TABLE I
COMPUTED NOISE PARAMETERS FOR THE FUJITSU FHX01A HEMT

	Common-Emitter	Common-Base	Common-Collector
Fopt (min NF)	2.800 (522 deg)*	3.000 (580 deg)	2.758 (510 deg)
R	34.0*	36.9	58.7
Gopt	.01919*	.01890	.01187
Bopt	.01509*	.01278	.01693
Avail Gain	5.416 (7.34 dB)	10.49 (10.2 dB)	1.867 (2.71 dB)
M	2.208 (640.2 deg)	2.21 (641.0 deg)	3.79 (1098 deg)
Mmin	2.20134 (638 deg)	2.20134 (638 deg)	2.20134 (638 deg)
Gs	.01927	.01732	.00460
Bs	.01668	.01172	.01403
Avail Gain	5.547 (7.44 dB)	11.34 (10.5 dB)	-6.52 (see text)
F	2.804 (523 deg)	3.007 (582 deg)	2.864 (540 deg)

*These data, and the common-emitter S parameters (below), taken from the data sheet for the transistor, were used to calculate the rest of the table:

$$S_{11} = -0.0773 + j0.51119 \qquad S_{12} = 0.03388 - j0.09515$$
$$S_{21} = 0.1840 - j1.4170 \qquad S_{22} = 0.66384 + j0.21569.$$

V. CONCLUSION

Formulas were derived to transform noise parameters when the terminals of a three-terminal amplifier are interchanged. It was shown that the minimum noise measure must be the same for the common-emitter, common-base, and common-collector configurations. A practical example was given to confirm this invariance. High-gain amplifiers with the minimum noise figure can be built with any of the three configurations or combination thereof. The choice of configuration can be (and is) determined by factors such as ease of stabilization or bandwidth.

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Microwave Shielding Effectiveness of EC-Coated Dielectric Slabs

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***Abstract*—The purpose of this paper is to derive correct formulas for the microwave shielding effectiveness (SE) of a thin metallic layer deposited on top of a dielectric slab. For coatings much thinner than the skin depth, the following holds: (a) In a half-wave geometry, SE is a function of the sheet resistance only, SE (in dB) $= 20 \times \log(1 + 188.5/R_s)$ if R_s is in ohms per square; (b) in a quarter-wave geometry, SE (in dB) $= 20 \times \log[(1 + \epsilon_r)/(2\sqrt{\epsilon_r}) + 188.5/(\sqrt{\epsilon_r}\, R_s)]$, where ϵ_r refers to the dielectric constant of the substrate. These formulas provide upper and lower limits for the effective shielding performance of an electroconductively coated dielectric slab.**

I. INTRODUCTION

Thin metallic films or stacks deposited upon glass substrates are known to attenuate incident radio-frequency radiation and, therefore, can be used to protect sensitive components against electromagnetic interference effects. At microwave frequencies, the case of interest is that of a uniform plane wave normally incident upon a "thin" electroconductive (EC) layer backed by a "thick" dielectric slab as illustrated in Fig. 1. The two relevant papers that have appeared in this TRANSACTIONS [1], [2] do not constitute a satisfactory treatment of the shielding effectiveness of such configurations. Liao's formula [1], which rests upon a procedure developed by Lassiter [3] for investigating the near-field situation, is basically incorrect and holds only under very special conditions. The work of Hansen and Pawlewicz [2], on the other hand, applies only to free-standing thin metallic sheets. My purpose here is to present a comprehensive but simple treatment of the microwave attenuation induced by an EC-coated plane-parallel dielectric and, in particular, to provide useful solutions for assessing the shielding effectiveness in an engineering-type environment.

Manuscript received August 14, 1989; revised October 23, 1989.
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IEEE Log Number 8933004.

