of 1000 greater than the measurements performed by MRA Labs, the authors are more confident that the materials will have similar dielectric constants to their stated values when measured at 433 MHz.

One interesting property of the DiLabs materials worth noting is that they age over time, and the dielectric constant of these materials decreases similarly to an inverse time function. The aging of the BL and BJ materials is shown in Fig. 2.13, and the aging of the BJ material is shown in Fig. 2.14. Out of these three materials, BJ has the greatest stability over time with only a little over 3% variance 11 years after it was sintered. BN has the worst temporal stability with a 15% variance over the same time period. Extrapolating past the boundaries of the plotted data would reveal that

Figure 2.13. Dielectric constant aging characterized as capacitance change for the BL and BJ materials.

Figure 2.14. Dielectric constant aging characterized as capacitance change for the BN material.

the variance of the dielectric constants in these materials is increasingly negligible. Between 11 to 111 years after the ceramics are sintered, BN's dielectric constant will vary by an additional 2%, BL's dielectric constant will vary by an additional 1%, and BJ's dielectric constant will vary by an additional 0.5%.

In order to protect its intellectual property DiLabs does not ship these materials as powders to its customers. Instead, DiLabs presses and sinters the materials prior to shipping, which could be either a convenience or a detriment to the ceramic designer. Since DiLabs does its own ceramic pressing, only a few shapes and sizes of presses are available, limiting the ceramic geometries that can be created without having to purchase custom presses for DiLabs to use. However, DiLabs has perfected their pressing and sintering technique for their custom powders, so less experimentation time is required by the ceramic designer to manufacture the ceramics correctly without cracks and other asymmetrical material properties.

DiLabs uses a machine assembly process that mixes a liquid binder into the high dielectric constant powder and rolls the mixture into 30 mil thick sheets prior to punching out shapes that fit into their dye presses. Because of this process, it is difficult for DiLabs to make ceramics thicker than 30 mils. Therefore, the authors requested from DiLabs many 30 mil thick cylinders with approximately the same radius as the cylinders made out of HF-402 in order to stack the 30 mil cylinders into the previously mentioned Makrolon lattices.

Substrate Verification

There are two main methods by which the artificial magnetic conductor's resonance may be observed. The first and most ideal method is a quantitative measurement of the reflection coefficient from the surface of the artificial magnetic conductor. In order to measure the reflection coefficient at the surface of the artificial magnetic conductor, an accurate calibration kit must de-embed the S-parameters from the desired measurement surface. The second method is a qualitative measurement of the gain of an antenna placed near the magnetic conductor's surface. If the gain improves when the antenna lies near the surface of the magnetic conductor, then the magnetic conductor is reflecting the antenna's fields with positive reinforcement. Additionally, if the antenna gain is absorbed by the surface, then the substrate does not act as an artificial magnetic conductor, but it may still be able to provide placement immunity for antennas near its surface. Simulations have shown that if the loss tangent of the high dielectric constant materials is greater than or equal to 2%, then the substrate will act more like an absorber than an artificial magnetic conductor.

Parallel-Plate Waveguide Calibration Kit for Quantitative Measurement

In order to de-embed the phase of the reflection coefficient from the surface of the substrate, a parallel-plate waveguide (PPWG) calibration kit was designed and fabricated based on a one port triple short technique. One of the benefits of the PPWG over traditional rectangular waveguide is that the PPWG has a scalable size based on the ratio between its height and width, whereas rectangular waveguide has a fixed size for a certain frequency range. For example, rectangular waveguide at 433 MHz would require a cavity with dimensions 533 x 267 mm, but the PPWG was designed to have a height similar to the diameter of the HF-402 pucks and a width that gave the PPWG a 50 Ω characteristic impedance (12.5 x 50 mm). Therefore the measurement surface required by the PPWG is much smaller (and less expensive) than the surface required by the rectangular waveguide.

Two separate transitions from coaxial cable to PPWG were designed according to [7] with center frequencies near 350 and 550 MHz. The simulated 3 dB insertion loss bandwidth of two back-to-back PPWG transitions is 277-461 MHz for the lower frequency transition and 251-877 MHz for the higher frequency transition. The topology of two back-to-back transitions is shown in Fig. 2.15.

Figure 2.15. Topology of two back-to-back PPWG transitions.

