A Feasibility Study to Monitor Soil Moisture Content Using Microwave Signals

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***Abstract*—A buried leaky coaxial cable is used as a sensor to continuously monitor the average moisture content of irrigated soil at a desired depth. The two-phase feasibility study consists of an analytical and an experimental investigation. The modal equation in the cable is expressed in terms of the surface impedance at the inner conductor boundary and an effective surface impedance at the outer leaky conductor boundary. The effective surface impedance is related to the soil moisture content through its measured attenuation. The results of the experimental work conducted in the field are in good agreement with the analytical results. It is shown that the recorded phase difference between the test signal from the buried cable and the reference signal can be used to monitor the soil moisture content.**

I. Introduction

The measurement of soil moisture content is of considerable interest to the agricultural community. Unfortunately, soil moisture content is not an easy variable to measure in field conditions. The commonly used methods have significant shortcomings. For example, gravimetric methods are slow, tedious, and necessitate the disturbance of the soil [1]. Soil conduction methods are inherently inaccurate and unreliable [1]. Radiation methods such as neutron thermalization and gamma-ray attenuation are not simple to operate and are also limited by safety considerations [1]. Moreover, these measurements have a common disadvantage in that they are all based upon measurements at one specific location.

Improved soil moisture measurement techniques which are suited to field conditions are required. The radio-wave method described in this paper has several important advantages. The soil moisture sensor covers a large portion of an irrigated field; thus an average measurement is determined. Once the sensing system is installed it is not necessary to disturb the soil in the measurement process. The system can be operated on a continuous basis; thus, an automated irrigation system could control the distribution of water to the field as prescribed by the user.

Before introducing the radio-wave method, it is useful to briefly review some basic concepts in soil engineering [2]–[5]. The soil is regarded as a matrix which supports the plants and provides water and minerals. It consists of water, mineral particles, organic matter, and living organisms. A given volume of soil V consists of a volume of water V_w, a volume of solids V_s, and a volume of air V_a, of mass M_w, M_s, and M_a, respectively. Soil moisture content θ_m defined on the basis of weight is given by

$$\theta_m = M_w / M_s. \tag{1}$$

The soil moisture content is not a very reliable measure of the availability of the water to the plants. A more useful concept has been defined and is referred to as the soil water potential ψ. The soil water potential is the amount of work required to move a given amount of water from one point to another, compared to the amount of work required to move the same amount of pure, unbound water the same distance. The soil water potential ψ is defined as a negative quantity, and is usually expressed in units of energy per unit mass. The soil water potential ψ is composed of three components: the matric potential ψ_m, the osmotic potential ψ_0, and the pressure potential ψ_p. The three components of soil water potential are additive, thus

$$\psi = \psi_0 + \psi_m + \psi_p. \tag{2}$$

Normally, the water characteristics of a given soil are defined by the curve relating the matric potential ψ_m to the soil moisture content θ_m. A typical relationship is illustrated in Fig. 1. There is a small amount of hysterisis between the wetting and drying curves. In Fig. 2, ψ_m is plotted as a function of θ_m for different soil textures [5].

The wilting point is defined by the soil moisture content at which the plants experience permanent wilting. This usually corresponds to a matric potential of -1500 J/Kg. Field capacity is defined by the moisture content of soil which is thoroughly saturated with water, allowing for all the water that can drain off by gravitational force to do so. This corresponds to a matric potential of -10 J/Kg. Available water is the water contained by the soil between the wilting point and field capacity. The available water provides an estimate of the moisture that is normally available to the plants.

The measurement of soil moisture using radio waves is based on the strong dependence of the complex permittivity of soil $\epsilon_1 = \epsilon_1' - j\epsilon_1''$ on the moisture content. A small portion of the power transmitted through a leaky coaxial cable radiates into the soil surrounding it. For a monochromatic signal (assuming an $\exp(j\omega t)$ time dependence), the complex propagation coefficient $\gamma = \alpha + j\beta$ of the azimuthally symmetric mode is influenced by the environment surrounding the leaky cable. Thus, γ is influenced by changes

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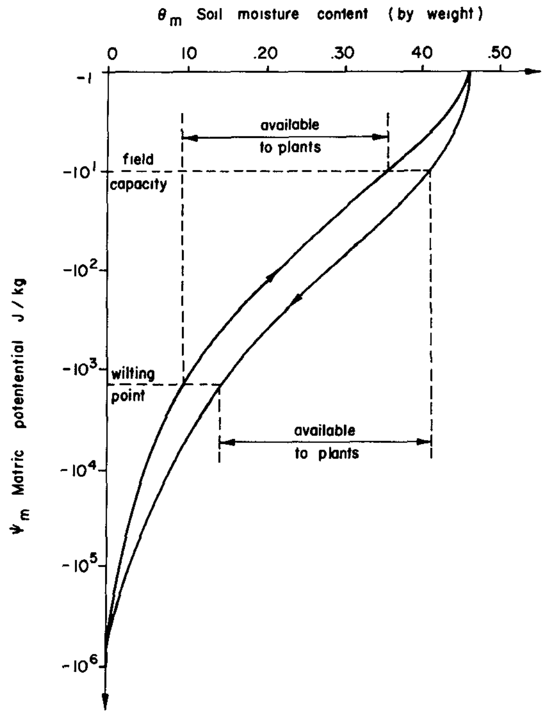

Fig. 1. Typical relationship between soil moisture content and matric potential (from Milthorpe [3]).

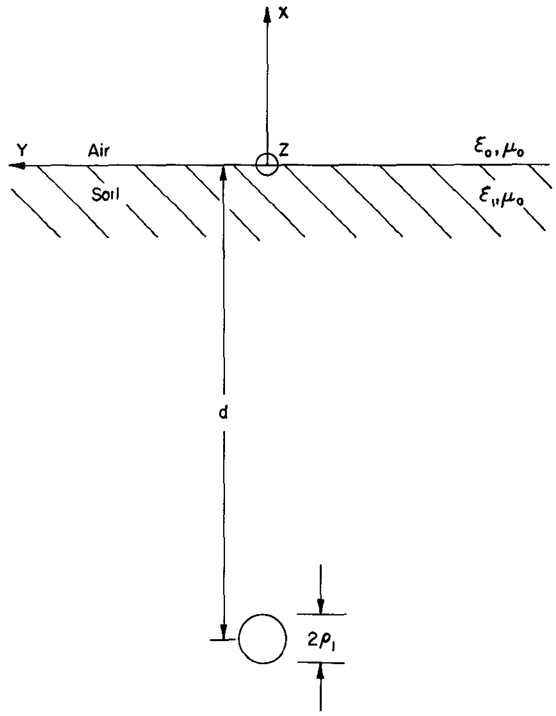

Fig. 3. Leaky coaxial cable buried at depth d below the air–earth interface.