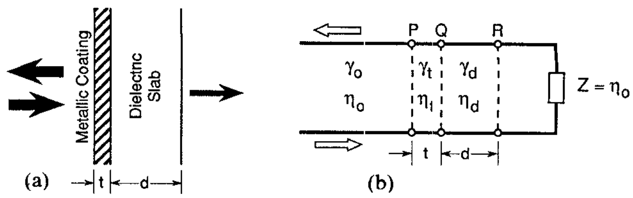

Fig. 1. Electroconductively (EC) coated dielectric slab (a) Electromagnetic shielding results from reflections at impedance discontinuities and absorption in the metal layer. (b) Equivalent transmission-line model including characteristic constants.

The shielding effectiveness (SE) is best defined in terms of the reduction in field intensity [SE (in dB) $= -20 \times \log(E_t/E_i)$] resulting from reflections and losses that occur upon inserting the "barrier" [4]. In the context of conventional transmission-line theory as formulated by Schelkunoff [5], which I will use to describe the propagation of a plane electromagnetic wave through the multilayer structure sketched in Fig. 1(a), the ratio of transmitted to incident electric fields corresponds to the voltage transmission coefficient T_V; the shielding effectiveness (in decibels) is therefore given by

$$SE = 10 \times \log[1/(T_V T_V^*)]. \tag{1}$$

The transmission coefficient T_V can be obtained on the basis of postulating that the metallic layer and the dielectric slab are both equivalent to sections of a transmission line as modeled in Fig. 1(b), that is, inserted into a transmission line of characteristic impedance η_0 terminating in a load impedance $Z = \eta_0$. The discontinuities at points P, Q, and R thus delineate two transmission-line sections of length t and d, each with its own set of characteristic constants. At this point, it is recalled that, in an isotropic medium of permeability μ and permittivity ϵ, the propagation constant of an electromagnetic wave of circular frequency ω is

$$\gamma = \sqrt{i\omega\mu(\sigma + i\omega\epsilon)} \tag{2}$$

where i stands for $\sqrt{-1}$ and σ designates the electrical conductivity. The intrinsic impedance of that medium is

$$\eta = \sqrt{\frac{i\omega\mu}{\sigma + i\omega\epsilon}} \tag{3}$$

which yields $\eta_0 = \sqrt{\mu_0/\epsilon_0} = 377\ \Omega$ for free space.

II. Metallic Sheet

To start with, let us consider the special case of an infinite metallic sheet of thickness t and apply transmission-line theory to describe the propagation of a normally incident plane electromagnetic wave. Following Schelkunoff [5], the voltage transmission coefficient can be expressed in a manner such as

$$T_V = \frac{p\exp(-\gamma_t t)}{1-q\exp(-2\gamma_t t)} \tag{4}$$

where p and q are transmission and reflection factors that combine the effect of the two surfaces, whereas γ_t refers to the propagation constant of the metallic sheet. The two factors p and q,

$$p = \frac{4\eta_0\eta_t}{(\eta_0+\eta_t)^2} \tag{5}$$

$$q = \frac{(\eta_0-\eta_t)^2}{(\eta_0+\eta_t)^2} \tag{6}$$

involve the intrinsic impedance of the conducting medium, η_t, in addition to the impedance of free space, η_0. At microwave frequencies, we have $\sigma \gg \omega\epsilon$ for metals and semiconductors, and the propagation constant reduces to

$$\gamma_t = (1+i)\sqrt{\omega\mu\sigma/2} = (1+i)/\delta \tag{7}$$

if one injects the concept of a skin depth. Similarly, the impedance η_t reduces to

$$\eta_t = (1+i)\sqrt{\omega\mu/(2\sigma)} = (1+i)/(\delta\sigma). \tag{8}$$

Assume now that the metallic layer always satisfies the two following conditions: (a) it is electrically thin, which means much thinner than the skin depth ($t/\delta \ll 1$), and (b) it has a low intrinsic impedance compared with the impedance of free space ($|\eta_t|/\eta_0 \ll 1$). Since the expressions p, $\gamma_t t$, and $1-q\exp(-2\gamma_t t)$ then yield very small numbers, the voltage transmission coefficient can be approximated by writing

