A LIGHTWEIGHT ARTIFICIAL DIELECTRIC FOR MITIGATING UNWANTED REFLECTIONS AND SCATTERING AT THE SOIL-AIR DIELECTRIC BOUNDARY ENCOUNTERED BY GROUND-PENETRATING RADAR

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ABSTRACT

A Lightweight Artificial Dielectric for Mitigating Unwanted Reflections and Scattering at the Soil-Air Dielectric Boundary Encountered by Ground-penetrating Radar

The development of an artificial dielectric is documented. The components of the finished product are 10 mm hollow polypropylene balls and 6 mm conducting beads. A method to separate – re-deploy the materials efficiently to achieve consistency of results is described. The material is placed over rough soil to make it look like smooth soil at the surface when illuminated by one antenna pair of a Focused Array Radar (FAR).

Experiments were conducted at Northeastern University, Boston, MA using an HP 8510 and Agilent 8714ES network analyzers, and Geo-centers, Inc., Newton, MA using one antenna-pair of a Focused Array Radar (FAR) system.

The HP 8510 used a paraffin-filled can as an open-ended waveguide to match artificial dielectric to soil, the Agilent 8714ES used WR284 waveguide to find the permittivity of air, and a 3-1/8 inch E.I.A. coaxial transmission line to find permittivities of numerous materials (air, artificial dielectric of 2 different mixtures, dry sand, dry soil, bag moisture soil). The single antenna pair (send/receive) of a FAR system was used to verify the efficacy of the artificial dielectric under actual-use conditions.

The method used to characterize the real relative permittivity values in the coaxial system is described in detail. This method provided 10% accuracy across 700 to 1300 MHz.
Keywords: land-mine detection, artificial dielectric, ground penetrating radar, focused-array radar (FAR), cross-correlation, correlation coefficient, polypropylene, beads, real relative permittivity.
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1. THE REFLECTION PROBLEM FOR GROUND PENETRATING RADAR AT THE AIR / SOIL BOUNDARY

1.1 Introduction

According to the Landmine Report 2007 [1] there were 5,751 documented landmine casualties in 68 countries during calendar 2006 including 1,367 killed, 4,296 injured and 88 of unknown severity. This is better than the longstanding estimate of 15,000 to 20,000 new landmine casualties per year, worldwide, which serves as an indication of improving conditions against a backdrop of hundreds of thousands killed or injured over the past decades. About eighty per-cent are civilians with 1/3 of that fraction women and children [2]. Landmines are designed to maim, but many die anyway for lack of medical attention [3]. According to another study [4] it will take 450 to 500 years to actually clear all of the landmines that have been placed, at the current de-mining rate. Assuming no new landmines are placed, this means four to five centuries of living in fear, four to five centuries of inaccessibility to wide tracts of land, hampering relief efforts, disrupting infrastructure and indiscriminately destroying breadwinners, livestock and wildlife alike, for those living near landmines in any of the 75 to 80 countries with a landmine problem. It only takes 2 or 3 landmines to render an entire field off-limits. There are currently 300,000 to 400,000 people living with landmine injuries and it has been estimated that there are currently 80,000,000 live landmines waiting for more victims, world-wide. These are some of the reasons that improving the process of finding landmines, which can be manufactured for as little as $3.00 each, is at the top of the list among technical challenges in need of a solution.
The goal of this thesis is to explain how to create a lightweight artificial dielectric and demonstrate that it meets two criteria: simulates soil under radar test conditions by comparing return signals in the time domain turning moderately rough soil into level soil, and matches the real permittivity of soil under microwave laboratory conditions. This goal was planned to be achieved through conducting an energetic search for the materials which would demonstrate these properties, then looking for indications of progressive difficulty to carry out that the material was working as expected. By this is meant trying lots of different ways to obtain this artificial material, then eliminating those which did not pass an initial test. The initial test was easier to conduct than the subsequent tests (open-ended waveguide connected to a network analyzer followed by radar tests at a soil bed, then designing/building/testing a microwave measurement fixture and selecting the proper algorithm to extract the information of interest). The initial search for the materials was guided by theoretical papers, and previously created similar materials. This information was adapted to the task at hand, namely remake an artificial dielectric that would mitigate the ground reflection of a ground penetrating radar.

This thesis is organized as follows, an introduction discusses the need for improving landmine detection, the goal of this thesis and organization. Prior work is briefly described, FAR (Focused Array Radar) defined with some explanations of which difficulties this technique solves or doesn’t solve. After the introduction, the nature of ground reflections are discussed, from the standpoint of an incident unified plane-wave upon a lossless/lossy dielectric boundary. Although the actual radar experiments employed near-fields incident on lossless/lossy dielectric boundaries, this discussion bounds the problem at one end as the radar is operated higher and higher above the soil
surface. This concludes the first chapter. The next chapter describes the actual artificial
dielectric that was selected using partially-paraffin-filled open ended waveguide on a
network analyzer. By doing same or different A/B tests, it could be seen that certain
compositions of artificial dielectric were indeed matching certain soil samples. This
artificial dielectric was then tested under actual radar conditions with a soil bed in the
laboratory in chapter 3. The deployment cycle (separating the components, reconstituting
the artificial dielectric and placing the artificial dielectric in position at the soil surface) is
perfected in this chapter. Not all techniques worked equally well. In chapter 4 the
characterization of the artificial dielectric, meaning the determination of real relative
permittivity across the frequency band of interest, 700 to 1300 MHz, is described. First,
the measurement system, meaning the fixture, adapters for connection to the network
analyzer, the algorithm for extracting the information of interest, had to be validated.
Then, the actual measurements of materials of interest is reported, finishing with the real
relative permittivity of the best artificial dielectric of the radar tests, as well as what
reflection coefficient it would have if it were spread out on level soil of similar electrical
characteristics to the soil used for the radar tests, and illuminated by a normal-incidence
uniform plane-wave. From an earlier paper, the goal of achieving magnitude of gamma
equal to or smaller than 0.098 across the frequency band of interest had been suggested
[5]. Following chapter 4 is the conclusion, and then there are 3 appendices: 1). Early
attempts at a solution, and rejected methods; 2). Parts lists and sources 3). GulfWax®
MSDS.

For an overview of why finding landmines is difficult the Office of Science and
Technology assessment provides a useful starting point [4]. This study, which opens the
technical discussion with Ground-Penetrating Radar, or GPR, includes appendices treating no less than 13 different approaches to the landmine detection-without-casualties problem. Yet, no single method has proven capable of finding with certainty all types of landmines under all types of conditions without any false alarms. Why is finding landmines hard, in general and specifically for RADAR: 1). Wide variety of physical materials enveloping landmines. For example, there are some magnetic soils [6]; 2). Lack of a pattern or “memory” to landmine positions. In many cases, records have been lost or destroyed, or the landmines were not precisely placed to begin with even if there was a map at first; 3). Variety of different types of landmines, made of different types of materials in different geometries, and the fact that this is not known a priori; 4). The small volume of an individual landmine compared with the volume of possible locations; 5). Pressure-triggered means they can’t be poked with sticks from above, they have to be encountered from the side if using contact methods; 6). Resemblance to surroundings, a sort of electrical camouflage in the case of radar detection; 7). Mobility is constrained by the presence of landmines, while conducting a search for the landmines; 8). Particular problems with GPR include false alarms caused by a similarity between return signals caused by targets (landmines) and harmless natural objects (rock, roots, water pockets, etc.) [4]; 9). Performance can be highly sensitive to complex interactions among mine metal content, interrogation frequency, soil moisture profiles and the smoothness of the ground surface boundary; 10). Trade-off between low-frequencies for penetration and higher frequencies for resolution; 11). It has been observed that while target resolution increases with increasing frequency, the target features are harder to separate from the background clutter of the rough ground interface. This difficulty in separating
information contained in the signal is not due to increasing electrical losses of the ground itself, but rather to the increasing effects of ground clutter with increasing frequency. The higher the frequency the rougher the ground surface appears in the UHF range of operations [7]; 12). Surface height variations similar to burial depth create other difficulties [7]; 13). Dielectric constant and conductivity of target similar to surrounding soil [7]; 14). Size of target comparable to thickness of soil above it [7]; 15). The air/soil interface presents a larger impedance mismatch than target/soil, by a wide margin [7], which translates into a much greater return for the unwanted signal; 16). For an idealized geometry with a perfectly planar ground surface interface, dielectric objects with corners could be identified, distinguished from naturally occurring objects (rocks, etc.) and the orientation determined. This is not possible in general because of the ground reflection [7]; 17). Taken together: the greatest impediment to detecting and classifying dielectric targets buried in natural soil backgrounds with GPR is the random clutter field generated by the rough ground surface. This is demonstrated by computational modeling of GPR wave propagation in air/soil and scattering from buried dielectric targets [7]. In summary, the transmitted radar signal is affected by the dielectric constant of the propagation medium, inhomogeneities of the medium, such as pockets, layers, depth profile or other variation dependent on location, target contrast, target depth, soil type (different permittivity for similar moisture content), moisture content and some of these variations can interact in complex ways. So, why use GPR? Before we treat that question, a brief sketch of the background behind landmines and radar can establish a starting point to that portion of the discussion.
The first pressure triggered landmine was described by H. Freiherr von Flemming in 1726 – the “fladdermine” (flying mine). “It consisted of a ceramic container with glass and metal fragments embedded in the clay containing 0.90 kilos of gunpowder, buried at a shallow depth in the glacis of a fortress and actuated by someone stepping on it or touching a low strung wire [8].”

The first operational radar was the “telemobiloscope” patented by Christian Hulsmeyer in 1904 [9]. “The first public demonstration of his ‘telemobiloscope’ took place on the 18th May 1904 at the Hohenzollern Bridge, Cologne, Germany. As a ship on the river approached, one could hear a bell ringing. The ringing ceased only when the ship changed direction and left the beam of his ‘telemobiloscope’. All tests carried out gave positive results.”

During WW I an inductive bridge was used for detecting landmines, and by the Korean war ground penetrating radar (GPR) was used by the Americans, to locate underground tunnels [10]. Later, Energy-focusing Ground Penetrating Radar (EFGPR), or Focused-Array Radar (FAR), would be developed for landmine detection.

The abstract from the 1995 patent by Sheldon Sandler and others describes FAR as:

A method and apparatus for resolving a radar return by an object or layer below the earth’s surface from a radar return of an air-to-earth interface. More specifically, a method and apparatus for generating short pulse width, broad bandwidth RF signals and for effecting control and timing of transmitted and received radar signals in order to focus an antenna illumination pattern. The apparatus includes an antenna array including a plurality of antenna elements which transmit and receive the RF pulsed signals and a custom control module which generates the control signals fed to the antenna array to delay the RF pulse signals with respect to each other, upon transmission and receipt, in order to form transmit and receive antenna illumination patterns [11].

And, from an abstract taken from Rappaport and Reidy:
The Focused Array Radar (FAR) is a unique time-domain radar system which uses adjustable time delayed signals in a wide multi-element array which focuses transmitted and received signals to detect targets in lossy soil. By making use of specially-designed folded rhombus antenna elements—which are both ultra-wideband and more omnidirectional in the forward direction than a comparable dipole—the FAR optimizes the trade-off between target resolution and penetration depth. These proprietary antenna elements, patented by GEO-CENTERS, INC., faithfully radiate sub-nanosecond pulses with frequency response varying from about 700 MHz to 1.3 GHz, so targets in wet soils within 60 cm of the surface and as small as 8 cm can be resolved. The array signals are focused by establishing the time delay from each element to each sample point in the soil medium, taking into consideration the differing propagation speed in air and various soils, as well as ray path refraction at the air/ground interface. These delays are applied in roughly ten picosecond intervals to the transmitted signal and used to time gate the received signal. By using time delays for a focused wideband pulse, the phases of each frequency component of the radar signal are in effect properly specified for constructive interference. Also, as a result of the time-gating, the large ground surface reflection signal is avoided [12].

FAR is the type of GPR used to get measurements for this paper.

So, why use GPR instead of one of the other 12 methods introduced above? The new ideas for applying GPR to landmine detection are still coming in and even though it is a mature technology the ability to model and separate the landmine signature from the clutter and signatures of natural objects is still in its infancy [4]. There is, therefore, room for future improvement. Why not stereographic GPR with a holographic headset-display? There are a number of advantages to using GPR: 1). It is complementary to conventional metal detectors because of the capability to find non-metallic mines [4]; 2). Generating an image is often possible based on dielectric differences between the target (whether landmine or natural object) and surrounding soil; 3). GPR can be combined with other technologies; 4). GPR can be made lightweight and easy to operate with a scan-rate comparable to other technologies [4].
Design considerations include finding the right trade-off between low-frequencies for greater penetration depth and higher frequencies for greater resolution. Lower frequencies get poorer resolution and higher frequencies are more severely attenuated, although changes in soil moisture might alter the center frequency of choice. Although increased soil moisture increases absorption, it may provide additional contrast of background permittivity, which may actually enhance the signal [4]. For deeper targets, only lower frequencies would work at all. So, achievable resolution would be expected to decrease with depth.

But, what about the ground reflection? FAR, itself, takes the first step in reducing this hindrance. For the purposes of this paper, “ground reflection” shall imply “unwanted ground reflection”. By focusing the energy at the target, ground reflections are reduced. But, FAR does not solve the problem of ground reflections for all cases. In the case of dielectric (non-conducting) targets: the largest single source of undesirable signal is the ground surface itself. Since the ground presents a larger impedance mismatch with the air above it than with the low-contrast, nonmetallic target within it, its contribution to clutter is quite significant [7] even after FAR-techniques are employed.