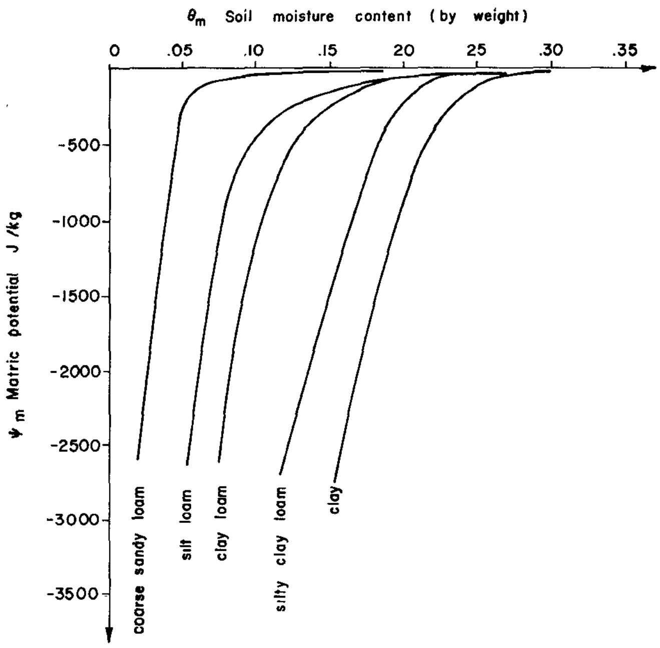

Fig. 2. Soil matric potential as a function of moisture content and texture (from Newton [5]).

in the soil permittivity, which varies with its moisture content. Changes in γ can be measured by observing changes in the transmission coefficient S_{21} for the buried cable, or changes in its input reflection coefficient Γ_{in}.

II. Analysis

A. Geometry of the Problem

Consider an infinitely long coaxial cable buried at a depth d below the surface of the earth and parallel to the z-axis, as illustrated in Fig. 3. The permittivity above the air–earth interface (x–0 plane) is ϵ_0 (free-space permittivity), and below the interface it is $\epsilon_1 = \epsilon_1' - j\epsilon_1''$. The magnetic permeability above and below the interface is μ_0 (free-space permeability) [6]. The skin-depth (reciprocal of attenuation coefficient) in the earth is given by [7]

$$\delta_1 = \left[\frac{2}{\omega\mu_0\sigma_1}\right]^{1/2} \tag{3}$$

in which

$$\sigma_1 = \omega\epsilon_1'' \tag{4}$$

is the conductivity of the earth.

Propagation along a buried coaxial cable at very low (VLF) and extremely low (ELF) radio-wave frequencies is of considerable practical interest [8]. At these wavelengths, the guiding structures are usually within one skin-depth of the earth surface. Consequently, the air–earth interface affects the value of the propagation coefficient γ for the azimuthally symmetric mode in the coaxial cable [9].

Buried waveguide structures operating at higher radio-wave frequencies (around 100 MHz) have also been used in groundwave radar detection systems. In many of these systems the waveguide structure is a leaky coaxial cable, either with a braided outer conductor or an outer conductor with periodic slots. These structures are also located within a couple of skin-depths of the earth's surface to provide broad area coverage and to reduce the loss into the surrounding soil [10].

For the application considered in this work, the leaky coaxial cable is buried at a depth of 0.5 m below the earth surface and is excited by a 0.9-GHz signal. Thus, the cable is located many skin-depths δ_1 (between 2 and 12, depending on θ_m) below the earth surface. Since $d \gg \delta_1$, the air–earth interface may be ignored and the coaxial cable is

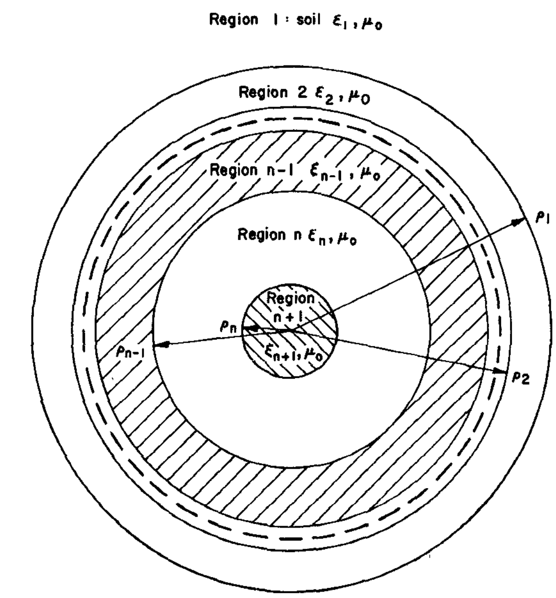

Fig. 4. Concentric geometry of the leaky cable.

considered to be centered along the z-axis (see Fig. 4) in an infinite medium (ϵ_1, μ_0). The structure has, in general, $n+1$ concentric regions $[(\epsilon_i, \mu_0), i=1,2,\cdots,n+1]$ which are separated by n interfaces $\rho=\rho_i, i=1,2,\cdots,n$. In this work, $n=5$, $i=6$ (the inner conductor), $i=5$ (dielectric layer), $i=4$ (the slotted outer conductor), $i=3$ (flooding compound), $i=2$ (protective outer jacket), and $i=1$ (the earth). The inner conductor of the cable, region $n+1$, is assumed to be a good conductor ($\sigma \gg \omega\epsilon$). Thus, it is represented by a surface impedance $Z_n = E_z/H_\phi$ at $\rho=\rho_n$. For good conductors, the surface impedance is given by [7]

$$Z_m = (1+j)\left[\frac{\omega\mu_0}{2\sigma_m}\right]^{1/2} \tag{5}$$

where σ_m is the conductivity of the metal (in this work $\sigma_{n+1}=\sigma_m=5.8\times10^7\ (\Omega-m)^{-1m}$ is the conductivity of copper). The surface impedance at an interface $\rho=\rho_i$, $i\leqslant n-1$ is defined by the ratio $-E_z/H_\phi$ evaluated at $\rho=\rho_i$. The surface impedance $Z_i (i\leqslant n-1)$ is defined such that $\mathrm{Re}(Z_i)=R_i \geqslant 0$ in a passive medium.