When the PPWG calibration kit was calibrated to the HF-402 based substrate's measurement surface using the triple short technique, a resonance phenomenon inside the PPWG made the constructed substrate appear to resonate at 550 MHz, as shown in Fig. 2.16. Since the authors doubted the accuracy of the phase measurement shown in Fig. 2.16, they decided to modify the triple short calibration kit into a through-reflect-line (TRL) calibration kit to measure the two-port S-parameters of transmissions through the HF-402 material. This allowed them to retrieve the dielectric constant of that material at 433 MHz (as shown in the next chapter) to determine whether or not the resonance they were previously measuring should exist based on simulated models.

Figure 2.16. Resonance at 550 MHz that appears with the triple short calibration.

UHF Antenna for Qualitative Measurement

In addition to the PPWG verification method above, measuring antenna performance over the AMC substrate provides an alternate test of substrate performance. A meanderline dipole antenna operating at 550 MHz was designed, fabricated, and tested to prove the antenna concept. An operating frequency of 550 MHz was chosen based on the expected in-phase reflection frequency, taken from initial measurements of the AMC substrate. Once the final substrate has been built and validated, the antenna frequency can be adjusted with minimal effort.

A double-sided parallel-strip line (DPSL) feed was chosen for integration with printed circuit techniques and easy transition to coaxial transmission line. Fig. 2.17 shows the antenna geometry and the simulated and measured return loss. Fig. 2.18 contains the simulated antenna gain in free space and over a perfect magnetic conductor (PMC). The measured and simulated results agree, and the simulated gain indicates an approximate 3 dB gain increase over the PMC, as expected. The

Figure 2.17. Meanderline dipole antenna geometry and return loss.

Figure 2.18. Simulated antenna gain in free space and over a PMC.

antenna pattern remains omni-directional over a hemisphere when placed over a perfect magnetic conductor.

The meanderline dipole antenna was designed to stay matched in both free space and over a PMC. The simulated return loss of the antenna over a PMC is shown in Fig. 2.19, and the measured return loss of the antenna over the fabricated substrate is compared to the ideal antenna performance in the same figure. The fabricated substrate is not a PMC due to its narrow-band in-phase reflection with fields that penetrate the substrate. By contrast, a PMC has broadband in-phase reflection, and all reflection occurs at the initial interface. Simulation of even a simple dipole over the AMC substrate indicates that the antenna-substrate interaction is significantly more complex than initially assumed. Simply simulating the antenna over a PMC is not sufficient to predict antenna performance when installed over the AMC substrate. The authors are currently working on understanding and compensating for the antenna-substrate interaction, while continuing to measure and verify the substrate's performance.

Figure 2.19. Simulated return loss of the antenna over a PMC versus the measured return loss of the antenna over the fabricated AMC.

As shown in the next chapter, the dielectric constant of the HF-402 material appears to be much lower at 433 MHz than it is at 1 kHz, which would cause the designed antenna to appear shorted out by the ground plane behind the AMC as shown in Fig. 2.19. Additionally, preliminary gain measurements of the antenna over the AMC were much lower than the measured gain of the antenna in free space, which also corresponds well with the hypothesis that the ground plane behind the AMC is reflecting out-of-phase fields back to the antenna.

Chapter 3

Dielectric Constant Measurement

As the previous chapter demonstrates, the qualitative and quantitative measurements of the HF-402 substrate provided conflicting results. Furthermore, the number of unknowns involved in the quantitative measurements including material properties, coax-PPWG transition resonances, and the influence of the substrate geometry on the PPWG measurement surface prevent accurate characterization of the problem. To address this issue, the PPWG calibration structures were adapted for two-port TRL calibration in order to directly measure bulk material properties. Once the material has been well characterized, substrate design and verification can proceed with confidence.

Based on [8, 9, 10, 11], a material extraction algorithm was created in MATLAB and ap-

Figure 3.1. Custom algorithm extracted material properties for HF-402, TMM10i, and air in PPWG.

plied to two-port PPWG measurements of the MRA Labs HF-402 to allow the authors to directly measure the dielectric constant of the ceramics used in the substrate fabrication over a broad frequency range. Material parameters extracted by the custom algorithm were compared with results extracted from a commercial code provided by Agilent, and both approaches produced similar results. Fig. 3.1 contains permittivity and permeability measurements for a row of 6.15 mm thick HF-402 pucks “PUCK”, a 9.3 mm thick slab of TMM10i “TMM”, and an 11 mm thick air gap. This plotted data was extracted from two-port PPWG measurements with the custom software that Loui designed.