1.2 The Nature of Ground Reflections

The return from the ground’s surface presents the biggest obstacle to improving FAR systems [5][7][12]-[31]. To see why this is the case, we must first review the propagation of electromagnetic waves in lossy conductors and their reflections at interfaces with low-loss dielectrics.
To start the discussion technically, we should first look at the source-free wave equations for lossy media. We will need the propagation constant. In phasor form [12], [32]-[36].

\[
\nabla^2 \mathbf{E} = \gamma^2 \mathbf{E} \quad (1.1)
\]

\[
\nabla^2 \mathbf{H} = \gamma^2 \mathbf{H} \quad (1.2)
\]

\[
\gamma^2 = j \omega \mu - \omega^2 \epsilon = j \omega \mu (\sigma + j \omega \epsilon) \quad (1.3)
\]

\[
\gamma = \alpha + j \beta \quad (1.4)
\]

\[\alpha = \text{attenuation constant [Nepers/m]}\]

\[\beta = \text{phase constant [radians/m]}\]

\[
\alpha = \omega \sqrt{\mu \epsilon} \left\{ \frac{1}{2} \left[ \sqrt{1 + \left( \frac{\sigma}{\omega \epsilon} \right)^2} - 1 \right] \right\} \frac{1}{2} \text{Np/m} \quad (1.5)
\]

\[
\beta = \omega \sqrt{\mu \epsilon} \left\{ \frac{1}{2} \left[ \sqrt{1 + \left( \frac{\sigma}{\omega \epsilon} \right)^2} + 1 \right] \right\} \frac{1}{2} \text{rad/m} \quad (1.6)
\]

Snell’s Law for one complex half-space on the transmission side of the interface appears as

\[
\gamma_1 \sin \theta_i = \gamma_2 \sin \theta_i \quad (1.7)
\]

\[
\sin \theta_i = \frac{\gamma_1}{\gamma_2} \sin \theta_i = \frac{j \beta_1}{\alpha_2 + j \beta_2} \sin \theta_i \quad (1.8)
\]

and the cosine value will be needed, as well
\[
\cos \theta = \sqrt{1 - \sin^2 \theta}
\]  \hspace{1cm} (1.9)

The intrinsic impedance of the lossy medium may be found by

\[
\eta_c = \sqrt{\frac{j \omega \mu}{\sigma + j \omega \varepsilon}}
\]  \hspace{1cm} (1.10)

Now, suppose there is a circular trench of semi-circular cross section cut into an otherwise level bed of soil, as shown in Figure 1.2. The reason the trench is given a semi-circular profile in cross-section is so that a downward-directed uniform plane wave (normal incidence to the plane of the soil surface) will experience all angles of incidence within the trench itself, from 0 to 90 degrees. The reason the trench follows a circular track is to make sure that both perpendicular as well as parallel E-field (to the reflection plane) will be encountered. This is to illustrate a general case geometry for the reflection coefficient equation-pair with a uniform distribution for all possible combinations of perpendicular and parallel E-field to the plane of reflection. (see equations 1.11 \& 1.12)

Illuminating this feature from above, with the Poynting vector directed towards the minus \( z \)-direction, with a sinusoidal signal of linear polarization, there will be a reflection caused by the air/soil interface described by the following equations in the case of plane-wave excitation.
Figure 1.1: The reflection plane at the interface between 2 media illuminated by a uniform plane-wave.

Figure 1.2: Uniform plane-wave excitation illuminating a circular trench of semi-circular cross section with 0 degree angle of incidence.
Figure 1.3: Close-up of the circular trench of semi-circular cross-section.

\[
\Gamma_1^{b} = \frac{E_1'}{E_1} = \frac{\eta_2 \cos \theta_i - \eta_1 \cos \theta_i}{\eta_2 \cos \theta_i + \eta_1 \cos \theta_i} \\
(1.11)
\]

\[
\Gamma_2^{b} = \frac{E_2'}{E_2} = \frac{-\eta_2 \cos \theta_i + \eta_1 \cos \theta_i}{\eta_2 \cos \theta_i + \eta_2 \cos \theta_i} \\
(1.12)
\]

Figure 1.4: Reflection Coefficient for an air/soil boundary with E-field perpendicular to the reflection plane.
Such well-ordered geometries would seldom apply to an actual case, but show the mechanism of first-order effects. Allowing multiple scaling of roughness features in a Monte Carlo simulation has revealed that small-scale roughness much less than a wavelength in prominence contributes to distortion of the signal reflected from the soil surface, while medium scaled roughness on the order of one-wavelength contributes mostly to a time-shift. Distortion effects, however, are sufficiently variable to disallow simply time-shifting and uniform scaling versus time to recover the original reflected signal [37]. Furthermore, although slant-angle plane-wave illumination backscatter
techniques can also help bring out the desired signal, this technique would not be as effective as directly eliminating the unwanted reflection [38]. The FAR system used later in this paper at Geo-centers is described in the literature [12].

Making matters worse, the relative permittivity of the landmine is often closer to the permittivity of the surrounding soil than the air above the surface. This ensures the desired reflection will be smaller than the unwanted reflection from the air/soil interface. Additionally, there is further degradation of the desired reflected signal away from the incident beam [39] [40].

Artificial dielectric is a material that can artificially look like a dielectric to a transmitted radar signal. For example, a loaf of bread made out of styrofoam, then sliced and held together by a band wrapped around the entire loaf, would exhibit dielectric behavior if pennies were held in-between the slices in a rectangular lattice.

Figure 1.6: (left) Reflection of simulated voltage versus time signal from flat air / soil interface (center) Signal reflected from a rough interface (right) overlaying the two signals shifted in time to maximize the cross-correlation showing how distortion compromises the fit even in the time-window with the greatest similarity between the signals [37].
These are the reasons to consider using artificial dielectrics to reduce the strong unwanted reflection at the air/soil interface:

1. The A.D. can allow maintenance of the distance between the mouth of the radar antenna and the interface.

2. The A.D. can evenly redistribute scattering sources at the surface.

3. The A.D. can remove medium-sized roughness as a feature of the soil surface.

4. The A.D. can allow pushing the target down a specified distance to enable time-windowing of the return signal.

5. In summary, A.D. can reduce distortion, magnitude and phase variations in the reflected signal without appreciably attenuating the signal returned from the target.

Further development could lead to continuous dielectric all the way from a dielectric-filled horn antenna into the ground without any abrupt air / soil boundary, *per se*. It could also lead to controlling the depth of target below the air / soil interface to allow time-windowing techniques to be applied.

This is not an impedance transformation based on A.D. layer thicknesses odd integral multiples of quarter-wavelengths long. Such considerations are beyond the scope of this paper.
2. DEVELOPING AN ARTIFICIAL DIELECTRIC

2.1 Theory

An artificial dielectric that would meet the criteria determined in the previous chapter would match the soil impedance at hand. Thus, the permittivities do not have to match \textit{per se} but rather

\[ \eta = \sqrt{\frac{j \omega \mu}{\sigma + j \omega \varepsilon}} \text{ Ohms} \]  (2.1)

where \( \eta \) equals impedance, \( j \) is the imaginary operator, \( \omega \) equals \( 2\pi f \), \( \sigma \) conductivity, \( \mu \) permeability and \( \varepsilon \) permittivity of the material under consideration. It was shown in Figure 1.4 and Figure 1.5 that conductivity has only a small effect on any possible reflection coefficient from even very wet soil, so the other two factors permeability and permittivity will receive the most attention. This simplifies the impedance formula to

\[ \eta = \sqrt{\frac{\mu}{\varepsilon}} \text{ Ohms} \]  (2.2)

There is a fundamental difference between dielectrics and conductors.

In conductors positive and negative charges are separated by macroscopic distances, and they can be separated by a surface of integration. This is not permissible for bound charges and illustrates a fundamental difference between bound charges in dielectrics and true charges in conductors [32].
While conductors are collections of free charges running around a lattice of non-neutral bound charges, a dielectric is macroscopically neutral but microscopically a collection of electric dipoles each defined by a polarization vector.

$$d_{\mathbf{p}}_i = Q_i \text{ C-m} \quad (2.3)$$

In words, this equation states that the dipole moment equals charge (either positive or negative) times separation distance. In the absence of an externally applied E-field the length goes to zero and there is no displacement of charge at the atomic level. In the presence of an externally applied E-field, charges are displaced from each other by length “l” which represents an increase in potential energy.

Macroscopically, it would not be practicable to add up individual polarization vectors at known locations, so a bulk polarization vector is introduced for a total of $N_e \mathbf{v}$

$$\mathbf{P} = \lim_{\Delta \mathbf{v} \to 0} \left[ \frac{1}{\Delta \mathbf{v}} \sum_{i=1}^{N_e \mathbf{v}} d_{\mathbf{p}}_i \right] = N_e Q_i \mathbf{v} \quad (2.4)$$

dipoles. This behavior applies to the largest class of dielectric materials, and is known as electronic polarization.

In free-space

$$\mathbf{D} = \varepsilon_0 \mathbf{E}_a \quad (2.5)$$

while in a dielectric material
\[ \mathbf{D} = \varepsilon_0 \mathbf{E}_a + \mathbf{P} \quad \text{C/m}^2 \quad (2.6) \]

Where \( \mathbf{P} \) also equals surface charge density for either positive or negative surface charges residing on opposite faces of a brick of dielectric material in the simple case with orthogonally applied E-field. The polarization vector is related to the applied E-field by

\[ \mathbf{P} = \varepsilon_0 \chi_e \mathbf{E}_a \quad (2.7) \]

which implies

\[ \mathbf{D} = \varepsilon_0 \mathbf{E}_a + \varepsilon_0 \varepsilon\chi_e \mathbf{E}_a = \varepsilon_0 (1 + \chi_e) \mathbf{E}_a = \varepsilon_\text{r} \mathbf{E}_a \quad (2.8) \]

where \( \chi_e \) is a dimensionless quantity called electric susceptibility and \( \varepsilon_\text{r} \) is the static permittivity. Normalization produces the relative permittivity more commonly known as the dielectric constant.

\[ \varepsilon_\text{r} = \frac{\varepsilon_\chi}{\varepsilon_0} = 1 + \chi_e \quad (2.9) \]

For the permeability term in equation (2.1), above, it is common to assume freespace conditions for many applications. Here, however, we will need a basic understanding of magnetic materials, in particular diamagnetic materials.
Following the general treatment of electric materials, we start with the atom and it’s dipole moment, in this case *magnetic dipole moment* with \( I \) equal to current created by an orbiting electron and \( ds \) equal to the area of its electric loop.

\[
dm_i = I_i ds_i \quad \text{A-m}^2
\]  
(2.10)

Considered in the context of bulk behavior, with all current loops aligned within parallel planes, for a quantity of magnetic moments, this equality becomes

\[
\mathbf{M} = \lim_{\Delta v \to 0} \frac{1}{\Delta v} \sum_{i=1}^{N_m} \Delta \mathbf{m}_i = n N_m (I ds)_{av}
\]  
(2.11)

where \( \mathbf{M} \) is equal to the magnetic polarization vector and \( \hat{n} \) the unit vector directed perpendicular to the parallel planes containing the electron orbits (electrical loops).

In magnetic material

\[
\mathbf{B} = \mu_0 (\mathbf{H}_a + \mathbf{M})
\]  
(2.12)

with \( \mathbf{M} \) also equal to the equivalent magnetic current density on one face of a brick of magnetic material in the simple case with orthogonally applied B-field.

The magnetization vector is related to the H-field by the *magnetic susceptibility*

\[
\mathbf{M} = \chi_m \mathbf{H}_a
\]  
(2.13)
With static permeability $\mu_s$ defined by

$$B = \mu_0(H_n + \chi_m H_n) = \mu_0(1 + \chi_m)H_n = \mu_s H_n$$

(2.14)

therefore, relative permeability $\mu_r$ can be defined by

$$\mu_r = \frac{\mu_s}{\mu_0} = 1 + \chi_m$$

(2.15)

### 2.2 Practice

In developing an artificial dielectric the same principles apply, “since electromagnetic fields of long wavelength do not discriminate between atomic and macroscopic dipoles [42].” Such an A.D. should be lightweight, moisture resistant, stable, resilient, definite morphology and readily reconfigurable to match a variety of impedances from 0.25 to 0.70 times the impedance of free-space.

Enter Lam, a colorful writer with a wry wit who coins phrases such as “the canonical artillery” and “the scarcity or outright lack of magnetic monopoles”. Two of his papers describe the relationships between permittivity and permeability versus volume fractions for simple cubic lattices of 2 species of conducting magnetic spheres [41] [42]. The restriction of placement of perfect geometric spheres in a uniform lattice to infinity allows for an analytical approach. Each sphere is electrically polarized and interacts with all of the other particles as well as the applied field generating individual multipole moments at each particle’s location. A linear matrix equation with explicitly worked-out coefficients contains all of the multipole interactions among the spheres, and can be...
solved by iteration. The solution appears as a series expansion in powers of the volume fractions of the spheres. Analytical formulas for relative permittivity and permeability were derived. The wavelength of the applied electric-field must be much greater than the center-to-center spacing between spheres, and also the radii of the spheres. The applied field must be quasi-uniform. Single-species can also be used and simplifies the formulas. (see equations 2.16 through 2.20) Volume fraction is defined as

\[ VF = \frac{\text{volume}_{\text{spheres}}}{\text{volume}_{\text{cell}}} \]  

(2.16)

For the simple cubic case, assuming each edge of the cubic cell has length ‘r’

\[ VF_{SC} = \frac{4}{3} \pi r^3 \]  

(2.17)
With \( r(\text{max}) = 0.5 \); \( \text{VF}_{\text{SC}}(\text{max}) = 0.5236 \). For the face-centered cubic case, using the same cell size with the length of each edge = \( 'r' \) there is an eighth of a sphere at each corner, as before, plus half a sphere centered within each face for a total of 4 times the volume.

\[
\text{VF}_{\text{FCC}} = \frac{16}{3} \pi r^3
\]  

(2.18)

The maximum radius of any sphere is now determined by \( \frac{1}{4} \)-th of a face diagonal

\[
r(\text{max})_{\text{FCC}} = \frac{\sqrt{2}}{4}
\]  

(2.19)

So, the maximum volume fraction for the face-centered cubic case works out to 0.7405.