It is also assumed here that for the azimuthally symmetric modes, the slotted or braided conductor, region $n-1$, may be represented by a homogeneous medium with an effective complex permittivity ϵ_{n-1}. This assumption is reasonable provided that the dimensions of the slots and the distance separating them is very small compared to the wavelength in the medium. In Section II-E, an "effective" permittivity is calculated for the region $\rho\geqslant\rho_{n-1}$ from the measured cable loss.

B. Field Solutions

All field quantities are assumed to vary as $\exp(j\omega t-\gamma z)$, where $\gamma=\alpha+j\beta$ is the complex propagation coefficient. Assuming azimuthal symmetry ($\partial/\partial\phi=0$), we seek the solution for the $TM_{0,m}$ modes. Thus

$$H_z = H_\rho = E_\phi = 0. \tag{6}$$

The transverse components H_ϕ and E_ρ can be expressed in terms of E_z[11]. Thus

$$H_\phi = -j\frac{\omega\epsilon}{h^2}\frac{\partial E_z}{\partial \rho}$$

$$E_\rho = \frac{-\gamma}{h^2}\frac{\partial E_z}{\partial\rho} = \frac{\gamma}{j\omega\epsilon}H_\phi. \tag{7}$$

The wavenumber k_i for the ith region is given by

$$k_i = \omega(\mu_0\epsilon_i)^{1/2}, \qquad \mathrm{Im}(k_i)\leqslant 0 \tag{8}$$

and the characteristic values h_i are given by

$$h_i = (\gamma^2+k_i^2)^{1/2}, \qquad \mathrm{Im}(h_i)\leqslant 0. \tag{9}$$

Thus, the scalar wave equation for E_z in the ith region

$$\nabla^2 E_z + h_i^2 E_z = 0 \tag{10}$$

reduces to

$$\frac{1}{\rho}\frac{\partial}{\partial\rho}\left[\rho\frac{\partial E_z}{\partial\rho}\right] + h_i^2 E_z = 0. \tag{11}$$

The solution to (11) can be expressed in terms of the Bessel and Neumann functions. Thus, in the ith region

$$E_z = A_i h_i^2[J_0(\rho h_i) + B_i N_0(\rho h_i)]$$

and

$$H_\phi = jA_i\omega\epsilon_i h_i[J_1(\rho h_i)+B_iN_1(\rho h_i)] \tag{12}$$

in which A_i and B_i are arbitrary constants.

C. The Modal Equation

Since E_z and H_ϕ are continuous at each interface, the surface impedance is also continuous. The surface impedances Z_i^+ and Z_i^- are

$$Z_i^+ = \frac{-E_z(\rho_i^+)}{H_\phi(\rho_i^+)} = \frac{-h_i[J_0(\rho_ih_i)+B_iN_0(\rho_ih_i)]}{j\omega\epsilon_i[J_1(\rho_ih_i)+B_iN_1(\rho_ih_i)]} \tag{13}$$

and

$$Z_i^- = \frac{-E_z(\rho_i^-)}{H_\phi(\rho_i^-)} = \frac{-h_{i+1}[J_0(\rho_ih_{i+1})+B_{i+1}N_0(\rho_ih_{i+1})]}{j\omega\epsilon_{i+1}[J_1(\rho_ih_{i+1})+B_{i+1}N_1(\rho_ih_{i+1})]} \tag{14}$$

where ρ_i^+ approaches ρ_i from medium i and ρ_i^- approaches ρ_i from medium $i+1$. On equating Z_i^+ with Z_i^- an expression for B_{i+1} is obtained in terms of Z_i^+

$$B_{i+1} = -\left[\frac{h_{i+1}J_0(\rho_ih_{i+1})+j\omega\epsilon_{i+1}Z_i^+J_1(\rho_ih_{i+1})}{h_{i+1}N_0(\rho_ih_{i+1})+j\omega\epsilon_{i+1}Z_i^+N_1(\rho_ih_{i+1})}\right] \tag{15}$$

for $i=1,2,\cdots,n-1$. Furthermore, since

$$Z_{i+1}^+ = \frac{-h_{i+1}[J_0(\rho_{i+1}h_{i+1})+B_{i+1}N_0(\rho_{i+1}h_{i+1})]}{j\omega\epsilon_{i+1}[J_1(\rho_{i+1}h_{i+1})+B_{i+1}N_1(\rho_{i+1}h_{i+1})]} \tag{16}$$

for $i=1,2,\cdots,n-1$, (15) and (16) are recurrence formulas that determine Z_{i+1}^+ in terms of Z_1^+ $(i=1,2,\cdots,n-1)$.

From the radiation condition (the field components remain bounded at infinity), E_z in the earth region is given by the Hankel function of the second kind of order zero

$H_0^{(2)}$, and H_ϕ is given by the Hankel function of the second kind of order one $H_1^{(2)}$. Thus, $B_1 = -j$ and Z_1^+ is given by

$$Z_1^+ = \frac{-h_1 H_0^{(2)}(\rho_1 h_1)}{j\omega\epsilon_1 H_1^{(2)}(\rho_1 h_1)}. \tag{17}$$

Also, since the inner-most region $n+1$ is a good conductor with conductivity σ_{n+1}, its surface impedance is given by (5), thus

$$Z_n^- = \frac{E_z(\rho_n^-)}{H_\phi(\rho_n^-)} = (1+j)\left[\frac{\omega\mu_0}{2\sigma_{n+1}}\right]^{1/2}. \tag{18}$$

Thus, the modal equation is obtained by imposing the resonance condition

$$Z_n^- + Z_n^+ = 0 \tag{19}$$

where Z_n^+ is given by (16) with $i = n-1$ and Z_n^- is given (18).

