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Measurement of Dielectric and Magnetic Properties of Soil

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Measurement of Dielectric and Magnetic Properties of Soil

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Abstract

The possibility of subsurface imaging using SAR technology has generated a considerable amount of interest in recent years. One requirement for the successful development of a subsurface imaging system is an understanding of how the soil affects the signal. In response to a need for an electromagnetic characterization of the soil properties, the Radar/Antenna department has developed a measurement system which determines the soils complex electric permittivity (ϵ) and magnetic permeability (μ) at UHF frequencies. The one way loss in dB is also calculated using the measured values of μ and ϵ .

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Introduction

The possibility of subsurface imaging using SAR technology has generated a considerable amount of interest in recent years. One requirement for the successful development of a subsurface imaging system is an understanding of how the soil affects the signal. In response to a need for an electromagnetic characterization of the soil properties, the Radar/Antenna department has developed a measurement system which determines the soils complex electric permittivity (ϵ) and magnetic permeability (μ) at UHF frequencies. The one way loss in dB is also calculated using the measured values of μ and ϵ .

There are many reports of measurements of the electric properties of soil in the literature, for example [1, 2, 3, 4, 5, 6]. However, most of these are primarily concerned with measuring only a real dielectric constant, ϵ' , and a conductivity, σ . Because some soils have ferromagnetic constituents, and because the soil near Albuquerque, NM appears to have some iron in it, it is desirable to measure both the electric and magnetic properties of the soil.

Electric Permittivity, Magnetic Permeability, and Conductivity

We will model the electric permittivity, ϵ , as a complex number, assuming the application of a time-harmonic field,

$$\epsilon = \epsilon_0 \epsilon_r = \epsilon_0 (\epsilon'_r - j\epsilon''_r), \quad (1)$$

where ϵ_0 is the electric permittivity of a vacuum. The relative electric permittivity, ϵ_r , arises from the interaction of the electromagnetic field with charge, located in the material in the form of electric dipoles and free monopoles. Generally, conductivity is used to describe the interaction with the free monopoles, and electric permittivity is used to describe the dipole interaction. For our purposes, we will combine both interactions into one interaction described by a complex relative electric permittivity, sometimes called the dielectric constant.

Balanis [7] provides a good discussion of the physics of dielectrics and shows that for time-harmonic fields (with time dependence $e^{j\omega t}$), a simple dielectric (one with only a single species of electric dipole) the relative dielectric constant can be modeled as

$$\epsilon_r = 1 + \frac{\frac{N_e Q^2}{\epsilon_0 m \omega_0^2}}{1 - \frac{\omega^2}{\omega_0^2} + j\omega \frac{d}{m \omega_0^2}}, \quad (1)$$

where N_e is the volume density of the dipole species, Q is the dipole charge, m is the mass associated with the moving charge, d is the damping coefficient due to friction, ϵ_0 is the permittivity of free space, ω is the applied-field radian frequency, and ω_0 is the resonant frequency of the dipole (which is generally much larger than microwave frequencies). A more thorough discussion of the subject is given by Von Hippel [8]. When the material is more complex and contains many dipole species and the static conductivity term is included, the relative dielectric constant would be

$$\epsilon_r = 1 + \sum_n \frac{\frac{N_{en} Q_n^2}{\epsilon_0 m_n \omega_{0n}^2}}{1 - \frac{\omega^2}{\omega_{0n}^2} + j\omega \frac{d_n}{m_n \omega_{0n}^2}} - j \frac{\sigma_s}{\omega \epsilon_0} = \epsilon'_r - j\epsilon''_r, \quad (2)$$

where σ_s is the static conductivity. Thus, we define the relative dielectric constant so that the imaginary part, ϵ'' , will also contain the conductivity term, $\frac{\sigma_s}{\omega \epsilon_0}$.

The magnetic permeability, μ , describes the interaction between the magnetic field and magnetic dipoles contained in a material. There are no magnetic monopoles, so a magnetic conductivity is not necessary. Similar to the electric permittivity, we will model the relative magnetic permeability, μ_r , as complex,

$$\mu = \mu_0 \mu_r = \mu_0 (\mu'_r - j\mu''_r), \quad (2)$$

where μ_0 is the magnetic permeability of a vacuum.

