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Scattering from Perfectly Conducting and Resistive Strips on a Grounded Dielectric Slab

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Abstract—The scattering properties of perfectly conducting and resistive strips are predicted for strips which are located on a dielectric slab backed by a perfectly conducting ground plane. The spectral domain Green’s function is used to relate the currents and fields on the strip, and the resulting integral equation is solved using the method of moments. Both TE and TM strips are examined using piecewise linear and pulse subdomain basis functions, respectively, to model the current on the strip. Calculated results are compared with results measured at the NASA Langley Research Center.

I. INTRODUCTION

The scattering properties of metallic strips in free space are well known and are discussed in many textbooks on electromagnetic theory. Balanis [1] discusses scattering from perfectly conducting strips in free space with both TE and TM polarizations. A variety of methods have been used in the past to examine such strips. These methods include physical optics, the physical theory of diffraction, and the integral equation/method of moments solutions.

In recent years, the study of nonperfectly conducting strips has received much attention. Senior [2], [3] has examined scattering from resistive strips with the use of diffraction techniques. Senior [4], [5] has also formalized the boundary conditions needed to correctly model resistive strips. Senior and Liepa [6] have used diffraction techniques in order to study strips with a nonconstant, or tapered, resistivity. Ray and Mittra [7] have used the method of moments to analyze strips with constant resistance loading on the strip edges. Haupt and Liepa [8]-[10] have shown that for resistive strips in free space, physical optics solution methods can give results that are close to the results given by method of moments solutions for certain cases, and the resistive taper placed on the strip can be used to control the scattering performance of the strip. Hall and Mittra [11] have extended the basic moment method analysis to include strips in an infinite array environment. Peters and Newman [12] have examined TM scattering by a resistive sheet located in a dielectric slab or on a dielectric half space with the use of the method of moments and spectral domain Green’s functions. Bailey [13] has used similar techniques in order to examine perfectly conducting strips in a dielectric slab backed by a ground plane as a way of predicting the properties of microstrip antennas. Superconducting materials can exhibit properties similar to those of resistive strips, and have recently been examined [14], [15].

This paper is concerned with the analysis of perfectly conducting and resistive strips placed on a dielectric slab which is backed by a perfectly conducting ground plane. The method of moments is used to calculate the scattering on the strip.

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to solve the electric field integral equation derived for the field on the strip. The spectral domain Green’s function is used to relate the currents and fields on the strip. Both TE and TM strips are examined, and calculated results are compared to results measured at the NASA Langley Research Center (LaRC). Although only single strips will be examined, the method may also be extended to include strips in an infinite array environment.

II. THEORY

A. TM Polarization

The geometry of an infinite strip is shown in Fig. 1. The strip is located on an infinite dielectric slab which is backed by a perfectly conducting ground plane. The strip is infinite in the direction and has width \( 2L \), in the \( z \) direction. The electric field is incident from the angle \( \theta \) measured from the \( z \) axis, and for TM polarization has only a \( \gamma \) component. Note that the angle \( \theta \) in Fig. 1 takes on both positive and negative values. In order to relate the tangential electric field on the strip to the current density on the strip, the boundary condition for the electric field on a resistive sheet is used. This is given by

\[
\mathbf{E}_{\text{tan}} + \mathbf{E}_{\text{scat}} = R_s \mathbf{J}
\]

where \( \mathbf{E}_{\text{tan}} \) is the tangential field at the strip location due to an incident plane wave, \( \mathbf{E}_{\text{scat}} \) is the scattered field due to the current on the strip, \( R_s \) is the surface resistance of the strip, and \( \mathbf{J} \) is the current density on the strip. The scattered field can be found by using the spectral domain Green’s function for a current element radiating on a grounded dielectric slab, and is given by

\[
\mathbf{E}_{\text{scat}}(x, y, z) = \int \int \int \mathbf{G}(x, y, z | x', y', z') \cdot \mathbf{J}(x', y', z') \, dx' \, dy' \, dz'
\]

where each component of the dyadic Green’s function is given by

\[
G_{ab} = \frac{1}{4\pi^2} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \tilde{G}_{ab} \cdot (k_x, k_y, d | d) e^{j k_x (x - x')} e^{j k_y (y - y')} \, dk_x \, dk_y
\]

and, in general, \( a \) and \( b \) can be each \( x, y, \) or \( z \). The exact terms of the Green’s function can be found in the literature, along with additional forms of the Green’s function which account for a dielectric covering above the strip [16].