D. Approximate Form of the Modal Equation

In order to facilitate the computations, an approximate analytical expression is derived in this section for the propagation coefficient γ. To this end, the modal equation (19) is expressed in terms of the surface impedances Z_n^- and Z_{n-1}^+.

On substituting (16) for Z_n^+ (with $i = n-1$) and (18) for Z_n^- into (19), the following exact expression is obtained for the modal equation in terms of Z_{n-1}^+ and Z_n^- at $\rho = \rho_{n-1}$ and $\rho = \rho_n$, respectively:

$$\begin{aligned}&[h_n J_0(\rho_n h_n) - j\omega\epsilon_n Z_n^- J_1(\rho_n h_n)]\\ &\quad\cdot[h_n N_0(\rho_{n-1}h_n) + j\omega\epsilon_n Z_{n-1}^+ N_1(\rho_{n-1}h_n)]\\ &= [h_n N_0(\rho_n h_n) - j\omega\epsilon_n Z_n^- N_1(\rho_n h_n)]\\ &\quad\cdot[h_n J_0(\rho_{n-1}h_n) + j\omega\epsilon_n Z_{n-1}^+ J_1(\rho_{n-1}h_n)].\end{aligned} \tag{20}$$

For the ideal coaxial cable $h_n = 0$. Thus, assuming small power loss at the boundaries ρ_n and ρ_{n-1}

$$|\rho_n h_n| \ll 1 \quad \text{and} \quad |\rho_{n-1}h_n| \ll 1 \tag{21}$$

and

$$|Z_n^-| \ll |\eta_n| \quad \text{and} \quad |Z_{n-1}^+| \ll |\eta_n| \tag{22}$$

where η_n is the intrinsic impedance for medium n

$$\eta_n = (\mu_0/\epsilon_n)^{1/2}, \qquad \mathrm{Re}\,\eta_n > 0. \tag{23}$$

On retaining first-order terms, (20) reduces to

$$\begin{aligned}&h_n[N_0(\rho_{n-1}h_n) - N_0(\rho_n h_n)]\\ &\quad= -j\omega\epsilon_n[N_1(\rho_n h_n)Z_n^- + N_1(\rho_{n-1}h_n)Z_{n-1}^+].\end{aligned} \tag{24}$$

Furthermore, since

$$\begin{aligned}[N_0(\rho_{n-1}h_n) - N_0(\rho_n h_n)] &\simeq \frac{2}{\pi}\ln(\rho_{n-1}/\rho_n)\\ N_1(\rho_n h_n)Z_n^- &\simeq \frac{-2}{\pi}\frac{Z_n^-}{\rho_n h_n}\\ N_1(\rho_{n-1}h_n)Z_{n-1}^+ &\simeq \frac{-2}{\pi}\frac{Z_{n-1}^+}{\rho_{n-1}h_n}\end{aligned} \tag{25}$$

it follows that

$$h_n^2 = \gamma^2 + k_n^2 \simeq j\frac{\omega\epsilon_n}{\ln(\rho_{n-1}/\rho_n)}\left[\frac{Z_n^-}{\rho_n} + \frac{Z_{n-1}^+}{\rho_{n-1}}\right] \tag{26}$$

and

$$\gamma = jk_n\left[1 - (h_n/k_n)^2\right]^{1/2}. \tag{27}$$

Noting that $k_n^2 = \omega^2\mu_0\epsilon_n$, hence $|h_n/k_n| \ll 1$, and

$$\left[1 - (h_n/k_n)^2\right]^{1/2} \simeq 1 - \frac{1}{2}(h_n/k_n)^2 \tag{28}$$

it follows that

$$\gamma = \alpha + j\beta \simeq jk_n + \frac{\omega\epsilon_n}{2k_n\ln(\rho_{n-1}/\rho_n)}\left[\frac{Z_n^-}{\rho_n} + \frac{Z_{n-1}^+}{\rho_{n-1}}\right]. \tag{29}$$

In the next section, the approximate form of the modal equation (29) is used to calculate the "effective" surface impedance $Z_{\mathrm{eff}} = Z_{n-1}^+$ from measured values of α.

E. Effective Surface Impedance

The leaky coaxial cable used in this study is manufactured by Andrew Corporation, and is referred to as RADIAX slotted coaxial cable. The inner conductor is a copper coated, solid aluminum wire. Because of the thickness of the copper coating (0.1 mm) (45 skin-depths at 0.9 GHz), the inner conductor is considered to be solid copper. The copper outer conductor is slightly corrugated to provide flexibility to the cable. The difference between the maximum and minimum radii of the outer conductor (peaks and valleys of the corrugations) is very small (0.762 mm) compared to a wavelength in the cable λ_c ($\lambda_c = 2\pi/\beta \simeq 23.6$ cm at 0.9 GHz). Therefore, for the model, ρ_{n-1} was assumed to be given by its average value $\rho_{n-1} = \rho_4 = 0.6223$ cm and $\rho_5 = 0.21717$ cm.

Electromagnetic power is coupled into the environment, through slots punched through the outer conductor of the cable (region $n-1$). Since the dimensions of the slots (3×6 mm) and the spacing between them (5 mm) are small compared to λ_c, the region $\rho \geqslant \rho_{n-1}$ is characterized by an effective surface impedance $Z_{\mathrm{eff}} = Z_{n-1}^+$ at $\rho = \rho_{n-1}$.

