solving the capacitance problem. Numerical results have been obtained and compared with previously published results. The accuracy and the relative efficiency of the method have been demonstrated.

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Theory of Coupled Open Transmission Lines and Its Applications

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***Abstract*—A technique is presented which is applicable to any uniform coupled open transmission lines such as coupled optical integrated circuits. The proposed technique is as follows.**

The electromagnetic fields of the wave propagating along a coupled line is expressed in terms of a linear combination of the fields associated with the individual lines, as a zero-order approximation. Inserting this trial field description into the variational expression for the propagation constant β, and applying the well-known Rayleigh–Ritz's procedure, accurate solutions for the propagation constants of the coupled lines are obtained.

This method can be applied generally to analyze coupled structures in microwave, millimeter wave, and optical wave circuitry. As an illustrative example, the coupling between two optical transmission lines consisting of lens-like dielectric media has been analyzed by means of the proposed technique.

I. INTRODUCTION

THE PROBLEM of coupling between open transmission lines is interesting both from academic and practical points of view in connection with the design and analysis of optical integrated circuits and components (see Fig. 6).

Though several papers concerning the coupling of open transmission lines have been reported [1]–[4], only the special case where the coupling occurs between two identical open transmission lines has been analyzed. To the authors knowledge, a technique adequate to treat the coupling between two different open transmission lines has not been given before.

This paper presents a theory which can be applied to two arbitrary coupled open transmission lines. The technique proposed is based on the variational method.[1] The procedure of the calculation is quite simple and straightforward as long as the electromagnetic fields associated with individual transmission lines are already known. As an example of the application of the proposed theory, the coupling between two dielectric lines consisting of lens-like media has been analyzed.

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II. VARIATIONAL EXPRESSION FOR THE PROPAGATION CONSTANT

The magnetic field $\boldsymbol{H}$ of any uniform open transmission line can be expressed as

$$\boldsymbol{H} = (\boldsymbol{h}_t + \boldsymbol{i}_z h_z) \exp[j(\omega t - \beta z)] \quad (1)$$

where $\boldsymbol{h}_t$ and h_z are the transverse and longitudinal components of the field, respectively, $\boldsymbol{i}_z$ is a unit vector in the longitudinal direction z, and β is the propagation constant. The variational expression for the propagation constant β in the z direction of a lossless uniform transmission line is given as[2]

$$\beta^2 = N/D \quad (2)$$

[1] The theory described in the present paper was originally reported at Radiation Science Research Committee on April 30, 1971, in Japanese. After preparing the manuscript of the present paper, the authors found two related articles. One is Marcuse's work [5] where the coupling problem is treated with a perturbation method, and the other is Snyder's paper [6] in which the problem is solved by a modal-expansion approach.

[2] The variational expressions for the propagation constant of the guided waves have been derived by several authors in different forms. The expression (2) is a slight modification and generalization of Kurokawa's original one [7]. The variational expression (2) has the advantage that it can be easily applied even though the material constants involved change discontinuously in the transverse cross-sectional surface of the guide.

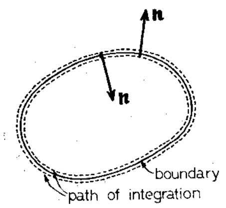

Fig. 1. Contours along which the line integral is evaluated.

$$N = \int \left[\epsilon \left(\omega^2 \mu \boldsymbol{h}_t - \nabla \times \frac{1}{\epsilon} \nabla \times \boldsymbol{h}_t \right)^2 - \omega^2 \frac{1}{\mu} (\nabla \cdot \mu \boldsymbol{h}_t)^2 \right] dS + 2 \int \left[\left(\frac{1}{\mu} \nabla \cdot \mu \boldsymbol{h}_t \right) \left\{ \boldsymbol{n} \cdot \left(\omega^2 \mu \boldsymbol{h}_t - \nabla \times \frac{1}{\epsilon} \nabla \times \boldsymbol{h}_t \right) \right\} \right] dl \tag{3}$$

$$D = \int \left[\omega^2 \mu \boldsymbol{h}_t^2 - \frac{1}{\epsilon} (\nabla \times \boldsymbol{h}_t)^2 \right] dS + 2 \int \left[\left(\frac{1}{\epsilon} \nabla \times \boldsymbol{h}_t \right) \cdot (\boldsymbol{n} \times \boldsymbol{h}_t) \right] dl. \tag{4}$$

The surface integral must be carried out over the whole transverse cross section, while the line integral must be evaluated along closed contours in Fig. 1 selected on opposite sides of the boundary across which the material constants change discontinuously. $\boldsymbol{n}$ is a unit vector normal to the boundary and directed as shown in Fig. 1.

It should be mentioned, in this connection, that we can assume any appropriate trial function $\boldsymbol{h}_t$ as long as both $(1/\mu) \nabla \cdot \mu \boldsymbol{h}_t$ and $\boldsymbol{i}_z \cdot (\boldsymbol{n} \times \boldsymbol{h}_t)$ are continuous across the boundary between the different media.

Since (2) is a variational expression, we can obtain a more precise value of the propagation constant β from (2), using the zero-order approximate expression for the field component $\boldsymbol{h}_t$.