From the paper by Lam [42] a set of equations is obtained with underlying assumptions and derivation described above, which define real relative permittivity and permeability with respect to packing fraction \( (p) \) which is the same as \( '\text{VF}' \) or Volume Fraction.

\[
\frac{\varepsilon}{\varepsilon_0} = 1 + \frac{3p}{A(p)}
\]  

(2.20)

\[
\frac{\mu}{\mu_0} = 1 + \frac{3p}{A(p)}
\]

For sc: \( A(p) = -\frac{1}{R_1} - p + 1.3047 R_2 p^{(10/3)} + 0.0723 R_3 p^{(14/3)} - 0.5289 R_3 p^{(17/3)} + 0.1526 R_3 p^6 \)
For fcc: 
\[
A(p) = -\frac{1}{R_t} - p + 0.0753R_t p^{(10/3)} + 0.2420R_t p^{(14/3)} + 0.0558R_t^2 p^{(17/3)} + 0.0231R_t p^6
\]

for effective dielectric constant, assuming conductivity goes to infinity:

\[R_n = -1\]

for effective magnetic permeability, assuming conductivity goes to infinity:

\[R_n = \frac{n}{1+n}\]

Figure 2.2: Real Relative Permittivity values for sc and fcc lattices of metal spheres of identical radii in a free-space host medium according to Lam equations (Equation 2.20).
Figure 2.3: Real relative permeability values for sc and fcc lattices of metal spheres of identical radii in a free-space host medium according to Lam equations (Equation 2.20).

Figure 2.4: Relative Impedance values for an artificial dielectric comprising metal spheres of identical radii in a free-space host medium.
Some comments are in order on the nature of the artificial dielectric described in Figure 2.1, Figure 2.2, Figure 2.3 and Figure 2.4. The dielectric properties are those of electronic polarization as long as the radius of the spheres is held below about $\frac{\lambda}{4}$ at the smallest operating wavelength [43]. This is to provide adequate margin from resonance effects occurring when the diameter reaches $\frac{\lambda}{2}$. Above this value macroscopic behavior no longer would mimic microscopic behavior of the material. Additionally, it’s important to keep the metal spheres within a lattice spacing of about $\lambda$ to avoid diffraction effects that also occur in natural dielectric materials under similar conditions. The magnetic properties exhibit diamagnetism. The H-field applied encounters and is disturbed by metallic spheres which are not penetrated by the E-field or the H-field below several skin depths. By going around the spheres equal magnitude but opposite phase currents are established which generate an opposing H-field. The super-positioning of the applied and resultant H-fields produces a reduced combined H-field as a result, ergo diamagnetism. The real relative permeability of diamagnetic materials is less than ‘1’.

In order to realize an actual material based on the principles outlined above, after many failed attempts using a form made of pre-expansion styrofoam beads covered with conductive mixtures and powders, some 6 mm diameter craft beads were located. These were hollow plastic balls with a very thinly plastic-coated conductive shell near the surface. Normally, they would be intended for use in crafts projects. It was discovered that by combining them with hollow polypropylene balls of 20 mm diameter, the real relative permittivity could be varied to any value between about 2 to 5. Other values may have been possible, but were not tried. These hollow balls were normally intended to
create a floating thermal blanket in hot tubs and other large containers filled with liquids exhibiting a temperature different than ambient. To streamline discussion, the conductive beads will be called “beads” in this paper, and the hollow polypropylene balls either “balls” or “spheres”. In June 2000, 10 mm hollow polypropylene balls replaced the larger ones, and were used with the conductive beads for all results shown in this paper. These results were obtained on an HP 8510 network analyzer producing $S_{11}$ curves from an open ended circular cavity half-filled with paraffin at the shorted, probe-fed end.

Table 2.1: Artificial Dielectric elements mixed in various proportions.

<table>
<thead>
<tr>
<th>Quantity 5000</th>
<th>Quantity 7 L</th>
</tr>
</thead>
<tbody>
<tr>
<td>10 mm diameter hollow polypropylene balls from McMaster-Carr. ($177.17 per 5000)</td>
<td>6 mm insulated, conductive beads from Michaels, Inc. Bar Code: 0-16318-92545-2 ($2.99 per 400)</td>
</tr>
<tr>
<td>Specific Gravity of bulk volume in a vacuum: 0.257. Measured diameter 10.0 mm Natural bulk density 965.9 per liter</td>
<td>Specific Gravity of bulk volume in a vacuum: 0.507. Measured Diameter: 5.8 mm Natural bulk density: 4924.4 per liter</td>
</tr>
</tbody>
</table>

In Figure 2.2, Figure 2.3 and Figure 2.4 it was assumed that under certain special conditions, such as hand-placement of beads with a tweezer, volume fractions up to a maximum of 0.74 could be achieved. In practice, however, when beads or spheres were poured into a 2 L graduated beaker without any shaking and counted, volume fractions of 0.50 were obtained for both beads and balls. It was observed, that any container would alter the free-space volume fraction by disturbing the underlying lattice (complicated though it may be) at the boundaries -- the so-called “surface effects”. Similar studies have been conducted at Princeton University with a packing fraction of 0.49 assigned the
designation “freezing point” [44] [45] and further research lead to the conclusion that M&M’s when poured and shaken, packed more densely than spheres [46].

The volume fraction ‘0.50’ was used as the target value whenever artificial dielectric was poured. 1 L of spheres at volume fraction 0.50 and 1 L of conducting beads at volume fraction 0.50 when poured together would not necessarily yield 2 L of mixture. This behavior was observed and noted as a possible error source for mixture specifications, but not investigated further.
3. THE GEO-CENTERS RADAR MEASUREMENTS

During June 10-17, 2000 numerous radar measurements were conducted at Geo-centers, in Newton, MA to find an artificial dielectric which would match soil. Using a pulsed radar with a proprietary antenna placed a few tens of centimeters above a soil test bed, intensity versus time plots were obtained for several different Artificial Dielectrics (see Figure 3.1) comprising 6 mm diameter conducting beads and either 10 mm or 20 mm diameter hollow polypropylene balls. All of the results reported in this paper use the smaller of the two sizes.

![Figure 3.1: Soil bed containing a disturbed surface “smoothed with AD composed of 6 mm insulated conducting beads and 20 mm balls, in a volume ratio of 1.73 to 1” [5].](image)

The radar system consisted of one antenna send-receive pair with attendant electronics, for a FAR (Focused Array Radar) or EFGPR (Extremely Focused Ground
Penetrating Radar). This system has been described in the literature [12]. The radiated energy is predominantly between 700 MHz and 1300 MHz. Only one antenna transmit–receive pair was employed.

Volume ratios refer to unmixed spheres, since mixing changed the total volume – i.e. 1 [Liter] of beads mixed with 1 L of balls did not yield 2 L of Artificial Dielectric. At first, AD was built up in layers made up of fractions of a liter at a time. But, towards the end, two different methods were devised to greatly speed up the cycle between Artificial Dielectric deployments.

Figure 3.2: (left to right) 1. The Shake-n-pour system, 2). Oriented for pouring conducting beads from one side of the heavy cardboard baffle and polypropylene balls from the other side, 3). The shipping can rotated until both species of spheres poured at the proper rate to insure uniform deployment, 4). The system as it would look when separating the mixture of spheres into beads and balls in preparation for the next deployment.

The first of these two enhanced methods was dubbed “shake ‘n pour”. To start, a shake ‘n pour can was constructed. This required one 13 L, by visual estimate, metal shipping can, some heavy cardboard and a 1 inch x 10 inch x 12 inch overnight shipping box. The heavy cardboard was cut to fit along the axis inside the shipping can with no gaps when taped, dividing the can into two equal half-right-cylinder volumes. On one half of the baffle, 7 mm wide slots were cut, 1 per 14 mm of length. The edges were
trimmed to allow 6 mm diameter beads to drop through. The other half of the baffle was left whole. Up to 5 L of mixed beads and balls could be separated with this device and re-deployed within 5 minutes. To separate a mixed quantity the AD was poured into one half of the can with the slots pointing down. Then, rotating the can by about 60 degrees and gently shaking, the beads could be made to drop through. To pour, the can was rotated another 120 degrees (180 degrees total) and fitted into a hole cut into the bottom of the overnight shipping box acting as a combiner. The lowest corner of the box had a 25 mm wide square hole cut out of it to act as a spout. When pouring, the ratio of beads to balls could be controlled by the angle of rotation of the can with respect to it’s axis. In this way, very even mixes of artificial dielectric could be rapidly generated. If done haphazardly, the homogeneity of the mix was lost, and good results could not be achieved. For with handling, the plastic balls had a tendency to float to the surface.

Figure 3.3: CAD Drawing of the heavy cardboard baffle used with the Shake-n-pour system.
The second enhanced technique for AD re-deployment was called “mix ‘n pour”. For this method, a tapered plastic cup that held about 225 ml of beads when heaped above the rim was used to obtain the bulk volume ratio of balls to beads. A variation of the original layered construction, taking an inverted cone as an example, the number of cups required per layer was calculated to maintain constant thickness. There was a lot of visual estimation involved with this approach since, for example, a layer of beads would have a tendency to sink down into a plastic ball layer, while the reverse was true for a layer of balls, which would tend to float above the beads. But, the ratio was controlled by counting how many heaping cups of each constituent element were used. Sometimes the surface was patted smooth, as a final step, and sometimes not for either of the 2 rapid techniques.

Before building AD, the soil was made flat, a radar signal recorded and then some soil removed in various patterns to provide a rough surface. A 2nd radar signal was then taken. Removed soil was kept in a clean container out of view of the radar, and the soil bed surface covered with a thin layer of plastic from the cleaners (for clothing). This allowed the AD to be removed in one quick bunching of the corners of the plastic sheet, and poured into the shake ‘n pour can, or other container for separation. This thin plastic sheet was used for all of the measurements. After radar signals were obtained for 3 different deployments of the same packing fraction of AD, except as noted otherwise, the soil was replaced and one final radar signature recorded to complete the before and after baseline.

The soil in the bed was profiled for dry specific gravity and dry moisture before any measurements were taken. Some of these values were checked after all radar measure-
Table 3.1: Soil Test-bed moisture and density profile with depth.

<table>
<thead>
<tr>
<th>Depth</th>
<th>Before radar measurements</th>
<th>After radar measurements</th>
</tr>
</thead>
<tbody>
<tr>
<td>Surface / Dry Sp. Gravity</td>
<td>Mean 0.80</td>
<td>0.86</td>
</tr>
<tr>
<td></td>
<td>STD 0.050</td>
<td>0.083</td>
</tr>
<tr>
<td>Dry moisture</td>
<td>Mean 0.093</td>
<td>0.097</td>
</tr>
<tr>
<td></td>
<td>STD 0.0047</td>
<td>0.012</td>
</tr>
<tr>
<td>16 cm deep/Dry Sp. Grav.</td>
<td>Mean 0.85</td>
<td>-NA-</td>
</tr>
<tr>
<td></td>
<td>STD 0.13</td>
<td>-NA-</td>
</tr>
<tr>
<td>Dry moisture</td>
<td>Mean 0.27</td>
<td>-NA-</td>
</tr>
<tr>
<td></td>
<td>STD 0.052</td>
<td>-NA-</td>
</tr>
<tr>
<td>26 cm deep/Dry Sp. Grav.</td>
<td>Mean 0.78</td>
<td>-NA-</td>
</tr>
<tr>
<td></td>
<td>STD 0.031</td>
<td>-NA-</td>
</tr>
<tr>
<td>Dry moisture</td>
<td>Mean 0.30</td>
<td>-NA-</td>
</tr>
<tr>
<td></td>
<td>STD 0.025</td>
<td>-NA-</td>
</tr>
</tbody>
</table>

...ments had been completed. In equations 3.1 and 3.2, \( \rho_d \) is “dry density”, \( m_d \) is “dry mass”, \( V \) is “volume”, \( \theta_g \) is “dry moisture ratio” expressed as a percentage and \( m_w \) is “wet mass”

\[
\rho_d = \frac{m_d}{V} \quad (3.1)
\]

\[
\theta_g = \left[ \frac{m_w - m_d}{m_d} \right] (100) \quad (3.2)
\]

Data was acquired for 2 different packing fractions, 0.125 and 0.286. Correlation coefficients were found referred to the original flat soil. These are given in Table 3.2, and the data plots are shown in Figure 3.4, Figure 3.5 and Figure 3.6.
Table 3.2: Correlation coefficients between curves in Figure 3.4, Figure 3.5 and Figure 3.6 referred to the original flat soil for packing fractions (p) of 0.125 and 0.286. Row 1: Three A.D.’s were built-up. Roughness was introduced by a 6.1 L inverted cone dug out of level soil, indicating that 6.1 L of soil was temporarily removed; Row 2: Three A.D.’s were built-up. Roughness profile 2 trenches of circular cross-section oriented along the H-field. About 10 L of soil was removed. Row 3: The last A.D. from ‘Row 2’ was finger-stirred on the surface, and re-measured.

<table>
<thead>
<tr>
<th>p</th>
<th>method</th>
<th>flat</th>
<th>rough</th>
<th>AD #1</th>
<th>AD #2</th>
<th>AD #3</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.125</td>
<td>Shake’npr</td>
<td>1.0000</td>
<td>0.9154</td>
<td>0.9924</td>
<td>0.9836</td>
<td>0.9870</td>
</tr>
<tr>
<td>0.286</td>
<td>Layers</td>
<td>1.0000</td>
<td>0.9194</td>
<td>0.9882</td>
<td>0.9901</td>
<td>0.9908</td>
</tr>
<tr>
<td>0.286</td>
<td>Layers</td>
<td>1.0000</td>
<td>0.9194</td>
<td>0.9913</td>
<td>0.9926</td>
<td>0.9934</td>
</tr>
</tbody>
</table>

Figure 3.4: 3:1 BALLS:BEADS (BULK VOLUMES) IN 6.1 L INVERTED CONE. Corresponds to Row 1 in Table 3.2. Legend: red/A.D. 1; magenta/A.D. 2; blue/A.D. 3; dashed black/rough ground; dotted black/original flat dirt; dashdot black/dirt added back to rough ground and re-leveled.
Figure 3.5: 0.75:1 BALLS:BEADS IN 10 L 2 H-FIELD DIRECTED CIRCULAR TROUGHS / LAYERED. Corresponds to Row 2 in Table 3.2. Legend: red/A.D. 1; magenta/A.D. 2; blue/A.D. 3; dashed black/rough ground; dotted black/original flat dirt; dashdot black/dirt added back to rough ground and re-leveled.