To calculate the "effective" permittivity ϵ_{eff} for the region $\rho \geqslant \rho_{n-1}$, the approximate form of the modal equation (29) is used. The wavenumber for the dielectric region n is

$$k_n = \omega(\mu_0\epsilon_n)^{1/2}, \qquad \mathrm{Re}(k_n) \geqslant 0, \epsilon_n = \epsilon_0(1.6 - j0.0008). \tag{30}$$

The impedance Z_n^- is given by (18), and the surface impedance Z_{n-1}^+ which characterizes the entire region $\rho > \rho_{n-1}$ depends upon the soil moisture content

$$Z_{n-1}^+ = R_{n-1} + jX_{n-1}. \tag{31}$$

Since the power per unit length coupled through the slots is very small compared to the total power transmitted through the cross section of the cable, the outer region $\rho > \rho_{n-1}$ is characterized by a good conductor. Thus

$$\epsilon_{\mathrm{eff}} = \epsilon'_{\mathrm{eff}} - j\epsilon''_{\mathrm{eff}} \simeq -j\epsilon''_{\mathrm{eff}} = -j\sigma_{\mathrm{eff}}/\omega \tag{32}$$

where $\sigma_{\text{eff}} = \omega\epsilon''_{\text{eff}}$ is the effective conductivity of the outer region. Therefore, the effective skin-depth is given by

$$\delta_{\text{eff}} = (2/\omega\mu_0\sigma_{\text{eff}})^{1/2} = (2/\mu_0\epsilon''_{\text{eff}})^{1/2}/\omega. \quad (33)$$

It follows that

$$X_{n-1} = R_{n-1} \simeq \text{Re}(\mu_0/\epsilon_{\text{eff}})^{1/2} = (\mu_0/2\epsilon''_{\text{eff}})^{1/2} = (\omega\mu_0/2\sigma_{\text{eff}})^{1/2}. \quad (34)$$

Thus, (29) reduces to

$$\gamma = \alpha + j\beta \simeq jk_n + \frac{(1+j)\omega\epsilon_n}{2k_n \ln(\rho_{n-1}/\rho_n)}\left[\frac{R_n}{\rho_n} - \frac{R_{n-1}}{\rho_{n-1}}\right]. \quad (35)$$

Take the real part of (35) and solve for R_{n-1} to get

$$R_{n-1} = \rho_{n-1}\left[\frac{\alpha - \text{Re}(jk_n)}{\text{Re}\left[\frac{(1+j)\omega\epsilon_n}{2k_n\ln(\rho_{n-1}/\rho_n)}\right]} - \frac{R_n}{\rho_n}\right] = (\mu_0/2\epsilon''_{\text{eff}})^{1/2} \quad (36)$$

and the effective surface impedance Z^+_{n-1} is given by

$$Z_{\text{eff}} = Z^+_{n-1} = R_{n-1} + jX_{n-1} = (1+j)R_{n-1}. \quad (37)$$

It should be pointed out that since Z^+_{n-1} is the surface impedance at $\rho = \rho_{n-1}$ (looking in the direction of increasing ρ) ϵ_{eff} accounts for the effects of the slotted outer conductor (region $n-1$), the flooding compound, (region $n-2$), the dielectric outer jacket (region $n-3$), and the soil (region $n-4=1$).

III. Experimental Work

To determine if the radio-wave method for sensing soil moisture could be practically implemented, a series of preliminary laboratory experiments was conducted. The results of these laboratory experiments were used to determine the experimental setup in the field where a series of measurements were made over an extended period.

The scaled microwave model used in the laboratory experiments consisted of a short length of leaky coaxial cable buried in sand. Three different types of cables were used to determine the most suitable choice for field applications. Furthermore, two methods of measurement were employed to determine the most feasible for use in the field. The first method considered was the transmission method, in which changes in the transmission coefficient S_{21} were measured. The second method, the reflection method, based on the measurement of changes in the input reflection coefficient

$$\Gamma_{\text{in}} = S_{11} + S_{12}S_{21}\Gamma_L/(1 - S_{22}\Gamma_L). \quad (38)$$

The locus of the transmission coefficient S_{21} as a function of θ_m is a spiral centered at the origin of the complex S_{21}-plane (Argand diagram). Changes in the phase of S_{21}, $\Delta\phi_T = -\Delta\beta l$ (where $\Delta\beta$ is the change in the propagation coefficient and l is the length of the buried cable) can be measured directly by a vector voltmeter.

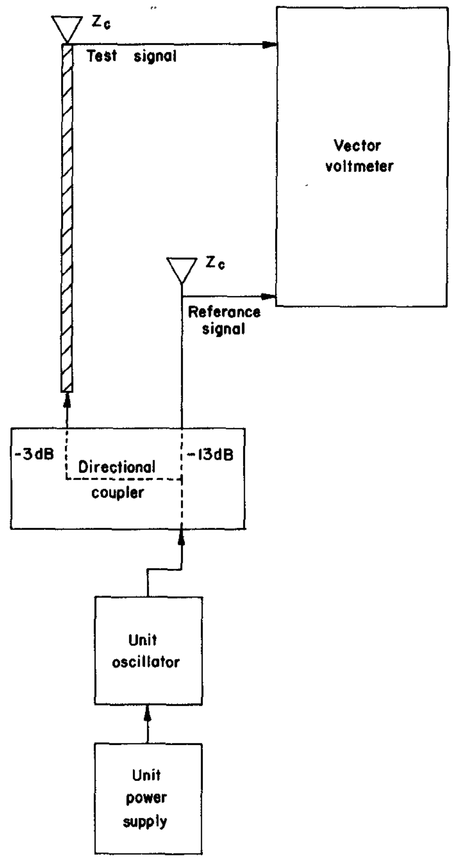

Fig. 5. Block diagram of experimental arrangement in the field.

The locus of the reflection coefficient Γ_{in} is a spiral displaced from the origin of the complex plane by the quantity S_{11}. The complex quantities S_{11} and S_{22} characterize the reflecting properties of an impedance-matched coaxial cable and Γ_L is the load reflection coefficient. For an ideal coaxial cable $S_{11} = S_{22} = 0$; however, due to the corrugations in the outer conductor of the cable (to provide flexibility), S_{11} is not negligible. The complex quantity S_{11} must be subtracted from the value of Γ_{in} to measure the phase change $\Delta\phi_R = -2\Delta\beta l$ directly. In the laboratory this can be accomplished through a translation of the origin of the polar display unit of a network analyzer. However, since a vector voltmeter was used as the detector in the field, it was decided to conduct the field work only in the transmission mode.