A rough calibration was performed with air and Rogers TMM10i material, which have known dielectric constants. The dielectric constant of air is approximately 1, and the dielectric constant of TMM10i is 9.8. The custom algorithm extracts $\epsilon_r = 1$ for air and $\epsilon_r = 7.7$ for TMM10i, giving a 27% error. The custom algorithm extracts $\epsilon_r = 11$ (with some variation over frequency) for MRA Labs HF-402. Reproducing the measurements in multiple lab environments with different extraction algorithms consistently produced $\epsilon_r = 10$ -14 for MRA Labs HF-402 material. Similar measurements for DiLabs BN material produced $\epsilon_r = 110$ -140, which is a significant improvement over HF-402 but is still significantly lower than the expected value of 4500.

Figure 3.2. PPWG moding and fringing field effects.

Since the extracted dielectric constant was significantly lower than the expected value for both MRA and DiLabs material, the two-port TRL calibration using PPWG was investigated further to determine whether unforeseen interactions were impacting results. Simulation of high dielectric constant blocks inside the PPWG revealed numerous issues. First, the material extraction algorithms calculate the effective dielectric constant of the sample. Since PPWG has substantial fringing fields in the air surrounding the structure, the effective dielectric constant differs substantially from the bulk material dielectric constant. This effect can be seen for Rogers TMM10i, which came out 27% lower than the known value due to fringing fields, and this also explains why TMM10i has 27% error while air is extracted accurately. Furthermore, the 50 mm PPWG width, required for a 50 Ω transmission line, allows standing waves in high permittivity samples in which the sample width is greater than a half-wavelength. Fig. 3.2 shows an example of the transverse resonance and fringing field effects for a PPWG transmission line. The issues described above apply to solid blocks of material, so material extraction becomes even more difficult for rows of pucks inserted inside a PPWG.

Coaxial measurements systems are currently being pursued in light of the obstacles to material parameter extraction with PPWG. A GR-900 kit manufactured by Maury Microwave has been chosen for several reasons. The coaxial transmission line has an outer diameter of 14.2875 mm, which is only 2 mm greater than the diameter of the HF-402 material and 0.8 mm greater than the diameter of the BN material. Additionally, the structure contains all fields inside the coaxial line, removing concerns about fringing fields and effective versus actual dielectric constant measurements. To give confidence in the coaxial measurement, material extraction algorithms have been applied to 3D electromagnetic simulations of high permittivity materials in both coaxial transmission lines and PPWG. Based on the two-port simulated data, the custom extraction algorithm produces bulk permittivities and permeabilities that are significantly different from the user specified values for PPWG measurement models. For coaxial measurement models, agreement between extracted and user specified material parameters is nearly perfect.

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Chapter 4

Conclusion

Simulated models show that high dielectric constant ceramic substrates can produce in-phase reflection at UHF frequencies when formed into the proper sizes and orientations. Additionally, simulated models have shown that high dielectric constant materials with dissipation factors greater than 2% act more like absorbers than in-phase reflectors in the previously discussed geometries. Although several commercial companies claim that their ceramics have high dielectric constants at 1 kHz to 1 MHz frequencies, it is difficult to determine the dielectric constant of those materials at UHF frequencies without developing accurate measurement hardware and extraction software.

Assuming that high dielectric constant materials can be found and properly measured, the authors believe that they will be able to develop either an artificial magnetic conductor or an absorber in the UHF range that is less than a tenth of a wavelength thick in order to provide placement-immunity for UHF antennas in remote locations. The authors also hope to fully understand the interaction of a UHF antenna with this type of substrate to be able to apply a UHF antenna directly to the surface of this substrate.

In the following year of this LDRD, the authors will focus on accurately measuring high dielectric constant materials in the UHF band while developing a low-loss interface between a UHF antenna and the artificial magnetic conductor substrate. These two objectives will allow the authors to build a new substrate that will either absorb or positively reinforce the radiation of an antenna on its surface. If time and money permits, the authors would still like to explore system level applications for this new substrate design.

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