$$T_V \simeq \frac{1}{(1-q)/p+2q\gamma_t t/p} \tag{9a}$$

which immediately leads to

$$T_V \simeq \frac{1}{1+\eta_0/(2R_s)} \tag{9b}$$

if R_s represents the sheet resistance [$R_s \equiv 1/(\sigma t)$]. The attenuation thus depends only on the ratio of the sheet resistance R_s to the characteristic impedance η_0, exactly as derived in [2]. In this connection, we may note that the voltage transmission coefficient (9b) is also that of a transmission line of characteristic impedance η_0 and terminal impedance $Z = \eta_0 R_s/(\eta_0 + R_s)$, which demonstrates that a thin metallic layer behaves essentially in the manner of a shunting resistance equal to the sheet resistance. In this light, the voltage reflection coefficient is given by

$$R_V \simeq \frac{-1}{1+2R_s/\eta_0} \tag{10}$$

which implies that, for very small sheet resistances ($R_s \ll \eta_0$), reflection is the dominant effect.

Returning now to (1) and substituting from (9b), we conclude that, to a very high degree of accuracy, the shielding effectiveness of a thin metallic sheet is given by

$$SE = 20\times\log[1+\eta_0/(2R_s)] \tag{11}$$

which confirms that the microwave attenuation in the far field is independent of frequency and that shielding can be enhanced by lowering the sheet resistance. In this regard, it should be emphasized that assumptions (a) and (b) relative to layer thickness and layer impedance imply that the sheet resistance must verify the conditions

$$\mu\omega t/2 \ll R_s \ll \eta_0^2/(\mu\omega t) \tag{12}$$

if (11) is called upon to provide a measure of the shielding.

III. Dielectric Slab

In the absence of any coatings, (4), (5), and (6) apply to a dielectric slab of thickness d with no other modification but substituting the parameters γ_d, η_d, and d for γ_t, η_t, and t, respectively. For nondissipative dielectrics, that is, upon setting $\mu = \mu_0$ and $\sigma = 0$, the intrinsic impedance reduces to

$$\eta_d = \sqrt{\mu_0/\epsilon} = \eta_0/\sqrt{\epsilon_r} \tag{13}$$

if ϵ_r represents the relative permittivity or dielectric constant. The propagation constant, in turn, reduces to

$$\gamma_t = \sqrt{-\omega^2\mu_0\epsilon} = i2\pi/\lambda \tag{14}$$

if λ refers to the wavelength in the solid medium:

$$\lambda = c/(\sqrt{\epsilon_r} f). \tag{15}$$

The voltage transmission coefficient of a dielectric slab may thus be expressed as follows:

$$T_V = \frac{4\sqrt{\epsilon_r}\exp(-i\beta)}{(1+\sqrt{\epsilon_r})^2-(1-\sqrt{\epsilon_r})^2\exp(-2i\beta)} \tag{16}$$

where β represents the "phase thickness," $\beta = 2\pi d/\lambda$, and measures the thickness of the slab in radians. This leads immediately to the correct formula for the transmittance of a perfect "window" at normal incidence,

$$T_V T_V^* = \frac{16\epsilon_r}{(1+\sqrt{\epsilon_r})^4+(1-\sqrt{\epsilon_r})^4-2(1+\sqrt{\epsilon_r})^2(1-\sqrt{\epsilon_r})^2\cos(2\beta)} \tag{17}$$

and describes the case as a function of the dielectric constant, the window thickness, as well as the wave frequency. This equation reveals that maxima and minima of $T_V T_V^*$ occur for phase thicknesses that are integral multiples of $\pi/2$. Specifically, for thicknesses d equal to integral multiples of $\lambda/2$ (half-wave configurations), there is no attenuation since

$$(T_V T_V^*)_{\max} = 1 \quad \text{when} \quad \beta = (N+1)\pi. \tag{18a}$$

In other words, the wave is "unaware" of the existence of the dielectric. For thicknesses $d = (2N+1)\lambda/4$, i.e., quarter-wave configurations, the incident wave experiences peak reflection losses; the transmittance is

$$(T_V T_V^*)_{\min} = [2\sqrt{\epsilon_r}/(1+\epsilon_r)]^2 \quad \text{when} \quad \beta = (N+1/2)\pi \tag{18b}$$

and decreases for dielectric constants $\epsilon_r > 1$.