Measuring the Dielectric and Magnetic Properties of a Material

At UHF and microwave frequencies, three of the possible measurement techniques used to determine the electrical properties of a material include capacitance measurement, measurement of the resonant frequency and circuit Q of a resonant cavity, and measuring the characteristics of a transmission line. The capacitance measurement procedure requires placing a sample of known size between two plates to form a capacitor. However, this method is of limited use above one GHz, and it does not provide information about the magnetic permeability of the sample. The resonant-cavity method can be very accurate at a spot frequency but does not provide continuous frequency information. In its usual form, it does not provide information about the magnetic permeability, either. The transmission-line method, however, provides several attractive characteristics including ability to measure the continuous frequency response of the sample, measurement of nonsolid materials when the transmission line is appropriately designed, and the magnetic permeability can be measured.

The complex dielectric permittivity, ϵ , and the complex magnetic permeability, μ , determine the electrical properties of a material, including velocity of wave propagation, impedance properties, and loss. Since there are two parameters, two measurements are required to determine them. These two measurements consist of measuring the reflection of a wave at the boundary of a known material and the unknown material (impedance characteristic), and of measuring the transmission through a known length of the unknown material (velocity of wave propagation and loss properties). A convenient method of obtaining these measurements is to place the material inside a length of coaxial transmission line and then to measure the S -matrix of the transmission-line section with a network analyzer. In the following, the equations are derived to calculate the complex ϵ and μ from the set of measurements.

The Wave Matrix and the S-Matrix

Measuring an unknown test sample requires cascading sections of transmission line which have different characteristic impedances. Rather than the usual forward and reverse wave voltages, V_i^+ and V_i^- , normalized wave amplitudes will be used

$$u_i^\pm \equiv \frac{V_i^\pm}{\sqrt{Z_i}}, \quad (3)$$

where Z_i is the characteristic impedance of the i^{th} port. Using normalized wave amplitudes allows the cascading of networks represented by wave matrices. The + superscript represents a wave entering a port, while the - superscript represents a wave leaving the port. The two-port network, shown in Fig. 1, can be represented by a generalized scattering matrix

$$\begin{bmatrix} u_1^- \\ u_2^- \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} u_1^+ \\ u_2^+ \end{bmatrix}, \quad (4)$$

and a wave matrix [9]

$$\begin{bmatrix} u_1^+ \\ u_1^- \end{bmatrix} = \begin{bmatrix} W_{11} & W_{12} \\ W_{21} & W_{22} \end{bmatrix} \begin{bmatrix} u_2^- \\ u_2^+ \end{bmatrix}. \quad (5)$$

The S-matrix representation is the form commonly measured with modern network analyzers, while the wave matrix is useful for cascading multiple networks. In the cascade configuration, u_1^+ and u_2^- represent waves traveling in the same direction through the cascade of two-port networks. Thus, the output of one two-port becomes the input of the next, and the total wave matrix is the product of the individual wave matrices.

Fig. 1 Representation of two-port network, showing the port variables used to characterize its performance.

The two sets of parameters are related by

$$S_{11} = \frac{W_{21}}{W_{11}}, \quad (6)$$

$$S_{12} = \frac{W_{11}W_{22} - W_{12}W_{21}}{W_{11}}, \quad (7)$$

$$S_{21} = \frac{1}{W_{11}}, \quad (8)$$

$$S_{22} = -\frac{W_{12}}{W_{11}}, \quad (9)$$

and

$$W_{11} = \frac{1}{S_{21}}, \quad (10)$$

$$W_{12} = -\frac{S_{22}}{S_{21}}, \quad (11)$$

$$W_{21} = \frac{S_{11}}{S_{21}}, \quad (12)$$

$$W_{22} = -\frac{(S_{11}S_{22} - S_{12}S_{21})}{S_{21}}. \quad (13)$$

The network is reciprocal, and reciprocity requires

$$S_{12} = S_{21}, \quad (14)$$

and

$$W_{11}W_{22} - W_{12}W_{21} = 1. \quad (15)$$

Measuring the Sample

In order to contain the sample within the section of airline, the ends must be plugged with a solid material, such as Teflon. Thus, the measurement of the sample requires cascading three networks, as shown in Fig. 2.

Fig. 2 Diagram showing three sections of transmission line cascaded for measuring the test sample.