Figure 3.6: 0.75:1 BALLS:BEADS IN 10 L 2 H-FIELD DIRECTED CIRCULAR TROUGHS / STIRRED. Corresponds to Row 3 in Table 3.2. Legend: red/A.D. 1; magenta/A.D. 2; blue/A.D. 3; dashed black/rough ground; dotted black/original flat dirt; dashdot black/dirt added back to rough ground and re-leveled.
4. CHARACTERIZING THE ARTIFICIAL DIELECTRIC

4.1 Preliminary Considerations

Having matched artificial dielectric to soil in the laboratory using radar over a soil bed, it became important to substantiate this claim through rigorous characterization of the artificial dielectric’s real relative permittivity (RRP) versus frequency over 700 MHz to 1300 MHz. There could be many ways to obtain this result, but all would be unlikely to succeed unless taking all of the following points into consideration, at least: the granularity of the artificial dielectric; the frequency range; all RRP values of interest; environmental conditions; compatibility with available laboratory equipment (e.g. Type-N connectors / sufficiency of equipment dynamic range, etc.); efficacy of the solution-algorithm (i.e. non-singular matrices); sample selection; economics; time window (availability of parts); broadband capability; possibility to prepare samples within necessary tolerances and material sample compatibility with the measurement set (e.g. liquids may make good samples for a beaker and probe, but not a free-space lens pair).

There could be many possible ways to characterize \( \varepsilon_r \). A brief search through the literature reveals dozens or even hundreds of methods [47]. Most of these won’t work for the task at hand for the following reasons:

1). The granularity of the material under test rules out methods used for liquids, such as inserting a thin probe into the material.

2). The sample-size must be sufficiently large to “see” the bulk properties of the material in the underlying lattice structure. This rules out small coaxial cells, and techniques more suitable for thin films.
3). The measurements must be broad-band to include the operating range of the radar whose energy is concentrated between 700 MHz to 1300 MHz. This disallows any methods dependent on c.w. resonant cavities.

4). The sample-size must not be so large that unwanted and detrimental higher-order-modes are introduced. This rules out GTEM cells, and the like.

5). Surface effects must not dominate the measurement. Suppose artificial dielectric is arranged in a rectilinear lattice, such as simple cubic, extending to +/- infinity in three directions (X,Y,Z). Conducting spheres of 1 mm radius are positioned 8 mm apart, center-to-center from the nearest neighbor. Now, suppose an imaginary cube of 1 m per edge is rotated and translated all around this space to random positions and orientations. There may be different numbers of whole conducting spheres enclosed, as well as fractional spheres, for the various placements. So, for complete specification of an artificial dielectric it is necessary to determine volume fraction, lattice description as well as the nature of the surface of enclosure. It’s often easier to discuss materials as if they are boundless, but these must eventually be tied to actual measurements based on samples affected by surface-related choices.

As a more mundane matter, with the actual plastic beads used, freezing them in a block of ice, then cutting to size would have produced misshapen spheres at the surfaces. These fractional forms would be open instead of closed geometries. The inner surface of the conducting shell would have been exposed changing the analytical formulas used to predict $\varepsilon_r$, for these formulas are geometry-dependent.

6). The lattice geometry must not dominate the measurement. For example, it may be possible to get different $\varepsilon_r$ values for materials “A” and “B” even though both have the
same volume fraction, if “A” is rectilinear (fcc, sc) and “B” curvilinear (coaxial). It has been demonstrated, analytically, that fcc gives different real relative permittivity values than sc at the higher volume fractions (< ca. 0.5).

7). Scaling, transposing effects must not dominate the measurement. For example, in going between the different forms described in “5” and “6”, above, one lattice may have only whole geometric conductor shapes, while another may introduce a population of altered conductor geometries. An artificial dielectric comprising entirely solid silver spheres, will have a different $\varepsilon_r$ than one comprising entirely solid-silver hemispheres or hemispherical shells. This could also apply to population-fraction variations, and variations in the distributions of non-uniform geometric species.

From the above examples, one may conclude that there are many opportunities for getting bad data. But, by keeping the measurements and simulations fairly simple it is hoped to avoid many of these deeper discussions while still building confidence in the data and results to follow.

8). All of these same principles apply in a more complicated fashion to freely poured collections of loose-composition artificial dielectrics.

The method used by Hipp [48] and others was based on determining the impedance of one end of a transmission line, typically coaxial cable, filled with MUT with the other end terminated in either a short or open. This method was not pursued, because the instrument available was not an impedance bridge, but a network analyzer. An Agilent seminar on measurements recommended not using a network analyzer to get impedance values because the network analyzer was intended to be used in a system with a characteristic impedance matching the network analyzer’s characteristic impedance.
The value of $\varepsilon_r$ was determined by measured s-parameters for room-environment air and Vaseline® in a 20 cm Hewlett-Packard type-N precision coaxial air-line using

$$\varepsilon_r = \frac{\varepsilon_r}{\varepsilon_0} = \frac{\beta^2}{\beta_0^2}$$

with good results. The rivets holding the type-N female connectors were drilled out, and the connectors unscrewed from the threaded ends of the outer conductor. The silver-coated center conductor was removed from the hand-force press-fit contact-socket. So, the entire hard-line could be disassembled into four separate parts. A thin shorting washer was made from copper foil and placed at the inside face of one of the type-N female end jacks. The precision air-line was re-assembled in working order with the short in place. The 8714ES was calibrated. The shorted jack was connected to Port 1 of the 8714ES with a good quality coaxial cable. Electrical Delay was dialed until the short became the smallest fuzzy point on the Smith Chart, consistent with $0+j0$ Ohms. This value of electrical delay was noted. Then, the near end was disassembled and the shorting washer removed. It didn’t matter which end was selected to be the “near end”. Vaseline® was forced into the outer conductor tube from the near end, with the inner conductor in place and centered by visual inspection. The far end was also left open. The Vaseline® was squeezed into the tube with as much pressure as could be generated by holding everything tightly by hand, and squeezing the lever on the manual bulk caulk gun. Vaseline® oozed out of the far end and was wiped quasi-square at both ends, just even with the step on the inner conductor to form the male contact in order to “kiss” the type-N connector after re-assembly with no gaps or compression of the MUT. Without using the bulk caulk gun experience indicated that no useful results could be obtained for Vaseline®. This is thought to be due to all of the voids and bubbles which naturally
form. Real relative permittivity values within 0.08 error of published values were obtained – published values over the frequency range 0.0001 MHz to 10000 MHz for air, ‘1’, for Vaseline® ‘2.16’ [49]. This method was rejected for use with the 3-1/8 inch E.I.A. coaxial cell, because it was thought extremely difficult to get a good short using this technique with the larger diameter outer conductor. But, the equation 
\[ \varepsilon_r = \frac{\varepsilon}{\varepsilon_0} = \frac{\beta^2}{\beta_0^2} \]
still applied to the method that was finally used. So, we start close to a solution right away, it is just a matter of figuring out how to de-embed the intermediate stages created by unknown connectors, transitions and in some cases cabling. Once the method was found for extracting this information, the un-calibrated state of the network analyzer itself became a contributor to these two intermediate stages. The method would work either with a calibrated or un-calibrated network analyzer, so the latter state was later chosen to save effort.

**Table 4.1: hp precision air-cell dimensions.**

<table>
<thead>
<tr>
<th>Description</th>
<th>Measurement</th>
</tr>
</thead>
<tbody>
<tr>
<td>Length (min)</td>
<td>158.00 mm</td>
</tr>
<tr>
<td>Length (max)</td>
<td>158.0 + 10.0 mm = 168.0 mm</td>
</tr>
<tr>
<td>Length (mean)</td>
<td>163.0 mm</td>
</tr>
<tr>
<td>Outer diameter of inner conductor</td>
<td>3.18 mm</td>
</tr>
<tr>
<td>Inner diameter of outer conductor</td>
<td>10.0 mm</td>
</tr>
</tbody>
</table>
Good measured results for air were obtained in July 2007 using WR284 with the 2-Line method, see Figure 4.2. These were the first good measured results of permittivity. One cell for holding MUT was already made and one additional cell was made by cutting a second commercially-made straight section of WR284, having the cut end milled on a milling machine to within +/- 0.001 inches and soldering a new circular flange onto the end with a basic propane torch like the ones used for plumbing. Solder was 50/50 lead/tin with the appropriate flux.
4.2 System Design, Test and Verification

After scanning the literature as well as conducting empirical trial and error for leading candidates described in the previous section, it was decided to use the 2-Line method in 3-1/8 inch coax [52][53][54]. This method also works for waveguide, but at the frequency range of interest, the waveguide would have been either huge or working below cutoff. Eliminating the troublesome center conductor was, therefore, tied to even bigger problems, see ‘Appendix A’. 

Figure 4.2: RRP for air in WR284 using the 2-Line method with measured S-parameters. This is the first good permittivity measurement obtained. Cutoff for TE_{10} occurs at 2.078 GHz.
The 2-Line method does not need any calibration standards to find the propagation constant in the MUT. The only drawback is, that as the name implies two lines are necessary. The lines must be of different lengths and of constant and equal line impedance versus length. Each line, in the method used in this paper, will be filled completely with MUT, and becomes a test cell. It is the propagation constant inside the test cell that is of interest. This method may be used for TE\textsubscript{10} mode in rectangular waveguide, TEM mode in coaxial cable, or other types of transmission lines. The mode must be pure, so for TEM must be operated below cutoff for H.O.M.’s. In rectangular waveguide the frequency band of operation typically lies between cutoff for TE\textsubscript{10} and cutoff for TE\textsubscript{01}.

The positions of the reference planes are not required, because cascade matrices are employed. A cascade matrix is built for each system from the S-parameter measurements of the entire system including the test cell filled with MUT, adapters, cables and uncalibrated ports on the network analyzers. Once the cascade matrices have been obtained for both cells, they may be combined into an eigenvalue equation, which produces the propagation constant. The propagation constants should be the same for both cells. Only 2-port systems are considered in this paper.

Once the propagation constant is found it may be set equal to an equation containing permittivity as the only unknown. In this way, the permittivity can be obtained. There are different formulas for this, depending on which mode is under consideration. In this paper, Equation 4.1 is used, since the mode under consideration is assumed to be pure TEM.
\[ \varepsilon_r = \frac{\varepsilon - \varepsilon_0}{\varepsilon_0} = \frac{\beta^2}{\beta_0^2} \quad (4.1) \]

The basic cascade matrix looks like this (Equation 4.2):

\[ T^i = \begin{bmatrix} e^{-\gamma_i} & 0 \\ 0 & e^{\gamma_i} \end{bmatrix} \quad (4.2) \]

Here the letter “i” designates the long test cell, “j” is used for the short test cell. Similarly, cascade matrices may be written for everything between the network analyzer ports and the test cell, including the uncalibrated ports, which then looks like this (Equation 4.3):

\[ M^i = XT^iY \quad (4.3) \]

The X and Y terms do not have superscripts, because they do not change for each test cell. Also, the cascade matrices for X and Y are of the same form as Equation 4.2. The designation “M” could stand for “measured” for this is the matrix that is obtained for the system from the measured S-parameters. The designation “T” is apt, as well, and could stand for “Test Cell”. Writing another equation for the short test cell, like Equation 4.3 except with “j” superscripts, the two equations may be combined into an eigenvalue equation.

\[ M^j X = XT^j \quad (4.4) \]
\[ M^\| = M^T \left[ M^T \right]^{-1} \]  \hspace{1cm} (4.5)

\[ T^\| = T^T \left[ T^T \right]^{-1} \]  \hspace{1cm} (4.6)

\( T^\| \) is diagonal and its diagonal elements are the eigenvalues of \( T^\| \) and \( M^\| \). Since \( M^\| \) is obtained from measurements, that is the matrix that will be used to find the eigenvalues, which are then equal to the eigenvalues of \( T^\| \), \( \lambda^\|_{1T} \) and \( \lambda^\|_{2T} \), where:

\[ \lambda^\|_{1T}, \lambda^\|_{2T} = e^{\pm \gamma(l_i-l_j)} \]  \hspace{1cm} (4.7)

So, in general,

\[ \gamma = \frac{\ln \left( \lambda^\| \right)}{l_i-l_j} \]  \hspace{1cm} (4.8)

which may be found by averaging the two eigenvalues from the measurements:

\[ \lambda^\| = \frac{1}{2} \left[ \lambda^\|_{1M} + \frac{1}{\lambda^\|_{2M}} \right] \]  \hspace{1cm} (4.9)

with

\[ \lambda^\|_{1M}, \lambda^\|_{2M} = \frac{(M^\|_{11} + M^\|_{22}) \pm \sqrt{(M^\|_{11} - M^\|_{22})^2 + 4M^\|_{12}M^\|_{21}}}{2} \]  \hspace{1cm} (4.10)

and
The measurements matrix for the short test cell is of the same form as Equation (4.2). In the case of lossless transmission lines of zero-length, it may be seen by inspection that both Equation (4.2) and Equation (4.11) reduce to

\[
M^i = \frac{1}{S_{2i}} \begin{bmatrix}
S_{12i} & S_{11i} & S_{1i} \\
S_{12i} & -S_{21i} & 1 \\
-S_{22i} & 1 & 1
\end{bmatrix}
\]