IV. Coated Dielectrics

In principle, we are now equipped to handle the case of an EC-coated dielectric slab without further ado. For a multiple-interface system as modeled in Fig. 1(b), Schelkunoff's procedure (see [5, p. 226]) yields

$$T_V = \frac{p\exp(-\gamma_t t)\exp(-\gamma_d d)}{[1-q_t\exp(-2\gamma_t t)][1-q_d\exp(-2\gamma_d d)]} \quad (19)$$

for the voltage transmission coefficient; the two gammas are propagation constants as defined earlier, but the factors p and q must be redefined to accommodate three impedance discontinuities. Specifically, in our notations these factors are

$$p = \frac{8\eta_0\eta_t\eta_d}{(\eta_0+\eta_t)(\eta_t+\eta_d)(\eta_d+\eta_0)} \quad (20)$$

for the transmission and

$$q_t = \frac{(\eta_t-\eta_0)(\eta_t-Z_d)}{(\eta_t+\eta_0)(\eta_t+Z_d)} \quad (21a)$$

$$q_d = \frac{(\eta_d-\eta_t)(\eta_d-\eta_0)}{(\eta_d+\eta_t)(\eta_d+\eta_0)} \quad (21b)$$

for the reflections associated with the metallic section and the dielectric section, respectively. Note that q_t includes the input impedance of the dielectric slab, Z_d, which is the impedance experienced at station Q when looking to the right. Since the output impedance is η_0 and the "line" is nondissipative, it is immediately seen that the relation

$$Z_d = \eta_d\frac{\eta_0+i\eta_d\tan(\beta)}{\eta_d+i\eta_0\tan(\beta)} \quad (22)$$

holds if β measures the phase thickness as specified in Section III. For $\beta=(N+1)\pi$, where $N=0,1,2,\cdots$, the input impedance equals the output impedance ($Z_d=\eta_0$), whereas for $\beta=(N+1/2)\pi$ we have $Z_d=\eta_d^2/\eta_0$; since both η_0 and η_d are pure resistances, these are the largest and smallest input impedances to be considered.

Assume now that the coating is electrically thin ($|\gamma_t|\times t \ll 1$) and made of low-impedance material ($|\eta_t|/\eta_0 \ll 1$). In that case, we may proceed as in Section II, characterize the coating by its sheet resistance, and write

$$T_V \simeq \frac{4\exp(-i\beta)}{(1/\eta_0+1/R_s+1/Z_d)[(\eta_0+\eta_d)+(\eta_0-\eta_d)\exp(-2i\beta)]}. \quad (23)$$

Upon substituting $\eta_0/\sqrt{\epsilon_r}$ for η_d, the input impedance Z_d can be expressed as

$$Z_d = \eta_0\frac{1+i\tan(\beta)/\sqrt{\epsilon_r}}{1+i\sqrt{\epsilon_r}\tan(\beta)} \quad (24)$$

which leads to the key result of this paper:

$$\frac{1}{T_V T_V^*} \simeq \frac{(1+\epsilon_r)+(\epsilon_r-1)\cos(2\beta)}{8\epsilon_r}\left|1+\frac{\eta_0}{R_s}+\frac{1+i\sqrt{\epsilon_r}\tan(\beta)}{1+i\tan(\beta)/\sqrt{\epsilon_r}}\right|^2 \quad (25)$$