There will be a wave matrix associated with each of the four interfaces and a wave matrix associated with each of the three lengths of transmission line. The measurable wave amplitudes are related by

$$\begin{aligned}
\begin{bmatrix} u_1^{(end)+} \\ u_1^{(end)-} \end{bmatrix} &= \frac{1}{T_{01}} \begin{bmatrix} 1 & \Gamma_{01} \\ \Gamma_{01} & 1 \end{bmatrix} \begin{bmatrix} e^{j\theta_1} & 0 \\ 0 & e^{-j\theta_1} \end{bmatrix} \\
&\cdot \frac{1}{T_{12}} \begin{bmatrix} 1 & \Gamma_{12} \\ \Gamma_{12} & 1 \end{bmatrix} \begin{bmatrix} e^{j\theta_2} & 0 \\ 0 & e^{-j\theta_2} \end{bmatrix} \\
&\cdot \frac{1}{T_{21}} \begin{bmatrix} 1 & -\Gamma_{12} \\ -\Gamma_{12} & 1 \end{bmatrix} \begin{bmatrix} e^{j\theta_3} & 0 \\ 0 & e^{-j\theta_3} \end{bmatrix} \\
&\cdot \frac{1}{T_{10}} \begin{bmatrix} 1 & -\Gamma_{01} \\ -\Gamma_{01} & 1 \end{bmatrix} \begin{bmatrix} u_2'^{(end)+} \\ u_2'^{(end)-} \end{bmatrix}
\end{aligned} \tag{16}$$

which can be written in terms of the measurable S -matrix elements

$$\begin{bmatrix} u_1^{(end)+} \\ u_1^{(end)-} \end{bmatrix} = \begin{bmatrix} \frac{1}{S_{21}} & \frac{-S_{22}}{S_{21}} \\ \frac{S_{11}}{S_{21}} & \frac{-(S_{11}S_{22} - S_{12}S_{21})}{S_{21}} \end{bmatrix} \begin{bmatrix} u_2'^{(end)+} \\ u_2'^{(end)-} \end{bmatrix}, \tag{17}$$

In the above

$$\Gamma_{01} = \frac{Z_{01} - Z_0}{Z_{01} + Z_0}, \tag{18}$$

$$\Gamma_{12} = \frac{Z_{02} - Z_{01}}{Z_{02} + Z_{01}}, \tag{19}$$

$$T_{10} = T_{01} = \frac{2\sqrt{Z_0 Z_{01}}}{Z_0 + Z_{01}}, \tag{20}$$

$$T_{12} = T_{21} = \frac{2\sqrt{Z_{01} Z_{02}}}{Z_{01} + Z_{02}}, \tag{21}$$

$$\theta_1 = \frac{2\pi f}{c} L_1 \sqrt{\epsilon_{r,end} \mu_{r,end}}, \tag{22}$$

$$\theta_2 = \frac{2\pi f}{c} L_2 \sqrt{\epsilon_{r,sample} \mu_{r,sample}}, \tag{23}$$

and

$$\theta_3 = \frac{2\pi f}{c} L_3 \sqrt{\epsilon_{r,end} \mu_{r,end}}, \tag{24}$$

where c is the speed of light in the vacuum. Note that the transmission coefficients (20-21) are not given by the usual voltage transmission coefficient, $T = 1 + \Gamma$, because the wave amplitudes

are normalized voltages (3), rather than voltages. However, this normalization does not affect the reflection coefficients.

The characteristic impedances are

$$Z_0 = 50 \Omega \quad (25)$$

for the usual S-parameter measurement test set, and

$$Z_{01} = \sqrt{\frac{\mu_0}{\epsilon_0}} \frac{\ln\left(\frac{b}{a}\right)}{2\pi} \sqrt{\frac{\mu_{r,end}}{\epsilon_{r,end}}}, \quad (26)$$

$$Z_{02} = \sqrt{\frac{\mu_0}{\epsilon_0}} \frac{\ln\left(\frac{b}{a}\right)}{2\pi} \sqrt{\frac{\mu_{r,sample}}{\epsilon_{r,sample}}}, \quad (27)$$

where a is the radius of the inner conductor and b is the radius of the outer conductor of the coaxial line. Note that for air-filled coaxial line with 50Ω impedance, such as GR-900 line,

$$\sqrt{\frac{\mu_0}{\epsilon_0}} \frac{\ln\left(\frac{b}{a}\right)}{2\pi} = 50 \Omega. \quad (28)$$