The current density on the strip is completely \( \gamma \) polarized and is modeled as a summation of pulse subdomain basis functions. Mathematically, this can be written as

\[
J_y = \sum_{m=1}^{M+1} \tilde{I}_m^y \Pi^m(x')
\]

where \( M + 1 \) is the number of subdomains on the strip and \( \tilde{I}_m^y \) is the amplitude of the \( m \)th subdomain. The pulse subdomain function \( \Pi^m(x') \) is defined as

\[
\Pi^m(x') = \begin{cases} 1 & (x_m - \Delta x) \leq x' \leq x_m \\ 0 & \text{otherwise} \end{cases}
\]

where \( \Delta x = 2L/(M + 1) \) and \( x_m \) is the coordinate of the \( m \)th subdomain.

After applying Galerkin’s technique, the terms of the impedance and resistance matrices are found in a straightforward manner, and a system of simultaneous equations results. However, in order to solve the system of equations, the terms of the excitation vector must also be evaluated. In general, the elements of the excitation vector are given by

\[
V_p^y = \int_{x_p - \Delta x}^{x_p} \tilde{I}_m^y \Pi^m(x') \cdot E_{0y} \, dx
\]

which is the reaction of the current on the strip with the incident field. Using reciprocity, this can be rewritten as

\[
V_p^y = -\frac{1}{Z_0} \left( \frac{8\pi}{j\omega} \right)^{1/2} E_p^{\gamma} \cdot E_0
\]

where the factor \(-1/Z_0(8\pi/j\omega)^{1/2}\) is the required strength of a line source to produce a unit amplitude plane wave and the factor \( E_p^{\gamma} \) has been suppressed. The incident field vector \( E_0 \) is assumed to be \( \gamma \) polarized and have unity amplitude. The \( E_p^{\gamma} \) term is the electric field at the point \((x, y, z)\) radiated by the \( p \)th current mode on the strip and can be evaluated by the method of stationary phase. The resulting expression for the electric field is

\[
E_p^{\gamma}(r, \theta) = \frac{-jZ_0 I_p^{\gamma}}{\sqrt{2\pi k_o}} \frac{e^{-j k_o r}}{\sqrt{r}} \cdot e^{j k_z (k_z d)} \cos \theta \sin (k_d d) \Pi_{p}^{\gamma}(k_d)
\]

where \( k_z \) is evaluated at the stationary phase point \( k_z = -k_o \sin \theta \). \( \Pi_{p}^{\gamma}(k_z) \) is the Fourier transform of the \( p \)th current mode, and \( k_d \) and \( k_z \) are the \( z \)-axis propagation constants in the dielectric and free space regions, respectively.

B. TE Polarization

For TE polarization, the electric field is incident from the angle \( \theta \) measured from the \( z \) axis and has only \( \gamma \) and \( \delta \) components. The current on the strip is completely \( \delta \) polarized, and is modeled as a summation of piecewise linear subdomain basis functions. This is written as

\[
J_x = \sum_{m=1}^{M} \tilde{I}_m^x \Lambda^m(x')
\]
where $M$ is the number of subdomains on the strip and $\Gamma_n^e$ is the amplitude of the $n$th subdomain. The piecewise linear subdomain function $\Lambda_n^m(x')$ is given by

$$
\Lambda_n^m(x') = \begin{cases} 
1 & x_m - \Delta x \leq x' \leq x_m \\
1 & x_m \leq x' \leq x_m + \Delta x \\
0 & \text{otherwise}
\end{cases}
$$