(4.11)

Let’s look, briefly, at an idealized example for the 2-Line method, as revealed by a simple test case. The simplest imaginable system of cascade matrices would contain one test-cell sandwiched between two sections of zero-length. Considering two such systems with cells of different lengths, but otherwise the same, leads to a useful example. The zero length stages are described by

\[
\begin{bmatrix}
1 & 0 \\
0 & 1
\end{bmatrix}
\]

(4.12)

which leads to \( M^i = T^i \) and \( M^j = T^j \). Suppose there are two test-cells of 30 mm and 10 mm respectively for 50 Ohm coaxial transmission line in air (inner conductor outer diameter 10 mm outer conductor inner diameter 23 mm), filled with lossless material of relative permittivity 1.39. From a CST STUDIO SUITE™ 2008 [50] simulation keeping the letter “i” referring to the longer test section, we find in a 50 Ohm system
\[ S_{11} = -0.1118 - 0.0766i \]
\[ S_{12} = +0.5600 - 0.8173i \]
\[ S_{21} = +0.5600 - 0.8173i \]
\[ S_{22} = -0.1118 - 0.0766i \]
\[ S_{11j} = -0.0168 - 0.0498i \]
\[ S_{12j} = +0.9462 - 0.3192i \]
\[ S_{21j} = +0.9462 - 0.3192i \]
\[ S_{22j} = -0.0168 - 0.0498i \]  (4.14)

Cranking through the matrix mathematics, we then find

\[
M^i = \begin{bmatrix}
0.5705 - 0.8326i & 0.0 - 0.1368i \\
0.0 + 0.1368i & 0.5705 + 0.8326i
\end{bmatrix}
\]
\[
M^j = \begin{bmatrix}
0.9488 - 0.3201i & 0.0 - 0.0526i \\
0.0 + 0.0526i & 0.9489 + 0.3201i
\end{bmatrix}
\]
\[
M^{ij} = \begin{bmatrix}
0.8006 + 0.6074i & 0.0 + 0.0998i \\
0.0 - 0.0998i & 0.8006 - 0.6074i
\end{bmatrix}
\]  (4.15)

with

\[
\lambda_{1M}^{ij} = +0.8006 - 0.5992i
\]
\[
\lambda_{2M}^{ij} = +0.8006 + 0.5992i
\]  (4.16)

Since the complex conjugates of these two eigenvalues are already equivalent, there is no need for averaging. Choosing \( \lambda_{2M}^{ij} \) the propagation constant may be found by

\[
\gamma = \frac{\ln(\lambda^{ij})}{l_i - l_j} = +32.1226i
\]  (4.17)
implying $\beta = 32.1226$. Comparatively, the phase constant for a lossless transmission line of relative permittivity 1.39 at 1.3 GHz would be found by

$$2\pi \sqrt{\varepsilon_r \cdot \frac{f[GHz]}{0.3}}$$

(4.18)

which equals 32.1003 for an error of 0.0007.

$$\frac{32.1226 - 32.1003}{32.1003} = 0.0007$$

(4.19)

A word about the two eigenvalues. For most results used in this paper, the eigenvalues were virtually identical at all frequencies with no easily discernable advantage to using one or the other. Therefore, in those cases the eigenvalue propagation constant designated as ‘1’ was used, and the eigenvalue propagation constant ‘2’ was ignored. See Figure 0.10 and Figure 0.18 in Appendix A.

In accordance with the plan to use large diameter hard-line coax for the test cells, calling around to various FM Radio Stations in the area, one was located that had had a “voltage event” leading to the scrapping of several tens of feet of 3-1/8 inch diameter transmission line. Just as with other sizes of coax more commonly found in the laboratory, like 0.141 inch, the 3-1/8 inch E.I.A. line is designated based on a nominal outer diameter of the outer conductor. The inner diameter of the outer conductor is specified as 3.027 inches and the outer diameter of the inner conductor is 1.315 inches for a characteristic impedance in air of 50 Ohms. After obtaining permission [55], some of the least damaged 3 foot lengths of line were promptly harvested with a hacksaw.
Connecting such large diameter test cells to Type-N connectors in the laboratory would require an adapter of some kind. If the adapter was too long, it could lead to time-delay problems with the network analyzer, or if the VSWR of the unit, or insertion loss, was too high there might not be sufficient signal for a good measurement given the system dynamic range and noise figure. Taking a closer look at possible measurement error due to excessive VSWR, the Type-N to 3-1/8 inch adapters would contribute the most to VSWR degradation, so simulations were performed using CST STUDIO SUITE™ 2008 [50] to see how much of an effect there would be. Excluding noise and dynamic range, for the 2-Line method a VSWR of 1.17:1 or better was shown to be more than sufficient to ensure good results. Simulation results are shown, below, with s-parameter inputs for worst-case Return Loss of 21.93 dB over 0 to 1.5 GHz.

\[
VSWR = \frac{10^{RL_{L}[dB]/20} + 1}{10^{RL_{L}[dB]/20} - 1} = 1.1741
\]  

(4.20)

Figure 4.3: MUT as specified and from 2-Line Method with simulation S-parameters as inputs are virtually identical.
Figure 4.4: Negligible error using s-parameters from simulation of realistic test geometry.

Figure 4.5: S-parameter inputs used to verify system VSWR specification.
Figure 4.6: Geometry for simulation of Type-N to 3-1/8 inch E.I.A. adapter. The voids in the CAD drawing were filled with perfectly conducting metal. Various permittivities were tried for the blue disc, to minimize VSWR.

Based on this simulation, 2 Type-N to 3-1/8 inch adapters and 4 pairs of 3-1/8 inch clamp-type flanges were ordered from Myat, Inc. An engineer at the company reported typical VSWR for this adapter of 1.08:1 over 100 MHz to 1500 MHz [56] and from diagrams on the website the unit appeared to be very compact. See Figure 4.12. Since the
equipment is manufactured for use at high power FM broadcast stations, low insertion loss was assumed.

Finishing the design of the measurement cells consisted in large part in finding a strategy that would produce the least measurement error for the least construction effort. Inner and outer conductor diameters are fixed, the only dimensions to find were the lengths of the 2 cells that would work the best, one long and one short which depends partly on the sample lengths. The effects of noise and dynamic range in the actual system were unknown, and these could be expected to vary with frequency. Also, effects of sample lengths and cell lengths were unknown although in general keeping these as short as possible would push all related resonance effects higher in frequency. In the optimistic view they would be sent above the frequency band of interest. Higher Order Modes presented another unknown, but keeping the faces of the samples flat and perpendicular to the direction of propagation ensured minimization [57]. Possible approaches were broken down into four types as expressed in Table 4.2. A main strength of the

<table>
<thead>
<tr>
<th></th>
<th>Shortest Sample Length</th>
<th>Longest Sample Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>Shortest Test Cell</td>
<td>A</td>
<td>B</td>
</tr>
<tr>
<td>Longest Test Cell</td>
<td>C</td>
<td>D</td>
</tr>
</tbody>
</table>

Table 4.2: Possible approaches to selecting test cell and sample lengths.

approach in ‘A’ would be to make all system resonances as high as possible in frequency. The arrangement in ‘B’ would maximize the amount of the signal that was due to the Material Under Test. The best thing about ‘C’ would be to ensure the maximum attenuation of unwanted H.O.M.s before reaching the measurement reference planes. Using ‘D’ would combine the good and bad effects of ‘B’ and ‘C’. It was decided to go
with plan ‘B’ since signal level was expected to be of greatest importance. Plan ‘A’ would also give strong signal levels, any given dimensional tolerance would be a larger percentage of a smaller sample. So, ‘A’ was ruled out for lack of availability of a precision machine shop.

Even if everything was made perfectly, unavoidable H.O.M’s would appear at 1727 MHz because of the coaxial transmission line geometry [58]. In Equation 4.21 below, ‘D’ is the inner diameter of the outer conductor, and ‘d’ is the outer diameter of the inner conductor, in inches. The units for ‘F’ are in megahertz. In the absence of discontinuities 95% of the theoretical cut-off frequency can be achieved with a pure TEM signal using off the shelf components.

\[ F_{\text{Cut-off}} = \frac{7500}{\sqrt{\varepsilon (D + d)}} \]  

(4.21)

In simulation there were problems whenever the difference between sample lengths was long enough to support more than ½ wavelength of signal. For an upper design-frequency of 1500 MHz and RRP of ‘6’, then, 37 mm would be the longest difference in sample lengths expected to yield good results. The effect at the low end of the frequency range, 700 MHz, or samples with an RRP close to ‘2’, or both, would be a reduction in the impact of the sample on the signal. Since bandwidths using this algorithm approaching 8:1 have been successfully achieved, this was not expected to be a big problem. The bandwidth here is only on the order of 2:1. Using the 37 mm length as a basic increment, the sample length for the design in the short cell becomes 37 mm and in the long cell 74 mm.
There is now sufficient information to specify the lengths of the inner conductors. But, before adopting the shear minimums it may be good to add one more basic increment to each length. In this way, the paraffin toroid could act as a barrier between M.U.T.s of general materials such as sand, garden mulch or artificial dielectric beads and plastic spheres. Adding another 1 mm for the expected minimum wander of the chopsaw we now have the first entry in Table 4.3:

Table 4.3: Chosen design lengths for measurement cell components.

<table>
<thead>
<tr>
<th>Inner Conductor</th>
<th>Outer Conductor</th>
</tr>
</thead>
<tbody>
<tr>
<td>Short Cell “Juliet”</td>
<td>75 mm</td>
</tr>
<tr>
<td>Long Cell “India”</td>
<td>112 mm</td>
</tr>
</tbody>
</table>

Cutbacks of 19.4 mm are specified for the inner conductors, or in this case 20 mm was added to each end of the inner conductor to find the corresponding outer conductor length. Because of chop-saw wander, all uncertainty regarding inner conductor connector clearance at the adapters was forced to be too much instead of too little. Bolting a measurement system together with insufficient cutbacks could cause severe damage to the adapters and cells. The long cell was subscript “i” in the literature and the short cell “j”. To avoid confusion dealing with hundreds of simulation results and measurement data files, the cells were named “India” and “Juliet” according to the phonetic alphabet used for 2-way radio communications.

Another innovation depended on whether the long sample would work just as well if cut into two pieces. In the extreme case, a gap of 1 mm between two toroids 37 mm long each was supposed. Allowing such a split would make the generation of samples modular, whereas a single long piece would not have any other use short samples could
be used as building blocks under a wider variety of conditions. In the actual long sample, this gap was reduced to a length as small as possible (~ 0.1 mm) and the common toroid between cells was left a millimeter or two longer than the precision piece.

Further refinements included a provision to keep the leading faces of the samples at the same clearance distance for both cells. Adapter ‘1’ was identified by a marking peculiarity and dedicated to only connect to Port 1 on the Network Analyzer in every case. It was then decided to make the standard clearance distance 18.6 mm from the dielectric surface in adapter ‘1’ since this would just get the sample onto the center conductor proper and off the adapter’s center conductor contacts, a logical reference point. The interface at the adapter ‘2’ end would maintain exact clearance distance by logical necessity, even though further simulations indicated this may not be an important characteristic to ensure. Another important condition met by this arrangement was assurance of center conductor release at adapter 2, not adapter 1. By always pushing the center conductor all the way onto adapter 1, leaving any free length margin at adapter 2, if the unit was held vertically, having a cell filled with sand, the adapter ‘2’ could be removed at the topside end like a lid without pulling the center conductor off adapter ‘1’. This would insure the maintenance of sand-free adapters. A call to Myat confirmed that cleaning the adapters would be difficult or time consuming, at least.

After all of this groundwork, 2 center conductors and 2 outer conductors were cut using a chop-saw, de-burred with a file, then varnish was sanded by hand down to shiny copper, all edges were carefully smoothed inside and out with emery cloth and #1 steel wool, all parts were cleaned of smudges and debris inside and out and flanges were
installed with a flat-blade screwdriver adding several millimeters of overall length. The difference in length between finished cells was determined to be about 37 mm with a

**Figure 4.7: Simulated S-parameters showing close similarity between cascade of test cell segments with and without a 1.0 mm air gap in the sample. Sample lengths are 74.0 mm in both cases, both cells are 167.0 mm long overall, and perfect geometries are assumed throughout; RRP = 2.23.**

**Figure 4.8: Close-up of data crossing points from Figure 4.7. Because both measurement systems are bi-lateral, only 4 curves are visible for 8 S-parameters.**
Figure 4.9: Block diagram used to obtain Figure 4.7 and Figure 4.8. The top system shows (left-to-right) port, coaxial transmission line, coaxial test cell filled with solid MUT, coaxial transmission line, port. The bottom system shows (left-to-right) port, coaxial transmission line, coaxial test cell filled with ½ length of solid MUT, 1 mm coaxial air-gap, coaxial test cell filled with ½ length of solid MUT, coaxial transmission line, port. In both systems the overall lengths of coaxial transmission line cascaded together are identical, as well as total sample lengths (74 mm).

dial vernier caliper. Later, multiple network analyzer measurements would find “India” to be 166.5 mm and “Juliet” to be 129.5 mm long. The center conductor was pushed onto adapter ‘1’ all the way until reaching the step, using a lid from a 32 ounce plastic jug of cranberry juice as a protective cover, since a bit of force was required against a fairly thin-wall of metal. The unit was bolted together creating the “India” cell. Eventually a system would be devised, in the absence of a torque wrench, to just lift one end of the complete system off the table surface with wrenches held “wing-on-wing” or both nearly horizontally opposed, as a poor man’s solution. Only the nut was actually turned, because it had a split-washer to prevent surface gouging of the adapter housing. In this way, repeatability approaching 0.1 mm in electrical length of the complete system was achieved. It’s important to note, that the housings had to be bolted on incrementally. No nut was given more than ½-turn until all had been given a ½-turn. Otherwise, instead of meeting with a flat interface, two flange rims providing RF contact
Figure 4.10: The mean from 3 real relative permittivity (RRP) results curves.