The field experiment was conducted with a slotted coaxial cable 110 m in length buried at a depth of 0.5 m, and excited by a 0.9-GHz signal. Based on the results of the laboratory work [13], the type of slotted cable selected was RADIAX RX4-1. A block diagram of the experimental arrangement is shown in Fig. 5. With the use of a directional coupler, the 0.9-GHz signal is transmitted through the buried cable to the vector voltmeter where it is compared with the reference signal. The buried cable is terminated by its characteristic impedance Z_c.

The phase difference between the test and reference signals was recorded automatically. All the electrical equipment was housed inside an instrument trailer. The instrument trailer was located at the edge of the experimental

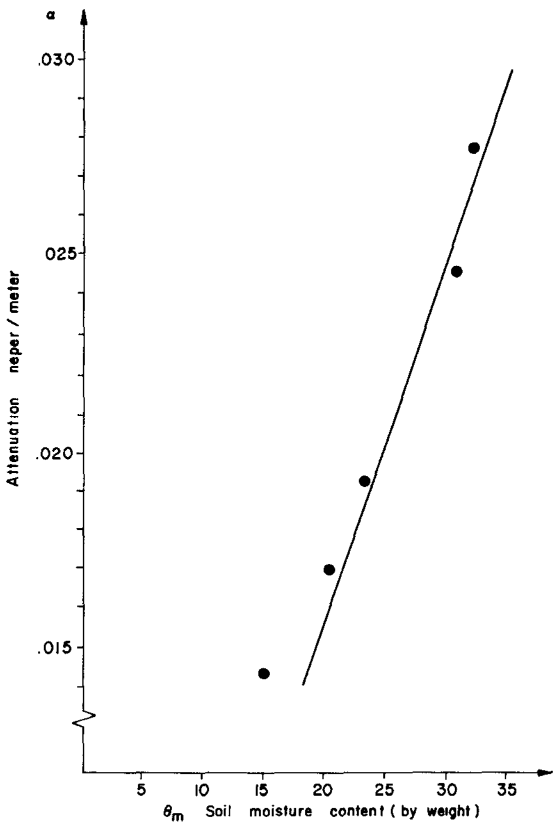

Fig. 6. Measured attenuation as a function of soil moisture content ($f = 0.9$ GHz).

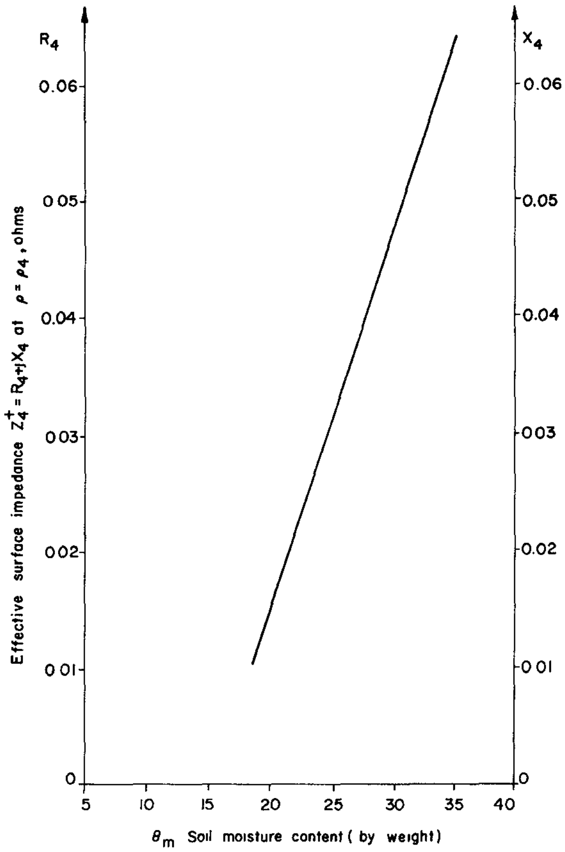

Fig. 7. Calculated effective surface impedance $Z_4^+ = R_4 + jX_4 = (1+j)R_4$ as a function of soil moisture content ($f = 0.9$ GHz).

site, minimizing the need for transmission lines between the equipment and the two ports of the buried cable.

The field site was planted with corn. The plants extracted some of the available water from the soil as in any practical situation. Gypsum resistance blocks were buried at 12 different locations over the experimental site. The gypsum blocks were placed at depths of 15, 45, and 75 cm. Measurements obtained from the gypsum blocks provided a measure of the average matric potential at the specified depths. These measurements, together with the data in Fig. 2, are used to determine the soil moisture content θ_m.

The attenuation α appearing in (36) is the total attenuation per unit length (due to absorption in the conducting walls and the dielectric ϵ_n, as well as the power radiated through the slots). By measuring α for the range of pertinent soil moisture conditions, the corresponding effective surface impedance Z_{n-1}^+ may be calculated using (36) and (37). Fig. 6 illustrates the measured attenuation of the cable as a function of θ_m. The corresponding variation of Z_{n-1}^+ as a function of θ_m is illustrated in Fig. 7.

A computer program was developed to evaluate $\gamma = \alpha + j\beta$ (as a function of moisture content) for the dominant azimuthally symmetric mode supported by the buried leaky coaxial cable. The exact form of the modal equation (20) was solved using a root-seeking algorithm based on Muller's method for complex functions [12] (see Fig. 8). Thus, the modal equation (20) and its approximate form (29) are in excellent agreement.