(10)

where $\Delta x = 2L/|M + 1|$. The elements of the excitation vector are similar to those described above, except the field $E^e_0$ is now $\theta$ polarized and is given by

$$
E^e_0(r, \theta) = \frac{-j \rho_0 E_0}{\sqrt{2 \pi k_e}} \left[ \frac{e^{-j k_e r}}{\sqrt{r}} - \frac{e^{-jk_e r}}{\sqrt{r}} \right]
$$

(11)

where $k_e$ is evaluated at the stationary phase point $k_x = -k_e \sin\theta$ and $\Lambda_n^m(k_e)$ is the Fourier transform of the $n$th current mode.

III. RESULTS

The matrix equations derived above for strips on a grounded dielectric slab have been solved for a number of test cases. For all of the test cases, the strip width was fixed at 2.0 in (5.08 cm) and the dielectric thickness was fixed at $d = 0.0787$ cm with a dielectric constant $\varepsilon_r = 2.33$ and a loss tangent $\delta = 0.001$. The results measured in the Experimental Test Range (ETR) at NASA LaRC are the radar cross section in units of dB/m$^2$ (dBsm) and have been transformed into scattering width by the relation [1]

$$
\sigma_{2DD} \approx \frac{\sigma_{3D}}{2d}
$$

(12)

where $l$ is the length of the strip being measured. The measured strips were made to be as long as possible in order to approximate the response of an infinite strip.

Initially, tests were performed on strips with a surface resistance which remained constant over the width of the strip. For TM polarization, the monostatic scattering width as a function of frequency is shown in Fig. 2 for a perfectly conducting (PEC) strip and a strip with a 55 $\Omega$ surface resistance. The incidence and scattering angles, shown in Fig. 1, were both fixed at $\theta' = \phi' = 90^\circ$. Close agreement between the measured and calculated data is noted in both cases. However, even with a surface resistance of 55 $\Omega$, the level of scattering has not been reduced significantly. Additional results are shown in Fig. 3 for strips which have a constant surface resistance of 377 and 754 $\Omega$ and also for strips which have a parabolic tapered resistance. The tapered resistive strips are both perfectly conducting in the center of the strip, with an increase in surface resistance towards the edges which reaches a peak of 377 $\Omega$ for one strip and 754 $\Omega$ for the second strip. As before, the incidence and scattering angles were fixed at $\theta' = \phi' = 90^\circ$. It is interesting to note that the strip with the higher resistive taper exhibits a larger scattering width across much of the frequency band. With the frequency fixed at 12.0 GHz, the bistatic scattering patterns for these strips are shown in Fig. 4. In this case, the incidence angle is fixed at $-90^\circ$ and the scattering width versus angle is shown for $\phi' = -90^\circ$ to $+90^\circ$. The strips with constant resistive loading show similar scattering patterns as that of the PEC strip, with the expected decrease in amplitude. The tapered strips, however, show different scattering patterns, with the more highly loaded strip showing a larger scattering width over most of the pattern.

Tests have also been performed on strips excited with TE polarization. The monostatic scattering width as a function of frequency for a PEC strip and an 11 $\Omega$ strip are shown in Fig. 5. As seen in Fig. 5, the response of the PEC strip contains a number of resonant peaks which have been totally suppressed, with a surface resistance of only 11 $\Omega$. Fairly good agreement between the measured and calculated values is seen for both of these cases. Additional results are shown in Fig. 6 for strips with constant surface resistance values of 377 and 754 $\Omega$ and also tapered resistance strips similar to those described above. The frequency responses of these strips, along with the PEC strip response, are shown in Fig. 6. The strips which have constant surface resistance show responses similar to those of TM polarization. Also, as was the case for TM polarization, the tapered strip with the higher peak resistance exhibits a higher scattering width across most
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Fig. 4. Bistatic scattering width for strips located on a grounded dielectric slab—TM polarization (frequency = 12 GHz, \( L_z = 2.54 \) cm, \( d = 0.07874 \) cm, \( \varepsilon_r = 2.33 \), loss tangent = 0.001, \( \theta^1 = -60.0^\circ \)).