Figure 4.11: Standard deviation for real relative permittivity value.
could have locked up with one inside the other creating a nasty deformity after tightening the bolts. This facet of unit assembly was confirmed by a telephone call to the manufacturer.

Three full sets of S-parameters were then obtained for air, and the RRP’s calculated along with the mean and standard deviation, as shown above. Using this mean of the length difference in actual dielectric would complete the process of verifying the test set.

Network Analyzer Procedure

1. Turn Agilent 8714 ES and related PC “ON” allowing 30 minutes of warm-up time.
2. Hit the “Factory Preset” button.
3. Set averaging to “16”.
4. Hit the “Calibration→Default 2-port” button.
5. Set number of frequency points to “1601”.
6. Make sure Test Set is “OFF”.

The PC was set up to take Re/Im data while recording all settings along with an image of the magnitude trace. The S-parameters were sent into MATLAB code, which found the propagation constants and real relative permittivity values. The length difference between cells was adjusted to find the value within 0.1 mm yielding the best mean for Air over 700 MHz to 1300 MHz. This value was determined to be 36.9 mm and was rounded off to 37 mm in Figure 4.10 and Figure 4.11.

Having gotten the measurement set to work nicely in air, the next task was to prepare the paraffin samples. But, before preparing the samples, the process of protecting the
adapters was started. A long strip of Nashua 398 duct tape was cut with a razor 18.6 mm wide and used to wrap the split spring center conductor contacts of adapters 1 & 2. Next, a toroid was made of duct tape with an inner hole diameter of 35 mm and outer periphery of 75 mm, starting with two strips taped together along the common edge forming a ca. 5 mm wide seam. It was a tight fit getting this disc over the center conductor, as intended, but for the outer edge, bolting one of the cells onto the adapter left a mark where the tape needed to be trimmed. This ensured making a reasonably good seal without interfering with the coaxial geometric property of the adapter and flange bolted together. The cell was then unbolted and removed from the adapter. The tape disc was trimmed along the outer edge guided by the compression mark left by the bolted-on flange. For the next steps followed to produce the paraffin samples, refer to Figure 4.12, Figure 4.13, Figure 4.14 and Figure 4.15. The adapter labeled “number 1” (not shown) was bolted on at the bottom in the last three of these figures which firmly held the center conductor in place.

Size 300x407 tin cans, also known as No. 300, are commonly available at grocery stores as 15.25 fluid ounce containers of fruit or vegetables, although actual net weight may vary [59]. One of these cans was obtained and after removing one lid, cleaned and dried. The label was not removed. It’s important to get the kind that can be opened at either end with a common can opener. Taking a small solid center punch (not spring loaded automatic type), the center point was marked with a tack hammer, and drilled to accept a small Greenlee hole punch draw bolt. Then, the 1-3/8 inch diameter Greenlee punch draw-bolt was inserted and the larger diameter hole punched through the tin can lid. See Figure 4.16. A strip of mold sealer was separated from the main piece, wrapped loosely around the can at perimeter “a” in Figure 4.14 and subsequently rolled back and
forth on the large breadboard to make a gasket that sticks to the can. Taking the outer conductor to measurement cell “India” (166.5 mm long overall) the can was tested to see if it would go inside the 3.027 inch diameter dimension. If not, it was rolled back and forth on the breadboard some more, until finally fitting inside the test cell with minimal clearance to eliminate sticking. This gasket now centered the can with the punched hole uppermost, as in Figure 4.14. The open bottom end of the can was moved sufficiently upward into the 3-1/6 inch outer conductor to ensure non-interference with ‘adapter 1’. The distance from the top of the rim on the top flange to the closest rim at the top of the can was measured and found to be about 40 mm. A special long center conductor, about 225 mm in length was carefully prepared, and installed all the way to the stop on the center conductor of ‘adapter 1’. ‘Adapter 1’ was bolted onto the bottom flange, to make sure the center conductor would clear the hold punched into the tin can upper lid. Then, ‘adapter 1’ and center conductor were removed and placed to one side. Two more strips of mold sealer were separated off, and rolled by hand on the bread board to make them grow about 2-1/2 times longer. Each one of these thinner strips would now encircle
Figure 4.12: The Type-N to 3-1/8 inch E.I.A. adapter manufactured by Myat, Inc. (reprinted with permission).
Figure 4.13: Isometric cut-away view of the 300x407 tin can placed inside a section of 3-1/8 inch E.I.A. transmission line to begin the formation of a paraffin mold.
Figure 4.14: The paraffin mold, see text for explanations.
Figure 4.15: After all other preparations are in place, pour RTV mixture to within 40 to 42 mm of the top of the mold cylinder. Allow to cure overnight (or 8 hours) then pour the paraffin. See text for complete details.
the can, see Figure 4.16. With the two elongated strips of mold sealer and the blunt ends of skewer sticks water-tight seals were formed at peripheries marked ‘b1’ and ‘b2’ in Figure 4.14. The purpose of making 2 seals, was to ensure no RTV or paraffin reached ‘adapter 1’. Next, install the “9” Fit-All S/J Washer o-ring on the center conductor positioned so that it will be just below the hole in the can lid when everything is bolted up, see Figure 4.17, and “c” in Figure 4.14. Bolt ‘adapter 1’ onto the cell and stuff
the center conductor full of cotton balls to prevent debris from getting inside ‘adapter 1’. Mold sealer strips were separated as needed and seals added around the center conductor and tin can periphery from the top. Mold sealer was molded by hand using mostly finger-tips. The finished assembly was placed in the Black and Decker Workmate 525 and leveled using mold sealer rolled into balls as supports at 3 points. See Figure 4.17.

AereoMarine 25 Silicone Moldmaking Rubber, available from John Greer & Associates, was mixed in sufficient quantity to generate about $1/10^{th}$ quart of mixture. This was poured into the paraffin mold until reaching about 40 millimeters from the top of the flange rim of the cylinder. See Figure 4.15, Figure 4.17 and Figure 4.18. This mixture was allowed to cure overnight. In the morning, GulfWax® paraffin was heated to 160 F on a kitchen stove. If the pour temperature was too high, the paraffin would shrink excessively, leading to a large air-gap around the sample. If the pour temperature was too low, the sample could have been difficult to remove from the mold, or in-homogeneities or fractures could have formed inside the sample. A pour pitcher was placed inside a 4-quart pan containing about a quart of water, or more, sitting on a stove burner. An oven thermometer sensor was situated inside the pour pitcher. Four quarter-pound sticks of GulfWax® paraffin were added to the pour pitcher. The stove was turned “ON” and adjusted so the heat would settle on 160 F melted paraffin temperature, steady-state. The paraffin was poured in 2 stages to reduce the circular groove which forms on the surface as a result of cooling. After the first pour a skin forms. As soon as this skin became thick enough it was poked through 12 or 13 times with a bamboo skewer [60]. When the entire block of paraffin reached a temperature of around 115 F the final 1 cm was poured, filling all the poked holes. Sixteen ounces equate to 453.6 grams and the first toroid came
out a couple mm over length at 127 grams, so it takes 0.28 pounds and therefore one 4-
ounce stick of paraffin may not be quite enough for one sample.

The mold was removed from the freezer and the center conductor pulled out from the
top. If the RTV has enough grip on the outer conductor inner wall, the O-ring will slip
off, leaving the mold intact. By rotating a little back and forth, remove the paraffin
sample off the top end of the center conductor. Measure the length of the sample with a
dial vernier caliper.

To remove any depressions that still exist in the sample, compromising the flatness
of one or both faces, push the center conductor back into the sample, to a depth of 39 or
40 mm. Heat an empty pour pitcher to 160 F, and with rotating motion, melt the grooved
face of the paraffin sample flat down to the metal inner conductor. The sample should
now be about 1 mm too long. Both measurement cells were stood on end, allowing the
alignment pins to dangle over the edge of the table top, and 37.0 mm difference verified.
The sample was then trimmed to 37.0 mm using a razor blade, see Figure 4.19 [61].

Figure 4.18: (left) RTV gasket. (right) poking holes in the paraffin.
A paper towel spool was then very carefully cut with a razor blade forming a cylinder as close to 18.6 mm in length as could be achieved. This cylinder was used as a stop at the center conductor contacts of ‘adapter 1’. With cell “India” assembled, the ordinary paraffin sample was pushed onto the center conductor to just touch this stop, and the precision 37.0 mm long paraffin sample was then added to the center conductor just touching the first sample. The samples were rotated to form the best fit making the smallest possible gap between them. Each sample had a good end, formed from contact with the RTV, and a poor end which was the top during the pouring process. The good ends were oriented as the inter-sample interface, further minimizing any gap effects. See Figure 4.20 for the results. Accepted RRP values are 2.25 at 100 Hz through 10 MHz and 2.22 at 3 GHz through 10 GHz [49].
Figure 4.20: RRP for GulfWax®.

Figure 4.21: Measurement error for GulfWax® compared to von Hippel’s results [49].
The accuracy degrades between 1.0 and 1.5 GHz which corresponds to the “best” return loss in the S-parameters for the longest test cell “India”, see Figure 4.22 and “Juliet”, see Figure 4.23. Using shorter samples or more than 2 lines should fix this problem. One of the main factors in getting these results was in adjusting the pour temperature to 160 F. At higher temperatures the paraffin sample would fall out of the mold without cooling off in the freezer for an hour. Paraffin shrinks when it cools and at 160 F pour temperature it shrinks just enough to remove the sample without damage after 1 hour in the freezer at 0 F.

An observation about GulfWax®, in 1954 when von Hippel’s work was published [49], this brand of paraffin was made by Gulf Oil Company at the Port Arthur Refinery [62]. A.R. von Hippel reports a melting temperature of 135 F. Today, GulfWax® is made by Exxon Mobil at the Baton Rouge Refinery, and melting temperatures on the MSDS are listed as 130 F maximum. See the MSDS in Appendix C. Actual measured values for a sample of 8 120,000 pound freight-car loads shipped during 2007 varied from 125.9 F to 126.7 F. Corresponding density values were not available.

Noise in the real relative permittivity values near d.c. is caused by the very small sample length expressed in wavelengths. If there are only fractions of a degree of electrical phase difference in the measurements, the value becomes dominated by small variations in the network analyzer data. This error source could be overcome by using longer differences in sample length or longer measurement times. The increase in error at frequencies which are well-matched in the long test cell, are due to resonance effects in the sample. When the sample electrically becomes ½-wavelength long the line no longer sees the sample as well. A solution would be to use shorter sample lengths or multiple
lines (more than two) [53]. It’s worth noting that this matching effect occurred only in the longer of the two lines, with the longer of the two samples.

Figure 4.22: $|S_{11}|$ for the long test cell.

Figure 4.23: $|S_{11}|$ for the short test cell.
Because these two extremes have conflicting solutions, longer versus shorter lines and samples, the length difference of 37.0 mm was chosen as a suitable compromise.

4.3 **Characterization/verification measurements in the laboratory**

**using a network analyzer**

Having finished testing the measurement system itself, to show that it was working properly, it then became possible to compare soil and artificial dielectric measurements directly in the laboratory. With soil of similar specific gravity and moisture content as that used in the Geo-centers Tests and artificial dielectric of the same composition, the match that was observed in the time domain reflections of the voltage signals could now be obtained as impedance versus frequency plots. There were a few related plots to obtain beforehand, for further proving of the system under real-world conditions.

The block diagram of the measurement system in the soil laboratory is as shown in Figure 4.24. Calibration of the system was not required, the procedure followed is also given below.

**Measurement Procedure for the Soil Laboratory System in Figure 4.24:**

Agilent 8714ES

1. Hit Preset. This should reset the network analyzer to the “Factory Preset” condition.
2. Hit the Cal\textsuperscript{Æ}”Default 2-port” buttons
3. Turn Averaging “ON” / factor=’16’
4. Menu\textsuperscript{Æ}Number of Points\textsuperscript{Æ}’1601’
5. System Options → System Config → Switching Test Set → Multiport ‘OFF’

![Diagram of system setup](image)

**Figure 4.24:** System used to obtain full s-parameters of two transmission lines of different lengths containing MUT.

Desktop PC
1. Log-in to the user account.
2. Launch Excel
3. Connect to the Network Analyzer over the LAN by clicking on the “Agilent” button in the bottom tray.
4. Select the Agilent 8714ES instrument.
5. Make sure to set up for getting all data with all measurement settings (all boxes checked). Make “Re/Im” ‘ACTIVE’.
6. “Get Data”
7. Save the data file to the memory stick.
8. after getting all 8 complex s-parameter data files, move the memory stick over to the laptop PC for data processing in MATLAB.
The 10 mm diameter hollow polypropylene spheres themselves were of interest, for these were not treated by the Lam equations, per se. These were measured and permittivity curves plotted, see Figure 4.25. For this measurement, there was no barrier of paraffin. The first cell was filled up to the top of the center conductor, and the fill margin measured with a dial vernier caliper. Four S-parameter measurement files were taken, then the 2nd cell filled to the same margin measured from the top surface of the RF lip and the other 4 files obtained. Working cutoff for H.O.M.’s is 1600 MHz in air.

Dry sand was next. For this material, an RRP value of 2.55 over 10 MHz through 3 GHz is reported [49]. Before preparing the sample, the test cell itself had to be readied. The precision 18.6 mm long stop was placed over the center conductor contacts of the bottom adapter. It was allowed to simply rest on the dielectric face of the adapter. Next, the copper center conductor was forced onto the contacts all the way, using a cap from a 32, 48 or 64 ounce plastic juice container at the far end as protection from the end of the thin-walled metal tube. One of the paraffin toroids was then inserted, rotating first to find a good fit to any out-of-roundness in the center conductor dimensions to minimize unwanted shaving of paraffin from the center hole. The paraffin barrier was then dropped onto the precision positioner over the center conductor, with the flatest side up. This is the side that would contact MUT. Using a flashlight and a couple dozen 6 inch applicators available at medical supply stores, Vaseline® was then applied to both gaps between the paraffin and the copper. After both gaps were sealed, verified by visual inspection, as much of the Vaseline® was removed using clean applicators as could be easily accomplished. In this way a very thin protective layer of Vaseline® was left behind to serve as a barrier to MUT dust or grains. Any MUT getting into either adapter could
cause problems by moving around between measurements affecting the line impedance, and could be very difficult to remove.