For convenience, normalized bulk impedances are defined as

$$z_{r,end} = \sqrt{\frac{\mu_{r,end}}{\epsilon_{r,end}}}, \quad (29)$$

and

$$z_{r,sample} = \sqrt{\frac{\mu_{r,sample}}{\epsilon_{r,sample}}}. \quad (30)$$

Then, for this case (GR-900)

$$\Gamma_{01} = \frac{z_{r,end} - 1}{z_{r,end} + 1}, \quad (31)$$

$$\Gamma_{12} = \frac{z_{r,sample} - z_{r,end}}{z_{r,sample} + z_{r,end}}, \quad (32)$$

$$T_{01} = (1 + \Gamma_{01}) \frac{1}{\sqrt{z_{r,end}}} = \frac{2\sqrt{z_{r,end}}}{z_{r,end} + 1} = \sqrt{(1 - \Gamma_{01}^2)}. \quad (33)$$

and

$$T_{12} = (1 + \Gamma_{12}) \frac{\sqrt{z_{r,end}}}{\sqrt{z_{r,sample}}} = \frac{2\sqrt{z_{r,sample} z_{r,end}}}{z_{r,sample} + z_{r,end}} = \sqrt{1 - \Gamma_{12}^2}. \quad (34)$$

Phase factors are defined

$$p_1 \equiv e^{j\theta_1}, \quad (35)$$

$$p_2 \equiv e^{j\theta_2}, \quad (36)$$

and

$$p_3 \equiv e^{j\theta_3}. \quad (37)$$

Thus, the following four (dependent) equations must be solved for Γ_{12} , and p_2 , which are then solved for $\epsilon_{r,sample}$ and $\mu_{r,sample}$

$$\frac{1}{(1 - \Gamma_{12}^2)} \begin{bmatrix} p_2 - \frac{\Gamma_{12}^2}{p_2} & -\Gamma_{12} \left(p_2 - \frac{1}{p_2} \right) \\ \Gamma_{12} \left(p_2 - \frac{1}{p_2} \right) & \frac{1}{p_2} - \Gamma_{12}^2 p_2 \end{bmatrix} = \begin{bmatrix} W_{11}^{sample} & W_{12}^{sample} \\ W_{21}^{sample} & W_{22}^{sample} \end{bmatrix}, \quad (38)$$

where

$$\begin{aligned} \begin{bmatrix} W_{11}^{sample} & W_{12}^{sample} \\ W_{21}^{sample} & W_{22}^{sample} \end{bmatrix} &= \frac{1}{S'_{21}} \begin{bmatrix} 1 & S'_{22} \\ S'_{11} & -(S'_{11}S'_{22} - S'_{12}S'_{21}) \end{bmatrix} \\ &= \frac{1}{(1 - \Gamma_{01}^2)} \begin{bmatrix} \frac{1}{p_1} & -\frac{\Gamma_{01}}{p_1} \\ -\Gamma_{01}p_1 & p_1 \end{bmatrix} \frac{1}{S_{21}} \begin{bmatrix} 1 & -S_{22} \\ S_{11} & -(S_{11}S_{22} - S_{12}S_{21}) \end{bmatrix} \begin{bmatrix} \frac{1}{p_3} & \Gamma_{01}p_3 \\ \frac{\Gamma_{01}}{p_3} & p_3 \end{bmatrix}. \end{aligned} \quad (39)$$

In (39), the measured S -matrix is

$$\mathbf{S} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix}, \quad (40)$$

and

$$\mathbf{S}' = \begin{bmatrix} S'_{11} & S'_{12} \\ S'_{21} & S'_{22} \end{bmatrix} \quad (41)$$

represents the measurement after correcting for the presence of the end sections, as shown. Because of the symmetry (it is assumed that the end sections are made of the same material),

$$S'_{11} = S'_{22},$$

while reciprocity imposes

$$S'_{12} = S'_{21}.$$

Two non-linear equations must be solved to obtain Γ_{12} and p_2 :

$$S'_{11} = \frac{\Gamma_{12} \left(p_2 - \frac{1}{p_2} \right)}{p_2 - \frac{\Gamma_{12}^2}{p_2}}, \quad (42)$$

and

$$S'_{21} = \frac{(1 - \Gamma_{12}^2)}{p_2 - \frac{\Gamma_{12}^2}{p_2}}. \quad (43)$$

When $|S'_{11}|$ is not too near zero, the solution for Γ_{12} and p_2 obtained as follows

$$\Gamma_{12} = \frac{S'^2_{11} - S'^2_{21} + 1}{2S'_{11}} \pm \sqrt{\left(\frac{S'^2_{11} - S'^2_{21} + 1}{2S'_{11}} \right)^2 - 1}, \quad (44)$$

$$p_2 = \frac{S'_{21} \Gamma_{12}}{\Gamma_{12} - S'_{11}}. \quad (45)$$

However, if $|S'_{11}| \sim 0$, an iterative solution should be used. We can obtain the m^{th} iteration for Γ_{12} from (43)

$${}_m\Gamma_{12} = \frac{S'_{11} ({}_m p_2^2 - {}_{m-1}\Gamma_{12}^2)}{{}_m p_2^2 - 1}. \quad (46)$$

and for p_2 from (42)

$${}_m p_2 = \frac{1}{S'_{21}} \frac{(1 - {}_{m-1}\Gamma_{12}^2)}{\left(1 - \frac{{}_{m-1}\Gamma_{12}^2}{{}_{m-1}p_2^2} \right)}, \quad (47)$$