and, hence, directly to the shielding effectiveness (see (1)). The shielding performance of an EC-coated dielectric slab thus depends on three parameters: R_s, ϵ_r, and β. First, it is immediately apparent that the performance can always be improved by lowering the sheet resistance. The dependence on dielectric constant and phase thickness, however, is seen to be quite complex considering that β involves the thickness d and the wavelength λ, which in turn is a function of the permittivity and the frequency. Fortunately, given an R_s value, variations in transmittance are confined to a relatively narrow range delimited by quarter-wave and half-wave thick substrates. This is illustrated in Fig. 2, which refers to an EC-coated slab of fused silica possessing a dielectric constant of 3.78 as in Liao's paper [1]. The thicknesses $d=(N+1)\lambda/2$ and $d=(2N+1)\lambda/4$ represent limiting conditions in the sense that the attenuation always lies on or between the two relevant curves, half-wave thicknesses providing maximum shielding for $R_s \leq 400\ \Omega/\square$. On the same figure, I am also displaying Liao's solution [1], thus demonstrating that his formula may hold for quarter-wave configurations (minimum shielding) but only if the sheet resistance does not exceed 10 $\Omega/\square$.

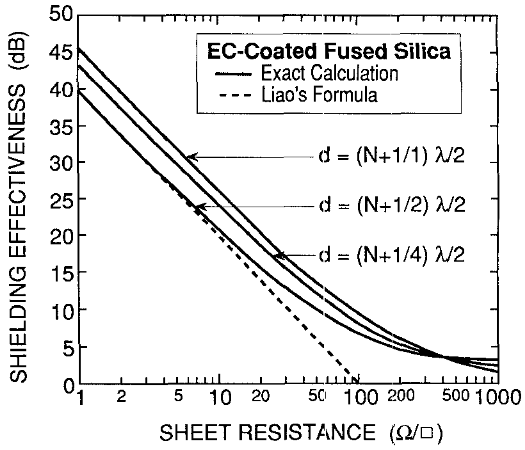

Fig. 2. Microwave shielding effectiveness of EC-coated fused silica as a function of the coating's sheet resistance The solid curves are correct results derived from the general solution (25). The broken line illustrates Liao's formula [1].

Actually, for half-wave configurations, the solution (25) reduces to a much simpler expression and yields

$$(SE)_{\lambda/2} = 20\times\log\left[1+\frac{\eta_0}{2R_s}\right] \quad (26a)$$

which is precisely the formula derived in Section II for free-standing metallic sheets and confirms that, under resonant conditions, the presence of the substrate has no effect on shielding. For quarter-wave configurations, on the contrary, we have

$$(SE)_{\lambda/4} = 20\times\log\left[\frac{1+\epsilon_r}{2\sqrt{\epsilon_r}}+\frac{\eta_0}{2\sqrt{\epsilon_r}R_s}\right] \quad (26b)$$

which depends on the dielectric constant in the manner discussed in [6].

V. Practical Example

Fig. 3 displays experimental shielding-effectiveness data as a function of sheet resistance ($9 \leqslant R_s \leqslant 125\ \Omega/\square$) for single-ply conductive glass specimens a quarter of an inch thick; the data are as recorded in table I of Lassiter's paper [3], for frequencies of 700 and 990 MHz, and are believed representative of plane-wave attenuations. On inserting a dielectric constant value of 4 ($\epsilon_r=4$), which is typical of glass at microwave frequencies, the formulas (26a) and (26b) generate theoretical upper and lower SE limits as drawn in Fig. 3. Indeed, these two curves bracket the results of most of the measurements, the lower frequency points

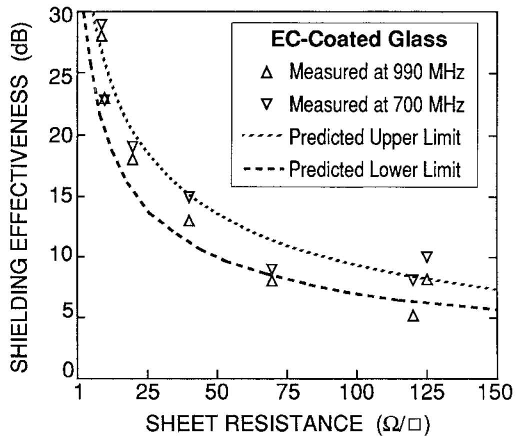

Fig. 3. Microwave shielding effectiveness of single-ply conductive glass as a function of the nominal sheet resistance; the data points are as listed in [3].

lying somewhat closer to the upper, no-substrate limit, thus validating the model and supporting some of the conclusions.