In preparing the sample, first dry sand was poured into one of the coaxial test cells and tapped repeatedly to try to establish a settled condition. Fill margin values were monitored with a dial vernier caliper. When the settled sand reached a predefined level the upper adapter was bolted on and measurements taken. Then, the top adapter was unbolted, and removed from the test assembly. The sand was poured out and weighed on a scale to find the mass. The volumes of both cells were then calculated, for the given fill margin. The mass of sand needed for the second coaxial test cell was determined based

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**Figure 4.25:** The RRP of 10 mm diameter hollow polypropylene balls poured into test cells made from two different length, vertically oriented, 3-1/8 inch E.I.A. coaxial lines leaving equal length air-gaps at the top.
on these volumes and lengths for the given settling state. After the correct mass of sand was poured into the 2nd test cell, it was tapped until the fill margin matched the previous cell’s value. In this way fill margins were held constant between cells, as well as the 37.0 mm MUT length difference between cells. That portion of the MUT which was held constant between cells was generally kept at 30.0 mm although this number could have varied somewhat without any negative effects as long as it was kept the same for both cells. Since the environment was dusty and harsh for delicate instruments, it was advisable to use the cheapest dial vernier caliper and scale that would suffice. See Figure 4.26.

Figure 4.26: The measured RRP of Dry sand.

Dry soil was characterized, next, using a similar procedure to dry sand. See Figure 4.27. One additional step was the drying of the soil. This was conducted in disposable
aluminum baking pan atop a kitchen stove set to the lowest setting that would guarantee boiling water. Steam was observed. Periodically, the soil was stirred. After 1 hour the dry specific gravity was determined. The soil was then placed back on top of the hot stove and allowed to dry for another hour. The new dry specific gravity was determined and compared to the previous value. These were within 5% of each other. It was decided to use the original number, and to use 1 hour on top of the stove for any drying requirements. The difference was close to or within the other errors involved in the measurements, such as scale readings and the length difference between coaxial test cells.

![Figure 4.27: Dry Parkis Mills Topsoil, sifted through a Number 4 sieve used for soil science. Specific Gravity: 1.04 g/cm³.](image)

Figure 4.27: Dry Parkis Mills Topsoil, sifted through a Number 4 sieve used for soil science. Specific Gravity: 1.04 g/cm³.
Next, soil at bag moisture was characterized. This was not from a fresh bag, but from one that had been opened, then re-sealed. After pouring through a Number 4 sieve the wet soil was measured using the procedure for dry sand. Then, it was placed in a 1 gallon sealable plastic bag and taken to a kitchen with a stove, where it was dried and specific gravity determined using the procedure described for dry soil, above. See Figure 4.28.

![Figure 4.28: MUT: Parkis Mills Topsoil / dry specific gravity 0.866 / dry moisture 12.1%](image)

**Figure 4.28: MUT: Parkis Mills Topsoil / dry specific gravity 0.866 / dry moisture 12.1%.**

As a check of this result, comparison was made with previous published results [48]. See Figure 4.29.

At Geo-centers, matching behavior was observed for artificial dielectric with packing fractions of 0.125 and 0.286, so both of these were characterized in the coaxial test cells
using the same paraffin barrier and cardboard spacer as in the sand and soil measurements described, above. There was one additional constraint on the sample preparation. As was observed at Geo-Centers, the polypropylene spheres and conductive beads could not just be dumped into the cell in a haphazard manner. In this case, they were placed a few at a time to ensure uniformity of bulk properties throughout the sample volume. They were layered, by first placing a layer of plastic spheres, then a layer of beads. If the layers were thin enough, the bulk properties should not be affected. See Figure 4.30.

**Figure 4.29:** Comparison between measured and published RRP values versus dry soil density.
After finding a propagation constant for artificial dielectric, the procedure for obtaining RRP was altered by non-unity of the real relative permeability. The value for $\mu_r$ was taken from the Lam equations.

$$\varepsilon_r = \frac{\beta^2}{\mu \beta_0^2}$$  \hspace{1cm} (4.22)

Figure 4.30: Artificial Dielectric being built up inside a coaxial test cell. Notice the cotton balls and 64 ounce plastic jug cap that were used to ensure no foreign matter entered the lower adapter (adapter 1, out of view). (left) A thin layer of conductive beads. (right) The last layer did not always come out consistent with a 37.0 mm length difference between MUT’s as in this case where a fractional layer was necessary. The plastic spheres were pressed down into the artificial dielectric before taking any data.

Having found the RRP’s for artificial dielectric of 2 packing fractions, the results were plotted. See Figure 4.31 and Figure 4.32.

The respective intrinsic impedances were then determined using

$$\eta_i = \sqrt{\frac{\mu}{\varepsilon}}$$  \hspace{1cm} (4.23)
where $\mu$-values for artificial dielectric was again determined using the Lam equations. These impedances are all plotted in Figure 4.33 and the corresponding reflection coefficients in Figure 4.34 verifying the Geo-centers Tests. At that time matching behavior was observed for both packing fractions, with $p=0.125$ producing better results.

![Characterization of Artificial Dielectric (p=0.125)](image)

**Figure 4.31:** Artificial Dielectric with packing fraction (p) equal to 0.125.

![Characterization of Artificial Dielectric (p=0.286)](image)

**Figure 4.32:** Artificial Dielectric with packing fraction (p) equal to 0.286.
Figure 4.33: Relative impedance versus frequency for soil similar to Geo-centers soil showing a good match to artificial dielectric.

Figure 4.34: Reflection Coefficients for A.D./Soil interfaces. Angles of incidence / transmission of 0 degrees assumed. A.D. 1 (p=0.125); A.D. 2 (p=0.286); soil is from Figure 4.28.
5. CONCLUSION

A lightweight artificial dielectric was developed which was used to fill in the depressions and trenches dug into a soil bed at an indoor radar test facility. The voltage versus time signal return data for the roughened soil with the artificial dielectric fill provided a better match to the signal returns for flat soil before roughening, than the signal returns for roughened soil without any artificial dielectric compared with undisturbed soil. The radar’s principle energy band covered 700 MHz to 1300 MHz.

A measurement system was devised for characterizing the materials properties of soil and artificial dielectric in the laboratory using a network analyzer. These measurements showed a reflection coefficient magnitude less than ‘0.1’ across the frequency band 700 MHz to 1300 MHz for the best case match between artificial dielectric and soil that were as similar as could be readily prepared to the original artificial dielectric and soil, assuming vertical incidence illumination by the radar.

For comparison, at an air/soil interface with a real relative permittivity of ‘3’ for the soil, the magnitude of the reflection coefficient would be ‘0.5’. Since energy goes by the squares, such a value would mean 25 times more energy reflected than an interface yielding a reflection coefficient of ‘0.1’. Therefore, the principles and methods explained here in this paper, and the results achieved open the door to reducing unwanted energy in the RADAR return signal to 1/25\(^{th}\) of the raw value.

Qualitatively, A.D. was shown to make moderately rough ground look like level soil under field conditions re-created in the laboratory, by visual inspection of the return signals and comparison of correlation coefficients.
REFERENCES


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[50] CST GmbH, Darmstadt, Germany.


APPENDIX A: ADDITIONAL WORK

A donation of discarded waveguide and coaxial components was received, including adapters and straight sections of WR284 waveguide with round flanges, larger straight sections, possibly L-band, and several type-N coaxial hard-line 10 dB isolators about 20 cm long between the connectors.

By November 2003, good simulation results using CST Microwave Studio® [50] had been obtained for a lattice of PEC spheres using the Nicholson-Ross-Weir algorithm [63][64][65]. The results were slightly lower than the analytical solution predicts, consistent with a small air gap around the sample to prevent short-circuiting against the conducting rectangular waveguide walls. WR284 dimensions were used for this simulation, and the formulas from Lam [42]. Above real relative permittivities of about 1.5 the beads interact with each other [66]. This interaction is accounted for in the Lam equations. Nicholson-Ross-Weir does not produce a unique solution because of the complex argument in the logarithm, but if the sample length is less than 180 degrees electrically, the principle value should be used. See equations (0.1) – (0.7).

\[ \Gamma_1 = X \pm \sqrt{X^2 - 1} \quad (0.1) \]

where \( \Gamma = \) reflection coefficient, the correct root is chosen by requiring \( |\Gamma_1| \leq 1 \).

\[ X = \frac{1 - V_1 V_2}{V_1 - V_2} \quad (0.2) \]

\[ V_1 = S_{21} + S_{11} \quad (0.3) \]
\[ V_2 = S_{21} - S_{11} \]  
\[ z_1 = \frac{S_{11} + S_{21} - \Gamma_1}{1 - (S_{11} + S_{21})\Gamma_1} \]

where \( z_1 \) is the transmission coefficient.

\[ \mu_r = 1 \quad \mu_e = 1 \]

\[ \varepsilon_r = \frac{\lambda_0^2}{\mu_r} \left[ \frac{1}{\lambda_c^2} - \left[ \frac{1}{2\pi L} \ln \left( \frac{1}{z_1} \right) \right]^2 \right] \]

Figure 0.1: Conducting PEC beads arranged in an FCC lattice. Volume Fraction (p) equals 0.175.

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The NRW approach generally uses a TRL calibration. The initials stand for “Thru-Reflect-Line” [67]. The “through” is a zero length line, or connecting the reference planes together. The “reflect” is usually a short circuit. It’s good to use the same short circuit, first at one reference plane then the other. The “line” is a length of transmission line of known line impedance from 30 degrees to 150 degrees over the frequency band of interest [68].

A parallel-strip TEM waveguide with type-N adapter was constructed out of wood and copper flashing. [69] – [70]. Input jacks were type-N female connectors. An Agilent E8358A PNA was made available for use. Home-made TRL standards were planned. The
project was abandoned after a couple weeks of intensive effort because of difficulties getting the home-made flanges to work. It also wasn’t clear if it would even be possible to make TRL standards without access to a proper machine shop and materials.

A material dubbed “lossless heavy clay” with $\varepsilon_r = 6.6$ and no losses was successfully used to get good 2-Line method results in a simplified simulation, see Figure 0.3 and Figure 0.4. The simulation did not include Type-N jack to rectangular waveguide adapters. Measurements of two clay-filled WR284 adapters connected together revealed a lot of resonance effects, and was determined to be unusable for making measurements of dielectric samples.

![Figure 0.3: Cut-away view of WR284 waveguide filled entirely with “lossless heavy clay”. The center section comprises the MUT. Good 2-Line method results were obtained for this arrangement over ca. 1 GHz to 1.5 GHz.](image)
Figure 0.4: Results using the 2-Line method in WR284 with simulated s-parameter inputs for “lossless heavy clay” of RRP = 6.6.

A quick gap-study was carried out in simulation, again using WR284 waveguide dimensions. The MUT was homogeneous (not A.D.) with RRP equal to 6.6. As expected,

Figure 0.5: WR284 simulation geometry with darker blue-green MUT shorter of two samples of $\varepsilon_r = 6.6$ used with the 2-Line method results in Figure 0.6. The lighter blue colored material is free-space. The a-dimension air gaps appear to the right and to the left of the sample. Not seen in this view, the x/y-directed surrounding space is PEC (Perfect Electric Conductor).
Figure 0.6: Simulated RRP for a homogeneous MUT specified as 6.6 with two 3-percent gaps in the a-dimension. The centered sample brick width is 0.94 of the waveguide a-dimension. There is no discernable harm done to the result by the presence of these gaps over ca. 0.8 GHz to 1.5 GHz.

Figure 0.7: WR284 simulation geometry with darker blue-green MUT shorter of two samples of $\varepsilon_r = 6.6$ used with the 2-Line method results in Figure 0.8. The lighter blue colored material is freespace. The b-dimension air gaps appear above and below the sample. Not seen in this view, the x/y-directed surrounding space is PEC (Perfect Electric Conductor).
Figure 0.8: Simulated RRP for a homogeneous MUT specified as 6.6 with two 3-percent gaps in the b-dimension. The centered sample brick width is 0.94 of the waveguide b-dimension. The error caused by the presence of these gaps is around 0.20 by visual inspection over ca. 0.9 GHz to 1.75 GHz.

A three-percent gap in the b-dimension caused much greater decrease in resultant RRP than a three-percent gap in the a-dimension of the sample. The latter case had very little effect.
Figure 0.9: 5 mm diameter conducting beads arranged in a simple cubic (sc) lattice. If the beads extended to infinity in all directions with this uniform lattice, ‘p’ would equal 0.2325.

Figure 0.10: RRP for sc lattice of 5 mm diameter conducting beads shown in Figure 0.9 with p=0.2325 if the lattice were extended to infinity with no gaps. This is the only case in this paper requiring both propagation constants to get a good result with the 2-Line method.
A question that arises naturally at this point, is what would happen if this lattice was curved around to fit a coaxial cross-section? This question was investigated, see Figure 0.9, Figure 0.10, Figure 0.11 and Figure 0.12. Starting with the lattice dimension between bead centers, which was uniform in the X/Y/Z directions for the rectangular lattice, it is possible to generate a sample suitable for a coaxial cell. The lattice edge dimension becomes the difference in radii, and the difference between bead centers along the arc formed by the radius for that “string”. A “string” of beads shares a radius. In this example, there are 3 strings.

Figure 0.11: The simple cubic lattice from Figure 0.9 transformed into a coaxial lattice.
**Figure 0.12:** Transforming the rectilinear simple cubic lattice into a coaxial lattice increases the RRP slightly over 2.078 GHz to 3.0 GHz. The dashed line represents the specified value.

