which, as Fig. 3 shows, at each of the $d/2a$ values used in the measurements is indistinguishably different from the measured mean. The empirical equation is a weighted least-squares fit to the data, in which the weights were chosen in inverse proportion to the lengths of the error bars in Fig. 3. A formula of this kind was chosen as, in addition to giving a good fit to the data, it has the correct form of functional behavior for both large and small arguments. Values in these regions, of course, lie outside the range of the measurements and represent an extrapolation in which there must necessarily be less confidence.

V. Radiation Conductance

At high frequencies, the transmission line will radiate from its open end. The excess charge on each wire in the vicinity of the open circuit may be thought of as generating a dipole moment, which can then be used to estimate an equivalent conductance G_f which should appear in parallel with C_f in Fig. 2 to give a complete equivalent circuit.

If there is a voltage V across the open circuit, then the equivalent dipole moment will be

$$p = C_f V d. \tag{6}$$

An expression for the power radiated by a dipole of moment p can be found in a number of standard texts (e.g., [5]) and by setting this equal to $V^2 G_f$, an expression for the radiation conductance is readily determined in terms of the fringing capacitance and the geometric constants of the line. Drawing together the previous results, it is thus easily shown that the normalized radiation conductance is

$$Z_0 G_f = \frac{(kd)^4 \left(\frac{\delta l}{d}\right)^2}{12\left(\frac{\pi Z_0}{\eta}\right)} \tag{7}$$

where k is the wavenumber. Since kd must be small in any practical transmission line, it is clear that the radiation conductance is always small and the end effect primarily capacitive.

VI. Conclusion

By means of experiments in an electrolytic tank, we have obtained data which allows determination of the effective position of the open circuit at the end of an open-circuited solid-conductor two-wire line. The data cover nearly all practical characteristic impedances, and an empirical formula is provided which allows easy input of the data to a computer. It has been shown that this is also sufficient information from which to estimate the radiation conductance associated with the fringing capacity in the equivalent circuit of the discontinuity. However, this conductance is so small in any practical transmission line that it is always safe to assume that the discontinuity effect is capacitive. The data presented are extendable to the case of a rod lying parallel to an infinite ground plane.

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An Estimate of the Interaction Impedance of a Vane-Loaded Helix Using Equivalent-Circuit Analysis

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Abstract —**The expression for the interaction impedance of a broad-band vane-loaded helical slow-wave structure has been found using Pierce's simple theory for a helix in free space, and the results of the equivalent-circuit analysis of the loaded structure. The interaction impedance has been found to be of the order of half the characteristic impedance in a typical loaded structure. The dependence of the small-signal gain on the thickness of the helix wire, the size of the vanes, and the location of the metal envelope has been predicted.**

I. Introduction

Recently, we presented in this Transactions an equivalent-circuit analysis of the vane-loaded helical slow-wave structure and, hence, predicted its optimum vane dimension leading to a 'dispersion-free' behavior [1]. The vane-loaded structure was estimated to have a higher value of characteristic impedance than an indentical vaneless structure. A more realistic parameter than the characteristic impedance, not considered in this analysis, is the interaction impedance of the structure. This is done here by suitably interpreting the expression for the interaction impedance of a helix in free space [2] with the help of the results of the equivalent-circuit analysis of the vane-loaded helical slow-wave structure [1].

II. Theoretical Formula

The analytical results for a vane-loaded helix may be so interpreted as to consider such a structure as embedded in a medium extended in dimensions to infinity and supposedly of relative permeability $\alpha_L = L/L_0$, and relative permittivity $\alpha_C = C/C_0$, where L and C are, respectively, the inductance per unit length and the capacitance per unit length of the vane-loaded structure [1], and L_0 and C_0 are the corresponding quantities for an identical helix in free space [1]. The impedance parameter F' of such a structure then may be written by replacing μ_0 and ϵ_0

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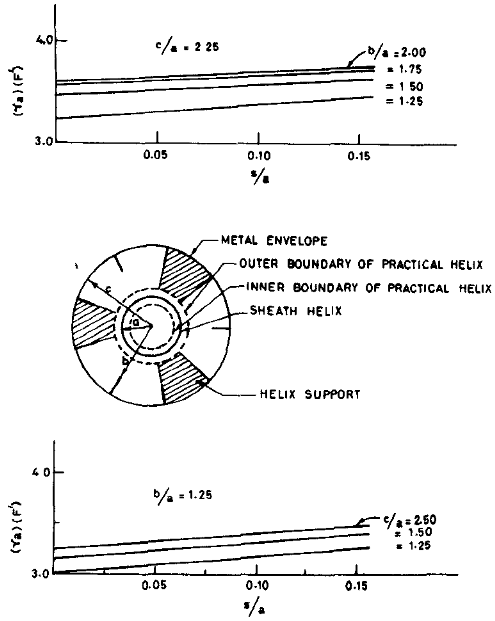