Fig. 6. Comparison of calculated monostatic scattering width as a function of frequency for strips located on a grounded dielectric slab TE polarization (\( L_z = 2.54 \) cm, \( d = 0.07874 \) cm, \( \varepsilon_r = 2.33 \), loss tangent = 0.001, \( \theta^1 = \theta^2 = -60.0^\circ \)).

Fig. 7. Bistatic scattering width for strips located on a grounded dielectric slab TE polarization (frequency = 12 GHz, \( L_z = 2.54 \) cm, \( d = 0.07874 \) cm, \( \varepsilon_r = 2.33 \), loss tangent = 0.001, \( \theta^1 = -60.0^\circ \)).

for a thin resistive sheet was used, along with the spectral domain Green's function, to derive an integral equation for the electric field on the strip. This method accurately accounts for the thickness and dielectric properties of the slab, as well as the width and resistive properties of the strip. Both TE and TM polarizations have been examined, and both monostatic and bistatic results have been presented. Calculated and measured results of monostatic scattering width have been presented for perfectly conducting strips and for strips with a constant surface resistance profile, and were seen to be in good agreement. Calculated results have also been presented for strips with a parabolic surface resistance taper. Interestingly, the strips with the higher surface resistance taper showed higher levels of scattering width.

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REFERENCES

Human Operator Coupling Effects on Radiation Characteristics of a Portable Communication Dipole Antenna

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Abstract—EM coupling effects of a human operator on antenna radiation characteristics, such as the antenna input impedance, radiation pattern, the radiation power (into free space), the power absorbed by the body, the radiation efficiency, etc., of a portable communication dipole antenna were investigated in detail. A realistically shaped 3-D man model and a proximate linear dipole antenna were used to model this problem. Coupled integral equations (CIE) and the method of moments (MoM) were employed to numerically solve this antenna-body coupling problem. Numerical examples are presented for the antenna located in front of the head (distance ranging from 5 to 1 cm) or adjacent to the abdomen (0.6 cm distance) at 840 MHz. It is found that, when coupled with the operator body, the antenna input impedance will have significant deviation from those in free space and different positions. Due to the operator body absorption effect, the maximum attenuation of the H-plane antenna gain may reach about 15 dB for the antenna at the head position and 25 dB for the abdomen position, toward the direction of the body side. Also, the antenna radiation efficiency is reduced to the range from 0.72 to 0.29 for the head position and 0.15 for the abdomen position, respectively. Moreover, the cross-polarization field is significant, especially in the E-plane of $\phi = 90^\circ$. This is important for the antenna RF design and communication link budget consideration of portable radio systems.

I. INTRODUCTION

Recently, portable communication systems (including two-way portable radios and portable cellular phones) have been extensively utilized by the general public. In these present and future portable communication systems, the carrier frequencies are all very high, and reach microwave frequency bands. For example, the current cellular phone systems are using an 800-900 MHz band. At such a high frequency, since the operator is very near the vicinity of the radiating antenna of the portable handset, the electromagnetic (EM) coupling between the antenna and the nearby body will significantly affect the antenna characteristics. These antenna characteristics include the antenna input impedance which varies from face to torso positions, the antenna current distribution, the antenna radiation pattern, and the antenna radiation efficiency which is reduced due to the radiated power absorbed by the body. Hence, in order to design a high-performance portable antenna, the EM coupling between the antenna and the operator body must be fully counted and analyzed in antenna design optimization. Moreover, the level of EM energy deposition from the radiating antenna into the nearby operator body (especially the head) have become a public concern as a potential health hazard. Hence, it is also necessary to analyze the RF dosimetry problem. Extensive research results have been reported on the energy deposition in a human body model exposed to the dipole antenna fields [1]-[3]. However, the detailed EM coupling effects between portable communication dipole antennas and a realistically shaped full-body model at the portable communication band (800-900 MHz) have been unavailable. To date, the effects of the EM coupling

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