Computing the value of Γ_{12} first in each iteration is slightly more efficient than computing p_2 first. The iteration starts with

$${}_{m=0}\Gamma_{12} = S'_{11},$$

and

$${}_{m=0}p_2 = \frac{1}{S'_{21}}.$$

The electrical length of the sample must not be near a multiple of a wavelength for (46) to be valid.

The values of Γ_{12} and p_2 are obtained either from the exact solution of (44) and (45), or from the iterative solution. Once these values have been obtained, the solution for the constitutive parameters of the unknown sample is completed as follows:

$$z_{r, sample} = z_{r, end} \frac{1 + \Gamma_{12}}{1 - \Gamma_{12}}, \quad (48)$$

$$c_{r, sample} \equiv \frac{1}{\sqrt{\mu_{r, sample} \epsilon_{r, sample}}} = j \frac{2\pi f L_2}{c \ln(p_2)}, \quad (49)$$

$$\epsilon_{r, sample} = \frac{1}{z_{r, sample} c_{r, sample}}, \quad (50)$$

and

$$\mu_{r, sample} = \frac{z_{r, sample}}{c_{r, sample}}. \quad (51)$$

The iterative solution has no ambiguity, but the exact solution, solving a quadratic equation in Γ_{12} , does have a sign ambiguity in (44). This ambiguity is resolved by requiring $\text{Re}(z_{r, sample}) > 0$. The choice of when to use the iterative or exact solution is somewhat arbitrary, but a useful choice is use the iterative solution when $|S'_{11}| < 0.5$.

The method described here is similar to that described in [10], but this method includes the effects of the end sections. When measuring dielectrics which are not solid, the end sections are necessary in order to contain the material. Note that Γ and T in [10] are

$$\Gamma = \Gamma_{12} \Big|_{\text{material 1 is air}},$$

and

$$T = \frac{1}{p_2}$$

in the method described here. Alternatively, the Nicolson-Ross algorithm described in [10] can be applied directly once S' , the S-matrix for the center section, has been obtained from the measured S-matrix.

The Measurement Configuration

The measurement system can be divided into a hardware section and a software section. The hardware consists of an HP 8510C network analyzer, a calibration kit, and sample holders. In order to allow soil samples with coarse grain structure, and to simplify the loading of the sample holder, a large air-filled coaxial transmission line, type GR900, is used. The Nicolson-Ross algorithm requires the sample length to be less than $1/2 \lambda$ to avoid drop-outs. The broad band width and wide range of dielectric characteristics required air lines of three different lengths to perform the measurements. The air line for the longest wave lengths, had a center conductor length of 19.986 cm. The air line used in the center of the band and at the lower end of the band for lossy materials

had a center conductor length of 7.495 cm. The shortest sample holder had a center conductor length of 2.998 cm. One of the soil samples was so lossy, Fig. 27, that only the short sample holder was used during its characterization. End caps fabricated from Teflon held the soil in the GR900 sample holders during the measurements.

The calibration procedure is key to obtaining valid absolute results from an automatic network analyzer such as the HP 8510C. Unfortunately, a commercial calibration kit for GR900 connectors is not currently available. However, the Sandia Primary Standards lab was able to generate correction coefficients which were loaded into the HP 8510C, which allowed a GR900 calibration kit to be assembled, permitting the network analyzer to be calibrated for GR900 measurements. The calibration software supplied by Primary Standards is capable of doing both a classic, (match, short, open) calibration and the newer TRL style of calibration. The software for converting the measured scattering parameters of the soil into ϵ and μ was written in Matlab.

Electric Permittivity and Magnetic Permeability of Sample Soils

Soil samples were collected from a number of sights, mostly in the local area. Samples were collected from the SAR test pit in Area 3 several times, to determine how rain fall would affect the soils dielectric properties. Table 1 gives the soil's name and where the results are plotted.