Fig. 1. $(\gamma a)(F')$, a quantity proportional to gain, as a function of the dimensions of the vane-loaded helical structure.

by $\alpha_L \mu_0$ and $\mu_C \epsilon_0$, respectively, in the expression for the impedance parameter F of an identical helix in free space [2]. This would lead to

$$F' = F(\alpha_L/\alpha_C)^{1/6}. \tag{1}$$

The interaction impedance, which is related to the impedance parameter [2], then may be expressed as

$$K' = K(\alpha_L/\alpha_C)^{1/2} \tag{2}$$

where K' represents the interaction impedance of the vane-loaded helix and K represents the corresponding quantity for an identical helix in free space. K may be suitably expressed using the dispersion relation as [2]

$$K = \frac{1}{2}\left[\frac{I_1(\gamma a)K_1(\gamma a)}{I_0(\gamma a)K_0(\gamma a)}\right]^{1/2} F^3 \cot\psi \tag{3}$$

where a is the mean helix radius, ψ is the helix pitch angle, I_n and K_n $(n = 0,1)$ are the modified Bessel functions of the first and the second kinds, respectively, and γ is the radial propagation constant of the structure.

III. Result and Discussion

The various structure dimensions shown in Fig. 1 and Table I are the mean helix radius a, being equal to the sheath helix radius, the radial coordinate of the tips of the vanes b, the radius of the metal envelope c, and the seperation s between the sheath helix and the dielectric supports, being equal to the helix-wire radius.

In Table I, the typical optimized vane-loaded $(b \neq c)$ and vaneless $(b = c)$ structures are compared with respect to the characteristic impedance $Z(=(L/C)^{1/2})$ [1] and the interaction impedance K. It is found (Table I) that the optimized vane-loaded structure is superior to the optimized vaneless structure with respect to having a higher value of the interaction impedance and, hence, the gain and efficiency of the device [2]. Table I also gives the characteristic impedances of these structures and the corresponding results for an identical helix in free space for the sake of comparison. In Fig. 1, $(\gamma a)(F')$, a quantity proportional to the gain of the device, is plotted against the various structural dimensions, showing qualitatively how the gain of the device would increase with s, b, and c.

TABLE I
The Characteristic and Interaction Impedances of the Optimized Vane-Loaded Structure Compared With Those of the Optimized Vaneless Structure

$\gamma a = 1.6$; $\cot\psi = 8$			
	Optimised vane-loaded structure: c/a=2.50, s=0; (b/a)opt=1.70	Optimised vaneless structure: b=c, s=0; (c/a)opt=1.25	Helix-in-free-space
$Z_{(Ohm)}$	105.6	62.4	140.6
$K_{(Ohm)}$	53.6	30.9	53.2

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Conversion Losses in GaAs Schottky-Barrier Diodes

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Abstract —**The conversion losses of a Schottky-barrier diode have been calculated for a set of realistic diode parameters. It is found that previous work overestimated the substrate losses by 30 percent. It is also shown that a lightly doped epitaxial layer will decrease the barrier capacitance and with properly designed thickness will avoid any resistance losses due to this layer. Parasitic losses can thus be reduced substantially.**

I. Introduction

The model of a Schottky-barrier diode we contemplate in the present work consists of a cylindrical contact of radius a on a thin cylindrical wafer of n-type semiconductor material possessing a radius $b \gg a$. The semiconductor wafer in turn consists of a layer of undoped or at least lightly doped GaAs epitaxially grown on a heavily doped n-type GaAs substrate several mils thick. Under high-frequency operation, the equivalent circuit of the structure just described consists of a spreading impedance Z_s in series with a combination of the barrier resistance R_b and the barrier capacitance C in parallel [1]. The current–voltage characteristic for a small ac signal may then be represented by

$$I = \frac{1 + i\omega C R_b}{R_b + Z_s + i\omega C R_b Z_s} V. \tag{1}$$

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