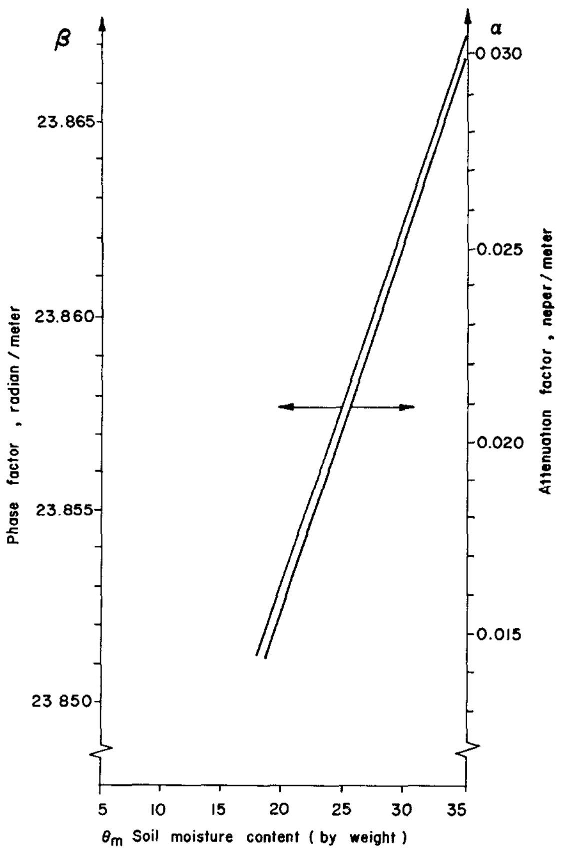

Fig. 8. Calculated propagation coefficient $\gamma = \alpha + j\beta$ (see (20), as a function of soil moisture content ($f = 0.9$ GHz).

The experimental work conducted in the field demonstrated that even sudden changes in the soil moisture

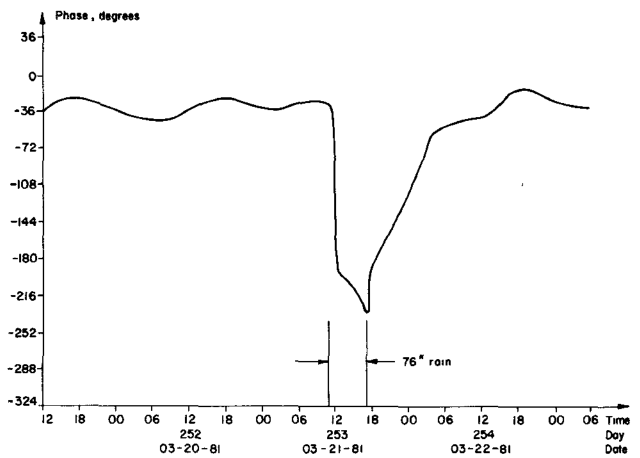

Fig. 9. Phase change $\Delta\phi_T = -\Delta\beta l$ due to rainfall.

content could be detected using the experimental arrangement shown in Fig. 5. When the soil moisture content increased (due to rainfall), a negative phase shift was recorded, and the amplitude of the test signal decreased. This corresponds to an increase in the phase factor β and the attenuation α. In addition, changes in the phase factor $\Delta\beta$ and the attenuation factor $\Delta\alpha$ were approximately equal, as predicted by the analysis (see Fig. 8). After a rainfall, the soil would begin to dry gradually. However, due to the shortage in rainfall during the season these measurements were conducted, the soil would dry up rapidly. As the soil moisture content decreased, a positive phase shift $\Delta\phi_T$ was recorded and the amplitude of the test signal increased. A typical example of this is shown in Fig. 9.

During periods of gradual changes in soil moisture, the phase measurement was seen to contain two forms of drift unrelated to changes in the soil moisture content. A principal source of phase drift was ambient temperature variations. The temperature variation of the soil (in the vicinity of the cable) affects the complex permittivity of the soil. This effect is negligible for temperatures between 15°C and 30°C [5]. However, variations in the ambient temperature of the buried cable directly affect its electrical properties. As the temperature of the cable changes, its physical length changes due to expansion or contraction of the copper used in its construction. In addition, the permittivity of the dielectric in the cable changes. These two effects combine to produce a change in the electrical length of the cable. The change in the cable's electrical length with temperature is characterized by the manufacturer by a phase-temperature coefficient. The phase-temperature coefficient can be reduced significantly by the manufacturer (see Conclusion). Another source of phase drift was frequency fluctuations of the signal of the unit oscillator. This effect could be minimized by replacing the oscillator used in the experiments by a frequency stabilized source.

IV. Discussion and Conclusion

The experimental investigation conducted in the field demonstrated that changes in the soil moisture content can be detected using a buried leaky coaxial cable as a sensor. The RADIAX cables were chosen since they did not absorb water and were not damaged by rodents under field conditions. Based on the laboratory work, the slotted cable with the smallest nominal attenuation (smallest slots), was found suitable for use in the field. The transmission method was used in the field work to sense changes in γ. For a change in soil moisture content $\Delta\theta_m$ there is a corresponding change in the complex propagation coefficient $\Delta\gamma = \Delta\alpha + j\Delta\beta$. An increase in θ_m resulted in a corresponding increase in α and β. Moreover, it was observed that the changes in the attenuation and phase factors were approximately equal ($\Delta\alpha \simeq \Delta\beta$).

The slotted cable was buried deep enough so that the air–earth interface could be ignored in the analysis. Since the measurements were only sensitive to changes in the soil moisture in a region of several skin-depths around the cable (10–20 cm) changes in soil moisture at the surface were not sensed. This was evident when light sprinkles of rain soaked only the top soil. In this case, the phase and attenuation measurements were unaffected, as were the independent measurements taken from the gypsum blocks.

The characteristic (modal) equation for the leaky coaxial cable is expressed in terms of the surface impedances at the inner conductor and an effective surface impedance that characterizes the region of the slotted outer conductor, the flooding compound, the outer protective jacket, and the earth. A computer program was developed to evaluate the principal root of the modal equation. The results of the analysis were seen to be in agreement with the observed behavior of the physical system.

There are four parameters which must be considered in the design of the radio-wave soil moisture sensing system, the length of the buried cable, its depth, frequency, and the type of cable (sensor).

To provide an indication of the soil moisture content averaged over a large portion of an irrigated field, a very long cable is required (around 1200 m). As a rule, farmers are primarily interested in the available water to the plant roots. Woody crops such as corn or sorghum have primary roots which extend to depths between 1–3 m [2]. Variations in the soil moisture content of the top soil are of little importance, except perhaps at the beginning of the growing season. Thus, the cable should be buried at the approximate depth of the mature plant roots, or at least several skin-depths deep so that effects of the air–earth interface are insignificant. The signal frequencies have both an upper and a lower limit. The lower limit is determined by the requirement that $d > \delta_1$ (the skin-depth in the soil). The maximum amount of attenuation in the cable that can be tolerated determines the upper frequency limit. The relationship between frequency, length, and depth for the type of slotted cable used in this work is shown in Table I. The length has been chosen to yield a maximum phase shift of π radians over the range of available moisture. The maximum and minimum skin-depths (occurring at wilting point and field capacity, respectively), are also listed. The depth has been chosen to be 150 percent of the maximum skin-depth. The RADIAX RX4-1 cable was selected in this work (Section III).

