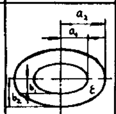
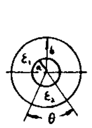
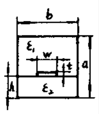
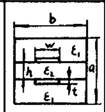
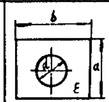
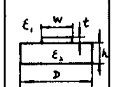
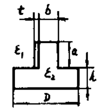
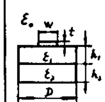
TABLE I
SAMPLE COMPUTATIONS

	elliptic coaxial line	Partially-filled coaxial line	microstrip in a box	dielectric-Supported-strip in a box	rectangular slab-line	microstrip	vertical microstrip	double dielectric microstrip
Structure	 $a_1 = 1\ 25$ $a_2 = 2\ 00$ $b_1 = 0\ 75$ $b_2 = 1\ 732$ $\varepsilon = \varepsilon_0$	 $a = 3\ 50$ $b = 8\ 00$ $\theta = 2\pi/10$ $\varepsilon_1 = \varepsilon_0$ $\varepsilon_2 = 3\ 0\varepsilon_0$	 $a = 2\ 02$ $b = 7.0$ $h = 1\ 00$ $W = 1\ 00$ $t = 0\ 01$ $\varepsilon_1 = \varepsilon_0$ $\varepsilon_2 = 9\ 6\varepsilon_0$	 $a = 5\ 00$ $b = 5\ 00$ $h = 1\ 00$ $W = 2\ 00$ $t = 0\ 001$ $\varepsilon_1 = \varepsilon_0$ $\varepsilon_2 = 2\ 35\varepsilon_0$	 $a = 1\ 0$ $b = 1\ 25$ $d = 0\ 51$ $\varepsilon = \varepsilon_0$	 $W = 1.00$ $t = 0\ 002$ $h = 1\ 00$ $\varepsilon_1 = \varepsilon_0$ $\varepsilon_2 = 9\ 6\varepsilon_0$ $D = 12\ 0$	 $a = 6\ 50$ $b = 2\ 00$ $h = 1.00$ $t = 0\ 001$ $\varepsilon_1 = \varepsilon_0$ $\varepsilon_2 = 2\ 55\varepsilon_0$ $D = 12\ 00$	 $W = 1\ 0$ $t = 0\ 02$ $h_1 = 0\ 60$ $h_2 = 0\ 40$ $\varepsilon_1 = 9\ 6\varepsilon_0$ $\varepsilon_2 = 2\ 65\varepsilon_0$ $D = 12\ 0$
$Z_0\ (\Omega)$ this method	37.74	45 68	46 04	65.02	50.43	51.62	198.30	63.06
$Z_0\ (\Omega)$ other method	37 43 [3, Table 2-3]	45 24 [4, Eq 5 3 1]	49 80* [4, Fig 3.13]	62 50 [5, Fig 7]	49 99 [4, Table 4 1]	49 79 [4, Table 3 4]		

*There are errors in [4, fig. 3.13]. These data are unreliable.

IV. EXAMPLES AND DISCUSSION

A general computer program has been written using the above formulation. Input consists of the structure parameters and the coordinates of the end points of each segment. Output consists of the charge distribution and the characteristic impedance. Some examples of computed results are given in Table I. Also given for comparison are some results obtained by other methods. Our solution is in good agreement with those obtained by other methods in most cases.

Although the computer program is written explicitly for the case of no infinite ground plane, it can be sued to approximate an infinite ground plane by making one conductor a wide, but finite, plane. The fact that parts of the ground plane far from the other conductor are missing should cause negligible error in the result if the width is taken large. Some examples for the case of a ground plane were computed using this approximation and are included in Table I.

The formulation of this paper can be extended to multiconductor transmission lines. For an N-conductor transmission line, there are $N-1$ quasi-TEM modes [6]. Each corresponds to the case for which all conductors but one are grounded. The ungrounded conductor is set at unit potential. The charge on each conductor then will be equal to an element of the capacitance matrix for the line [1]. The inductance matrix for the line is $\epsilon_0 \mu_0$ times the inverse of the capacitance matrix obtained by replacing all dielectrics by free space [1].

An alternative method for solving the matrix equation (17) for $[\alpha]$ is given in [7]. In that method, the constant k is eliminated and not determined.

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Correction to "Theoretical Considerations on the Use of Circularly Symmetric TE Modes for Digital Ferrite Phase Shifters"

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Recently, we have become aware of increased publication activity by authors who refer to the above early paper.[1] We felt that it is particularly timely, therefore, to inform those concerned that in the above paper a few formulas, unfortunately, are in error. Therefore, we would like to bring attention to the correct version of the formulas. Equation (6), on p. 422, should appear as

$$\gamma_0^2 = \omega\mu_0\left[(1+\chi)^2 - \kappa^2\right] = \omega\mu_0\Delta \tag{6}$$

while a_1 and d_2 in (9) and (10) should be

$$a_1 = -a_0(\alpha/3) \tag{9}$$

$$d_2 = -1/4. \tag{10}$$

Equations (18) and (20), on p. 424, should read

$$\frac{B_1(\alpha;\tau_1 x)}{H_1(\alpha;\tau_1 x)} \cdot \frac{F_1(\alpha;\tau_1 x) - F_2(\alpha;\tau_1 x)}{F_3(\alpha;\tau_1 x) - F_2(\alpha;\tau_1 x)} = \frac{B_1(\alpha;\tau_2 x)}{H_1(\alpha;\tau_2 x)} \cdot \frac{F_1(\alpha;\tau_2 x) - F_4(\alpha;\tau_2 x)}{F_3(\alpha;\tau_2 x) - F_4(\alpha;\tau_2 x)} \tag{18}$$

where

$$F_2(\alpha;\tau x) = \tau y\, J_0(\tau y)/J_1(\tau y)$$

and

$$-j\omega\mu_0 b H_z^I = y J_0(yr/b)\, J_1(\tau y). \tag{20}$$

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[1] D. M. Bolle and G S. Heller, *IEEE Trans. Microwave Theory Tech*, vol. MTT-13, pp. 421–426, July 1965.

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