Table 1 SOIL SAMPLES

Soil Type	Electric permittivity ϵ' / ϵ''	Magnetic Permeability μ' / μ''	Loss dB / m (one way)
Dry Native Soil	Fig. 3, pp. 15 / Fig. 4, pp. 16	Fig. 5, pp. 16 / Fig. 6, pp. 17	Fig. 7, pp. 17
Dry Sand	Fig. 8, pp. 18 / Fig. 9, pp. 18	Fig. 10, pp. 19 / Fig. 11, pp. 19	Fig. 12, pp. 20
Damp Native Soil	Fig. 13, pp. 20 / Fig. 14, pp. 21	Fig. 15, pp. 21 / Fig. 16, pp. 22	Fig. 17, pp. 22
Damp Sand	Fig. 18, pp. 23 / Fig. 19, pp. 23	Fig. 20, pp. 24 / Fig. 21, pp. 24	Fig. 22, pp. 25
Louisiana Soil	Fig. 23, pp. 25 / Fig. 24, pp. 26	Fig. 25, pp. 26 / Fig. 26, pp. 27	Fig. 27, pp. 27
Belen Sample #1	Fig. 28, pp. 28 / Fig. 29, pp. 28	Fig. 30, pp. 29 / Fig. 31, pp. 29	Fig. 32, pp. 30
Belen Sample #2	Fig. 33, pp. 30 / Fig. 34, pp. 31	Fig. 35, pp. 31 / Fig. 36, pp. 32	Fig. 37, pp. 32
Belen Sample #3	Fig. 38, pp. 33 / Fig. 39, pp. 33	Fig. 40, pp. 34 / Fig. 41, pp. 34	Fig. 42, pp. 35

The measurements for each soil type often contain data from more than one length of sample holder. Results vary slightly between sample holders due to variations in packing density and moisture content. The shortest sample holder was the quickest to load (less drying time) and the easiest to compress (higher packing densities). Increased moisture content and greater packing density produce greater loss in the samples measured with the short sample holder. The long sample holder was invaluable for measuring low loss soils at frequencies below 100 MHz, where the short sample holder was plagued by errors due to its short phase length and capacitance from air gaps. These errors are most easily observed in plots of μ_r'' such as Fig. 6 where the two shorter sample holders provide incorrect results at the lower end of the band. It is also important to remember that soil is not necessarily a homogeneous media. It may contain many different minerals and organic materials in a single sample. While each of the samples may have come from the same shovel full of soil, they are by no means identical.

In addition to electrical measurements, three of the damp soil samples were weighed before and after the moisture had been removed, and the percentage of water by weight was calculated. Three measurements were made on each soil sample, and the results were averaged. The sand sample, Fig. 18 through Fig. 22, contained 2.65% moisture while the native soil, Fig. 13 through Fig. 17, contained 8.2% moisture. A sample of soil from Toledo Bend in Louisiana, Fig. 23 through Fig. 27, was found to contain 25.6% moisture by weight. The damp sand and damp native soil came from the same location, the SAR test pit, after exposure to the same weather, but they had very different moisture contents. The difference in moisture content is explained by the difference in specific surface area between the soil types. Specific surface area is a metric that provides insight into the ability of a soil to bind water. Clay particles are often platy and thus contribute even more surface area per volume than their small size alone would indicate. The specific surface area of sand is often less than $1 \text{ m}^2/\text{gm}$ while clay may be as high as several hundred m^2/gm . The clay is thus able to "hold" more moisture. The ramifications of the disparity in moisture content are readily observable in the dielectric measurements.

Fig. 3 Native soil from Kirtland Air Force Base, SAR test pit, ϵ_r' (Bottom mid)

Fig. 4 Native soil from Kirtland Air Force Base, SAR test pit, ϵ_r'' Loss (Bottom mid)

Fig. 5 Native soil from Kirtland Air Force Base, SAR test pit, μ_r' (Bottom mid)

Fig. 6 Native soil from Kirtland Air Force Base, SAR test pit, μ_r'' Loss (Bottom mid)

Fig. 7 Native soil from Kirtland Air Force Base, SAR test pit, Loss dB/m (Bottom mid)

Fig. 8 Sand from Kirtland Air Force Base, SAR test pit, ϵ_r

Fig. 9 Sand from Kirtland Air Force Base, SAR test pit, ϵ_r'' Loss

Fig. 10 Sand from Kirtland Air Force Base, SAR test pit, μ_r

Fig. 11 Sand from Kirtland Air Force Base, SAR test pit, μ_r'' Loss

Fig. 12 Sand from Kirtland Air Force Base, SAR test pit, Loss dB/m

Fig. 13 Damp native soil from Kirtland Air Force Base, SAR test pit, ϵ_r

Fig. 14 Damp native soil from Kirtland Air Force Base, SAR test pit, ϵ_r'' Loss

Fig. 15 Damp native soil from Kirtland Air Force Base, SAR test pit, μ_r'