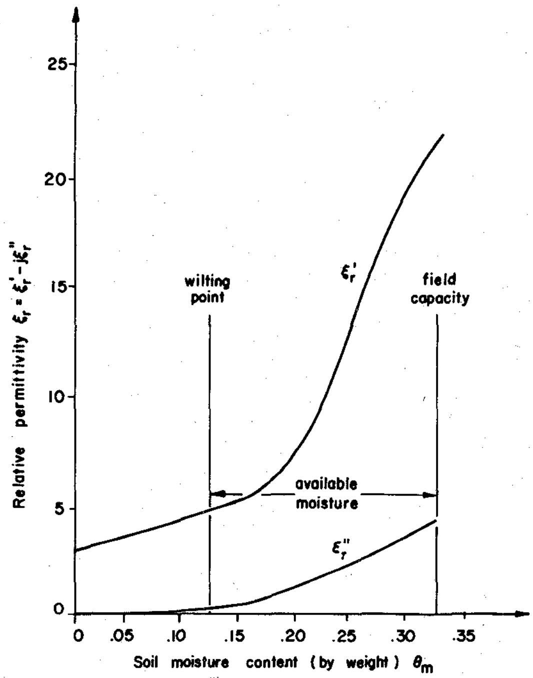

Fig. 10. Relative permittivity of clay loam as a function of soil moisture content at 1.4 GHz (from Newton [12]).

TABLE I

Frequency f GHz	Skin-depth m $\delta_{1,max}$	$\delta_{1,min}$	Length ℓ m	Depth m
0.1	2.10	0.30	2100	3.15
0.2	1.05	0.15	1050	1.58
0.3	0.71	0.10	700	1.07
0.4	0.53	0.08	525	0.80
0.5	0.42	0.06	420	0.65
0.6	0.35	0.05	350	0.55
0.7	0.30	0.04	300	0.45
0.8	0.27	0.04	260	0.40
0.9	0.25	0.04	230	0.38
1.0	0.21	0.03	210	0.32
1.1	0.19	0.03	190	0.29
1.2	0.18	0.02	175	0.26

Farmers are interested in the amount of available water to plants, rather than the soil moisture content. In the research conducted by Newton [5] and by Silva *et al.* [1], it was demonstrated that most soils can be characterized by a critical moisture content (near wilting point) at which the slope of the complex permittivity changes significantly. The transition point occurs at different soil moisture contents for different soil types, and their data suggests that a piecewise linear model for the soil permittivity is reasonable. Some of the date obtained from Newton's research at 1.4 GHz [5] is shown in Fig. 10. Note that the wilting point occurs just below the transition point. Usually, the greatest rate of change in the complex permittivity of soils occurs within the range of available water to plants. Since the radio-wave system senses the changes in soil permittivity, it provides an excellent method of monitoring the amount of available water to plants.

It was seen that the oscillator of the radio-wave sensing system must be frequency stable to minimize random phase fluctuations. If it is desired to monitor the changes in the attenuation coefficient, the power output must remain constant with time. To minimize the diurnal phase drift due to variations in soil temperature, a phase stabilized cable should be used to perform the sensing. The manufacturer of the slotted cable used in this work is presently capable of producing such cables.

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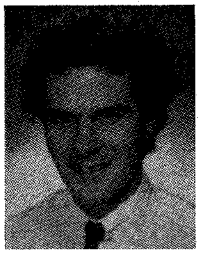

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Waves Guided by Conductive Strips Above a Periodically Perforated Ground Plane

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***Abstract*—This paper considers the propagation of waves along an array of conductive strips situated above a periodically perforated conductive plane. Each conductor has zero thickness and finite sheet resistance, and the dielectric is homogeneous. The surface current density on the conductors is approximated by a finite number of current elements having rooftop spatial dependence. The transverse electric field is expressed in terms of the current, and the electric field boundary condition is satisfied in an integral sense over the conductors. This generates a matrix equation whose solution gives the dispersion curve relating the propagation constant to frequency, as well as the current distribution.**

The simulation results are used to obtain equivalent transmission-line parameters applicable to printed circuit boards found in high-performance computers. A characteristic impedance is defined and it is shown that, with proper interpretation, the uniform transmission-line equations for propagation constant and characteristic impedance apply to such computer packages. The coupling between adjacent strips is calculated, and the effect of finite resistivity discussed.

I. Introduction

TRANSMISSION LINES in the form of conductive strips (signal lines) embedded in a dielectric and sandwiched between conductive planes are used in high-performance computers to carry signals between integrated-circuit chips. Often, many layers of conductor and dielectric are integrated into a compact package or printed circuit board [1], and may require interconnections between signal lines located on different layers. Arrays of holes are therefore made in the conducting planes and dielectric so that conductive elements can be inserted to electrically connect signal lines situated on different layers. The resulting transmission-line structure is, then, an array of signal lines situated between and insulated from ground planes which are perforated periodically with apertures.

One might assume that the structure can be approximated as a uniform transmission line. By calculating the appropriate capacitance, inductance, and resistance, through available means [2], [3], the propagation characteristics could then be determined. However, the validity of such an approximation must remain questionable until the structure is accurately analyzed.

In this paper, we present a numerical analysis based on rigorous electromagnetic theory for propagation along an array of signal lines situated above a single perforated ground plane in a homogeneous dielectric. We chose to consider an array of signal lines because the composite structure has two-dimensional periodicity and previous results can be applied. However, if adjacent signal lines are sufficiently separated to reduce their interaction, the resulting current distribution approaches that for an isolated signal line situated above the ground plane. Numerical results are given for the propagation constant, characteristic impedance, and the coupling between adjacent lines.

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