VI. Conclusion

Liao's formula [1] for the shielding effectiveness of metal-coated glass only applies to sheet resistances $R_s \lesssim 10\ \Omega/\square$ and does not take into account enhancements that may occur under "resonant" conditions. Metal-sheet reflections as evaluated in [2] dominate the loss mechanism for low-impedance films, but microwave absorption by the metal layer becomes increasingly significant for higher sheet resistance coatings. The analysis that was carried out in this paper assumes normally incident plane waves and should be applicable to any EC-coated optically transparent dielectric provided the thickness of the coating is much smaller than the skin depth. The two formulas presented in (26a) and (26b), which derive from the general expression (25), allow us to quickly assess the far-field microwave shielding performance in the sense that they yield upper and lower limits for the attenuation. For sheet resistances $R_s \lesssim \eta_0/(\sqrt{\epsilon_r}-1)$, either half-wave-thick or low-dielectric-constant substrates provide optimum shielding; for larger sheet resistances, on the contrary, quarter-wave thicknesses in conjunction with high-permittivity dielectrics result in enhanced attenuation. Substantial attenuation, however, cannot be achieved with high-resistivity coatings, irrespective of the coating's nature or the coating's design.

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Flexible Circular Waveguides at Millimeter Wavelengths from Metallized Teflon Tubing

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***Abstract* — Flexible waveguides for use at millimeter wavelengths have been fabricated by deposition of a metallic film onto the composite-modified inside surface of Teflon tubing. The attenuation characteristics in the range 80 to 115 GHz show losses of the order of 0.1 dB/cm. Bending, twisting, and rotating to the limit of plastic mechanical stability (curvature radius typically > 8 cm) have negligible effect on the attenuation, and bend angles ⩽ 45° produce relatively small changes in the insertion phase.**

I. Introduction

Considerable effort has been devoted to the design and characterization of dielectric waveguides for microwave frequencies, which are analogous to optical fibers. Dielectric rods of Teflon and polystyrene have been shown to operate at 71 and 74 GHz, successfully supporting propagation of hybrid modes [1]. Flexible millimeter waveguides consisting of Teflon core and Teflon cladding, as well as polyethylene multilayer waveguides, have also been demonstrated [2]. Polymeric dielectric waveguides are inexpensive, flexible, and low in weight, and are therefore attractive for a variety of practical applications. Unfortunately, they are subject to bending losses since the rod and cladding have similar dielectric constants. Recently, flexible Teflon tubing filled with a high-dielectric-constant powder of inorganic titanate salts has been fabricated, having attenuation low enough to be attractive for short-distance transmission at 10 and 94 GHz [3]. However, there are certain problems associated with dimensional imperfections and packing density irregularities, which result in scattering, reflections, and creation of multiple hybrid modes. Coupling to the TE_{11} circular metallic waveguide is also an issue.

In this paper we have focused on an alternative, which is a flexible composite consisting of thin-wall metal tubing encircled by a protective polymeric coating. Recent developments in polymer science enable materials in layered form to be tailored in both composition and size in order to meet specific requirements [4]. Upon doping, conjugated polymers can be made to exhibit semiconducting, metallic, or even superconducting properties not traditionally associated with these materials. This has resulted in a wide range of new basic research and a rich potential for applications in microelectronics and optoelectronics. We have created surface compatibility between the metal and the elastic coating by forming a film of conducting polymer on the Teflon surface. Such a modification makes Teflon accessible for metal deposition or electroplating, with good adherence at the interface. In this paper we present a brief summary of our fabrication and measurement techniques and of results obtained in the 80–115 GHz band.

Manuscript received August 23, 1989; revised October 18, 1989. This work was supported by the National Science Foundation under Grant AST 88-15406.

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IEEE Log Number 8933005.

0018-9480/90/0300-0324$01.00 ©1990 IEEE