Fig. 16 Damp native soil from Kirtland Air Force Base, SAR test pit, μ_r'' Loss

Fig. 17 Damp native soil from Kirtland Air Force Base, SAR test pit, Loss dB/m

Fig. 18 Damp sand from Kirtland Air Force Base, SAR test pit, ϵ_r'

Fig. 19 Damp sand from Kirtland Air Force Base, SAR test pit, ϵ_r'' Loss

Fig. 20 Damp sand from Kirtland Air Force Base, SAR test pit, μ_r'

Fig. 21 Damp sand from Kirtland Air Force Base, SAR test pit, μ_r'' Loss

Fig. 22 Damp sand from Kirtland Air Force Base, SAR test pit, Loss dB/m

Fig. 23 Damp lossy soil, Toledo Bend, Louisiana, ϵ_r

Fig. 24 Damp lossy soil, Toledo Bend, Louisiana, ϵ_r'' Loss

Fig. 25 Damp lossy soil, Toledo Bend, Louisiana, μ_r'

Fig. 26 Damp lossy soil, Toledo Bend, Louisiana, μ_r'' Loss

Fig. 27 Damp lossy soil, Toledo Bend, Louisiana, Loss dB/m

Fig. 28 Belen soil location #1, as collected and after drying ϵ_r'

Fig. 29 Belen soil location #1, as collected and after drying ϵ_r'' Loss

Fig. 30 Belen soil location #1, as collected and after drying μ_r'

Fig. 31 Belen soil location #1, as collected and after drying μ_r'' Loss

Fig. 32 Belen soil location #1, as collected and after drying, Loss dB/m

Fig. 33 Belen soil location #2, as collected and after drying ϵ_r'

Fig. 34 Belen soil location #2, as collected and after drying ϵ_r'' Loss

Fig. 35 Belen soil location #2, as collected and after drying μ_r'

Fig. 36 Belen soil location #2, as collected and after drying μ_r'' Loss

Fig. 37 Belen soil location #2, as collected and after drying, Loss dB/m

Fig. 38 Belen soil location #3, as collected and after drying ϵ_r'

Fig. 39 Belen soil location #3, as collected and after drying ϵ_r'' Loss

Fig. 40 Belen soil location #3, as collected and after drying μ_r'

Fig. 41 Belen soil location #3, as collected and after drying μ_r'' Loss

Fig. 42 Belen soil location #3, as collected and after drying, Loss dB/m

Summary and Conclusion

The transmission-line measurement system described here allows timely determination of the complex electric permittivity and magnetic permeability of soil with reasonable accuracy. The accuracy has been verified through comparison of measurements obtained with the GR900 transmission-line system and measurements performed by the Primary Standards lab. This comparison is documented in Appendix I. A number of soil samples have been measured on an opportunistic basis. The same soil can produce significantly different loss characteristics as seen by a comparison of Fig. 7 and Fig. 17 or Fig. 12 and Fig. 22, where the addition of 8% moisture doubled the transmission loss in the soil. Some soils, such as the sample from Toledo Bend, Fig. 27, are clearly not conducive to UHF propagation. The measurement system we have in place can provide cost-effective characterization of a soil before an expensive SAR flight. The *a priori* knowledge of a soil's transmission characteristics can help determine radar and flight parameters for a successful SAR data acquisition.

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Appendix I. Validation of Dielectric Measurement Process

This appendix documents the results of the verification process for the dielectric measurement system. The first phase of the validation process required characterization of material samples by the microwave section of the Sandia Primary Electrical Standards Department (1742). The complex permittivity and permeability measured by Primary Standards was then compared with our measurements. The second phase of the validation procedure involved determination of how well the end cap correction algorithm was working.

Primary Standards Phase

Primary Standards used two methods to determine the electrical properties of the materials in question. The first method involved the use of three reentrant cavities, one for each of the spot frequencies. A network analyzer (HP 8753) is used to measure the resonant frequency of the loaded Karpova cavity and its Q . The complex permittivity ($\epsilon = \epsilon' + j\epsilon''$) is then calculated using a program developed by primary standards. The cavity method requires solid samples of specific dimensions. The second method employed by primary standards was the transmission line method using GR900 coaxial sample holders. The samples were measured using an HP8510C network analyzer and the material properties were obtained using HP software (*HP 85071B Materials Measurement Software, Rev. 1.01*). This software package allows the use of several algorithms to determine the sample's properties. The Nicolson-Ross algorithm was selected as it is the only one which provides information on the magnetic properties (μ) of the sample.