The most promising simulations of the rejected approaches used waveguide below cutoff to get real relative permittivity from TE_{10} with the 2-Line method. Simulation settings included open boundaries at the ends of the waveguides. These could be replaced by horn antennas operating in their normal range which would be below cutoff in the feed waveguide. The entire system would have to be positioned inside an anechoic chamber in an empirical effort in order to get the same result. The adapters might have to be optimized for the insertion of below-cutoff energy, and the walls of the waveguides might have to be silver-plated. Cryogenic cooling might be needed to reduce all resistive losses. It was at this point that the search for a method was shifted from the waveguide approach, to a coaxial cable approach.
Figure 0.13: $\text{TE}_{10}$ in MUT of real relative permittivity 2.5 has a cutoff frequency of 1.3142 GHz in WR284 waveguide.

Figure 0.14: Boundary conditions showing that $Z_{\text{min}}$ and $Z_{\text{max}}$ were set to “Open” boundaries. The results in Figure 0.13 were obtained with this geometry.
Figure 0.15: Endless waveguide would solve the problem of system resonances caused by commercially-made Type-N to WR284 waveguide adapters used below cutoff. This is how the geometry in Figure 0.14 would look if one could see the open boundaries at the ends of the adapters.

Figure 0.16: As an alternative to endless waveguide, loads could be devised to match the energy below cutoff. This could be a very complicated task, as the impedance is frequency-dependent and existing loads are designed with a different frequency-dependency in mind. This is one possible approach to realizing the infinite waveguides shown in Figure 0.15.
Figure 0.17: This waveguide is surrounded by perfect electric conductor (PEC) to infinity in the +/- X/Y/Z directions. Therefore, this represents the measurement set-up using commercially available adapters with shorting end-plates. Poor quality measured results with this type of set-up are shown in Figure 0.18.

Figure 0.18: Actual realization of a materials properties characterization system with commercially-made Type-N to WR284 adapter data show unusable RRP obtained from the measured S-parameters. Pure TE_{10} in the paraffin sample covers 1.3853 GHz through 2.6333 GHz based on the recommended fundamental operating range divided by the square-root of 2.25. Similar poor results were obtained from simulations based on actual geometries. This approach worked in one paper [51][52][53].
APPENDIX B: PARTS LISTS, EQUIPMENT LIST AND LIST OF SOURCES

Measurement Set Parts List

QTY 1, 3-1/8 inch outer diameter hard-drawn copper coaxial transmission line inner and outer conductors with minimum lengths of 175 mm; Myat P/N 301-004-000 $30.00/ft + $110.00 base price 3-1/8" transmission line assembly, unflanged, customer specified length

QTY 2 3-1/8 inch to Type-N (female) coaxial adapter including hardware (3/8-16 bolts, split washers, nuts, QTY 6 sets, all stainless steel) Myat P/N 301-059

QTY 1 center conductor from Myat P/N 301-004-000, 225 mm long, or slightly shorter, well de-burred and polished especially on the outside, dedicated for use in the mold, only.

QTY 4 flanges, Myat P/N 301-014

Electronics Laboratory Equipment List

No. 4 soil scientist’s sieve

6 inch Puritan® applicators, cotton tipped, non-sterilized

Agilent Industries 8714ES Network Analyzer, Type-N Receptacle Ports

Agilent Industries 85032F Type-N Calibration Kit

Cardboard spool from Paper towels

Coddington Magnifier

2 combination wrenches (open-end & box-end) for 3/8-16 bolts

1 coaxial cable 36 inches long, type-n (male) at each end

1 coaxial cable 18 inches long, type-n (male) at each end

QTY 4 plastic protective caps for Type-N (female) connectors.
QTY 2 plastic protective caps for Type-N (male) connectors.

Data card to connect the Agilent 8714 ES with the PC for data taking

Dry sand

GulfWax® “Household Paraffin Wax” (UPC: 0-62338-00972-8) at the local hardware store canning section, distributed by Royal Oaks Enterprises, manufactured by Exxon Mobil Corporation, Irving, TX, USA.

Isopropyl alcohol with no oils or other additives

LAN availability

QTY 2 Microwave Distributors Midisco Type-N Keepers (NN-MF)

Memory stick

Parkis Mills Topsoil

PC op/sys Windows (with internet connection)

Q-tips

Software:
  CST STUDIO SUITE™ 2008
  MATLAB® Student Version 7.0.1
  Microsoft® Office 2003
  Agilent Technologies’ I/O Card software

Vaseline®

Workshop Parts List

4-quart pan

50/50 lead/tin solder

600 ml bulk caulking gun with re-usable dispenser bags used for caulking with 6 mm to 35 mm bead size polyurethane forcing cone manufactured by Cox. Allglassparts Part Numbers: Gun:: 1-C08-00121; Cone:: 1-C07-00100

“9” Fit-All S/J Washer / Danco Company / Concordville, PA 19331 (UPC: 0-3715-36655-2)
AeroMarine 25 Silicone Moldmaking Rubber (RTV)

Black and Decker Workmate® 525

Boxwood clay shaping set of tools, 6 inches long, Dick Blick Art Materials P/N 30304-1069

Combination square - 12 inch Sears P/N 39551, or similar to 12 inch Stanley Hand Tools P/N 46-222

Can opener

Candy Thermometer, digital

Crisco® Pure Vegetable Oil, ingredients: soybean oil, 48 [fluid ounces] (UPC: 0-51500-25362-5)

Dial Vernier Caliper

Drills, set, metal cutting

Electric drill with ¼ inch keyed chuck

Emery cloth

File

Flat-blade screwdriver

1-3/8 inch diameter Greenlee Punch (for sheet metal), Allied P/N 799-3007, Greenlee Type 730BB-1-3/8

Household Freezer operating at 0 F

Isopropyl Alcohol without oils or other additives

Jumbo Cotton Balls (QTY 100) , Walgreens (UPC: 3-11917-06333-1)

Kitchen Klassics 100 Bamboo Skewers 10 inch p/n 171212 (UPC: 0-90782-24220-0 )

Kitchen stove


Large wooden cutting board ( 18” x 24” )

Machinists Rule
Mold Sealer by Yaley Enterprises, No. #110281 (UPC: 0-52124-10082-1)

Nashua 398 Industrial Grade Duct Tape; or, Henkel Duck Tape (UPC: 0-75353-03660-0)
rated for use at temperatures up to 200 F

Number 300x407 tin can (clean, dry & empty with one lid intact and the other removed –
both ends should be of the same type)

Ocean Spray® Cranberry Juice Cap (32 ounce container)

Pam, non-stick spray

Paper Towels

Propane torch (for soldering copper water lines)

QTY 2 Pour Pitchers

Razor blades

Scissors

Soehnle/bretagne Art.-Nr. 67060 kitchen scale, mass to nearest gram, (German Bar-code: 4-006501-670601)

Steel punch (not spring loaded)

Steel wool

Tack hammer

Tacro 3601 M Drafting Compass

List of Sources

Agilent Technologies, Inc., 5301 Stevens Creek Blvd., Santa Clara, CA 95051

All Glass Parts, Inc. 18139 – 107 Avenue, Edmonton, Alberta, T5S-1K4 Canada; (780) 487-4888; http://www.allglassparts.com/contact/ {accessed on 16 February 2008}

Allied Electronics, Inc., 7151 Jack Newell Blvd. S.; Fort Worth, TX 76118; (817) 595-3500, (800) 433-5700 (catalog #408)

Aubuchon Hardware, Maynard, Massachusetts; (978) 897-2500
Byrne Home Health Center, 16 Main St., Natick, MA 01760; (508) 655-3656

C-Lec Plastics, Inc., 6800 New State Road, Philadelphia, PA 19135-1535

Chomerics North America, Parker Hannifin Corp., 77 Dragon Court, Woburn, MA 01801; (781) 935-4850

Dick Blick Art Materials, 401 Park Drive, Boston, MA 02215; (617) 247-6660

Federated Foods, Inc., 3025 W Salt Creek Ln., Arlington Heights, IL, USA; (847) 577-1200

Harvey’s Hardware Company, 1004 Great Plain Ave., Needham, MA 02492-2508; (781) 444-4515

Home Depot, 339 Speen Street, Natick, MA 01760; (508) 647-9600

John Greer & Associates, 4128 Napier St., San Diego, CA 92110; (877) 342-8860

Karstadt, Darmstadt, Germany

McMaster-Carr, 6100 Fulton Industrial Blvd. SW, Atlanta, GA 30336-2853

MSC Industrial Supply, Worcester, MA

Michaels, 321 Speen Street, Ste G, Natick, MA 01760-1506; (508) 655-2225

Microwave Distributors

Myat, Incorporated, 360 Franklin Turnpike, Mahwah, NJ 07430; (201) 684-0100

Peak Candle Supplies, Colorado (technical support)

Penn Engineering; 12750 Raymer St., North Hollywood, CA 91605

RCBS Operations, 605 Oro Dam Blvd., Oroville, CA 95965, (800) 533-5000

Robinson Ace Hardware, 1 Nicholas Road, Framingham, MA 01701; (508) 877-1888

Roache Brothers, Natick, MA

Sigma-Aldrich Corporate Offices, Sigma-Aldrich, 3050 Spruce St., St. Louis, MO 63103

Small Parts, Inc., 15901 SW 29th Street / Suite 201, Mirmar, FL 33027

Stan Rubinstein Associates, Inc., Foxboro, MA

Wal-Mart Stores, Inc., Bentonville, AR 72716-8611

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Walgreens, Inc., 200 Wilmot Road; Deerfield, IL 60015; (847) 914-2500

Western Test Systems, Inc., 2701 Westland Court, Unit B; Cheyenne, WY 82001
APPENDIX C: GulfWax® MSDS

MATERIAL SAFETY DATA SHEET

Royal Oak Enterprises, Inc.
1 Royal Oak Avenue
Roswell, GA 30076
(678) 451-3200

713-870-6884 for non-emergency
1-800-726-3015 for medical emergency
1-800-424-9300 for transportation emergency
901-395-0119 for technical information during business hours

NPPA Hazard Rating
Health 0
Flammability 1
Reactivity 0

SECTION 1 MATERIAL IDENTIFICATION

Name: Gulfwax Household Paraffin Wax

SECTION 2 INGREDIENT INFORMATION

<table>
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<tr>
<th>Chemical/Common Name</th>
<th>CAS No.</th>
<th>percent by Weight</th>
<th>Exposure Limits</th>
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<tr>
<td>Paraffin wax</td>
<td>64742-43-4</td>
<td>100</td>
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SECTION 3 HAZARDS IDENTIFICATION

Signs and Symptoms of Overexposure: Fumes from overheated paraffin may be mildly irritating to nose, throat, and eyes.

Potential Health Effects:
- Eye Contact: Not an eye irritant.
- Skin Contact: Not a primary skin irritant.
- Ingestion: Not orally toxic.
- Inhalation: Not toxic by inhalation.

Chronic Effects: The product has been tested as a whole for chronic effects. No ingredient is listed carcinogen (NTP, IARC, OSHA).

SECTION 4 FIRST AID MEASURES

Eye Contact (paraffin fumes): Flush with water. Call physician.
Skin Contact: Melted paraffin wax can cause thermal burns. Flush with cold water to cool affected area and contact physician immediately.
Ingestion: No specific procedure required.
Inhalation (paraffin fumes): Remove to fresh air. Call physician.

SECTION 5 FIRE FIGHTING MEASURES

The information contained herein is based on data considered accurate. However, no warranty is expressed or implied regarding the accuracy of this data or the results to be obtained from the use thereof.

Page 1 of 4
Flammability Properties
Flash Point: 385 °F (Penney-Martens Closed Cup)
Extinguishing Media: CO2, dry chemical foam.
Fire Fighting Instruction: Cool exposed containers with water. For large fires or fires in confined spaces, self-contained breathing apparatus is required.

SECTION 6 ACCIDENTAL RELEASE MEASURES
In Case of Leakage or Spillage: Allow to solidify if liquid. Sweep or scoop up.

SECTION 7 HANDLING AND STORAGE
Always melt by heating in a pan over boiling water, as in a double boiler. Never melt directly in pan over fire, hot plate, or in hot oven. Wax can catch fire if it is overheated. Do not heat above the flash point. Do NOT heat with open flames or open electrical coils. Acceptable heat sources are heat transfer fluids (steam, hot water, oil) or sealed electrical resistance heaters.

SECTION 8 EXPOSURE CONTROLS/PERSONAL PROTECTION
Ventilation Requirements: General ventilation.
Personal Protective Equipment: Chemical resistant gloves with thermal protection when working with melted paraffin.

SECTION 9 PHYSICAL AND CHEMICAL PROTECTION
Appearance: Colorless waxy solid
Odor: Nearly odorless
Boiling Point: Not applicable
Freezing Point: Not applicable
Solubility in Water: Insoluble
Vapor Density (Air = 1): Not determined
Specific Gravity: 0.85220/20°C (typical)
Acute Oral LD50: Not applicable
Acute Inhala. LC50: Not applicable

SECTION 10 STABILITY AND REACTIVITY
Storage Stability: X Stable
Unstable Conditions to Avoid: None known
Hazardous Polymerization: May Occur X Will not occur
Conditions to Avoid: None known
Incompatibility (Materials to avoid): None known
Hazardous Decomposition Products: CO2

SECTION 11 DISPOSAL
Proper Disposal Method: Discard with trash collection

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SECTION 12 ADDITIONAL COMMENTS

CERCLA (Superfund®) Reportable Quantity: Not applicable
SARA Section 313 Chemicals: None

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### PARAFFIN WAX

#### SPECIFICATIONS

<table>
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<tr>
<th>TEST DESCRIPTION</th>
<th>METHOD</th>
<th>MINIMUM</th>
<th>MAXIMUM</th>
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