Four different materials were measured by the Primary Standards Lab. Teflon and Nylatron were measured using the cavity method and complex permittivity (ϵ) was calculated. MF-112 and MF-116, ferrite materials manufactured by Emerson & Cuming Inc., were measured using both methods. A subset of the results for each of the materials is given on the following pages.

The NO GAP designation after the line length of the 2343 measurements indicates that the air gap between the sample and the inner and outer conductor of the coax was eliminated through the use of a dielectric grease. The first plot in each set contains the real portion of the measured data. The loss or imaginary portion of the material's permittivity (ϵ'') or permeability (μ'') is given in the next graph.

Teflon is the first material in the comparison. It was selected as a good material to test the lower bounds of the measurement system since its dielectric constant and loss tangent should be lower than almost any soil. Teflon also should provide a fairly constant response across the frequencies in question.

Fig. 1 Teflon in three sample holders of different lengths and a cavity (real portion)

Fig. 2 Teflon in three sample holders of different lengths and a cavity (imaginary portion)

Obviously the two measurement techniques agree very well for the complex permittivity of Teflon. A major weakness of the coax method is the large uncertainty associated with the loss portion (ϵ'') of low loss materials. A second weakness of the coax method when used with the Nicolson-Ross algorithm is that the results are not valid if the line length is $n\lambda_{\text{coax}}/2$. This is shown by the drop in the data for the medium length line at frequencies above 1,150 MHz. In general, the data after $\lambda_{\text{coax}}/2$ has not been plotted to limit confusion.

The agreement of the two methods is not as good for the real portion of the permittivity of the Nylatron samples. A possible reason for the discrepancy is that the samples 2343 measured are not from the same block of material as the samples Primary Standards measured.

Fig. 3 Nylatron in three sample holders of different lengths and a cavity (real portion)

The loss portion of the Nylatron measurement is in good agreement as seen in Fig. 4 on the following page.

Fig. 4 Nylatron in three sample holders of different lengths and a cavity (imaginary portion)

Fig. 5 Ferrite MF-112 permittivity (real portion)

The ferrite samples MF-112 and MF-116 were measured by Primary Standards using both the coax and resonant cavity method. To conserve space, only the data for MF-112 is presented in this Appendix.

The C/A designation after some of the Primary Standards Lab measurements indicates that HP's correction for air gaps has been applied. The spot frequency data does not take into account the permeability of the material and thus assumes all deviations from the empty cavity resonant frequency and Q are due to permittivity. The dielectric loss associated with this material is high, but not as high as some of the damp soil samples.

Fig. 6 Ferrite MF-112 permittivity (imaginary portion)

The cavity method did not provide any information on permeability and thus, is absent from Figs. 7 and 8. The permeability measurements compare very well as seen in Fig. 7 and Fig. 8.

Fig. 7 Ferrite MF-112 permeability (real portion)

Fig. 8 Ferrite MF-112 permittivity (imaginary portion)

End Cap Algorithm Phase

The soil samples are held in the air line by Teflon end caps placed at each end of the line. Measurement accuracy is enhanced by removing the effect of the end caps through a correction algorithm. The ability of the algorithm to mathematically remove the end caps was tested by measuring an empty airline and then inserting the washers (end caps) and remeasuring. The empty air line measurement is compared to the corrected airline and washer measurement in the following two plots. Two different widths of Teflon washers were used for the measurement. In all of the recent soil measurements the thin Teflon washers have been used to limit the effect of the washers. The data for the short air line is the worst case since the end caps take up a larger percentage of the total sample holder length for that case. Fig. 9 and Fig. 10 show the correction for the air line corrects to the same level irrespective of the type of end cap used.

Fig. 9 Comparison of short air line with and without Teflon washers (real portion)

Short Air Line (With & Without Teflon Washers)
Relative Dielectric Constant (ϵ_r') Loss

Fig. 10 Comparison of short air line with and without Teflon washers (imaginary portion)

The data for the loss portion of the previous measurement highlights the problem with the large uncertainty associated with low loss samples. The measurement clearly shows that using the short air line to measure ϵ'' values of less than -0.02 at frequencies below 150 MHz is not accurate.

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