III. COUPLING THEORY BASED ON THE VARIATIONAL METHOD

Let us consider the coupled system shown in Fig. 2 where the coupling occurs between line 1 and line 2. Assume that the electromagnetic field of the individual line is known. Let the transverse field components of line 1 and line 2 be $\boldsymbol{h}_{t1}$ and $\boldsymbol{h}_{t2}$, respectively, and the propagation constants be β_1 and β_2, respectively. As long as β_1 and β_2 are not markedly different, the transverse field component $\boldsymbol{h}_t$ of the coupled system is approximately represented in terms of the linear combination of $\boldsymbol{h}_{t1}$ and $\boldsymbol{h}_{t2}$ as

$$\boldsymbol{h}_t = m_1 \boldsymbol{h}_{t1} + m_2 \boldsymbol{h}_{t2}. \tag{5}$$

The coefficients m_1 and m_2 must be determined in such a manner that (2) has a stationary value when (5) is substituted into (2). Substituting (5) into (2), we get

$$\beta^2 = \sum_{i=1}^{2} \sum_{j=1}^{2} N_{ij} m_i m_j \Big/ \sum_{i=1}^{2} \sum_{j=1}^{2} D_{ij} m_i m_j \tag{6}$$

where

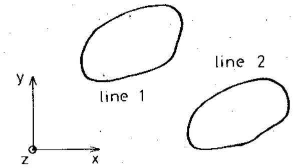

Fig. 2. Coupled system consisting of two arbitrary transmission lines.

$$N_{ij} = N_{ji} = \int \left[\epsilon \left(\omega^2 \mu \boldsymbol{h}_{ti} - \nabla \times \frac{1}{\epsilon} \nabla \times \boldsymbol{h}_{ti} \right) \cdot \left(\omega^2 \mu \boldsymbol{h}_{tj} - \nabla \times \frac{1}{\epsilon} \nabla \times \boldsymbol{h}_{tj} \right) - \omega^2 \frac{1}{\mu} (\nabla \cdot \mu \boldsymbol{h}_{ti})(\nabla \cdot \mu \boldsymbol{h}_{tj}) \right] dS + \int \left[\left(\frac{1}{\mu} \nabla \cdot \mu \boldsymbol{h}_{ti} \right) \cdot \left\{ \boldsymbol{n} \cdot \left(\omega^2 \mu \boldsymbol{h}_{tj} - \nabla \times \frac{1}{\epsilon} \nabla \times \boldsymbol{h}_{tj} \right) \right\} + \left(\frac{1}{\mu} \nabla \cdot \mu \boldsymbol{h}_{tj} \right) \cdot \left\{ \boldsymbol{n} \cdot \left(\omega^2 \mu \boldsymbol{h}_{ti} - \nabla + \frac{1}{\epsilon} \nabla \times \boldsymbol{h}_{ti} \right) \right\} \right] dl, \quad i = 1, 2, \quad j = 1, 2 \tag{7}$$

$$D_{ij} = D_{ji} = \int \left[\omega^2 \mu \boldsymbol{h}_{ti} \cdot \boldsymbol{h}_{tj} - \frac{1}{\epsilon} (\nabla \times \boldsymbol{h}_{ti}) \cdot (\nabla \times \boldsymbol{h}_{tj}) \right] dS + \int \left[\left(\frac{1}{\epsilon} \nabla \times \boldsymbol{h}_{ti} \right) \cdot (\boldsymbol{n} \times \boldsymbol{h}_{tj}) + \left(\frac{1}{\epsilon} \nabla \times \boldsymbol{h}_{tj} \right) \cdot (\boldsymbol{n} \times \boldsymbol{h}_{ti}) \right] dl \quad i = 1, 2, \quad j = 1, 2. \tag{8}$$

The condition for which (6) has a stationary value, that is, $\partial \beta^2 / \partial m_1 = 0$ and $\partial \beta^2 / \partial m_2 = 0$, is given by

$$\frac{m_1}{m_2} = - \frac{N_{12} - \beta^2 D_{12}}{N_{11} - \beta^2 D_{11}} = - \frac{N_{22} - \beta^2 D_{22}}{N_{12} - \beta^2 D_{12}}. \tag{9}$$

From the foregoing equations, β^2 can be expressed as

$$\beta^2 = \frac{1}{2(D_{11} D_{22} - D_{12}^2)} \left[(D_{11} N_{22} + D_{22} N_{11} - 2 D_{12} N_{12}) \pm \{ (D_{11} N_{22} - D_{22} N_{11})^2 + 4 (D_{11} N_{12} - D_{12} N_{11})(D_{22} N_{12} - D_{12} N_{22}) \}^{1/2} \right]. \tag{10}$$

It can be seen from (10) that, in general, the coupled system possesses two different propagation constants. They correspond to symmetric and antisymmetric modes of propagation.

Equation (10) can be approximated further to a simpler form with the help of the following relations that are satisfied with sufficient accuracy in most practical cases:

$$N_{11} = \beta_1^2 D_{11} \qquad N_{22} = \beta_2^2 D_{22}. \tag{11}$$

Then (9) and (10) can be reduced to the more convenient form

$$\beta = \beta_0 \pm c \tag{12}$$

$$\frac{m_1}{m_2} = \left(\frac{D_{22}}{D_{11}}\right)^{1/2} \left\{1 + \left(\frac{\delta}{F}\right)^2\right\}^{1/2} (\pm 1 + \Delta) \tag{13}$$

where

$$\beta_0 = \tfrac{1}{2}(\beta_1 + \beta_2) \tag{14}$$

$$\delta = \frac{\beta_1 - \beta_2}{\beta_1 + \beta_2} \tag{15}$$

$$c = \beta_0 F \left\{1 + \left(\frac{\delta}{F}\right)^2\right\}^{1/2} \tag{16}$$

$$\Delta = \frac{\beta_0 \delta}{c} \tag{17}$$

$$F = \tfrac{1}{2}(D_{11}D_{22})^{-1/2}\left(D_{12} - \frac{N_{12}}{\beta_0^2}\right). \tag{18}$$

We designate the mode with the upper sign in the above equations as mode a and the mode with the lower sign as mode b. Also, we denote the propagation constants of mode a and mode b as β_a and β_b, respectively, and abbreviate the ratios m_1/m_2 for mode a as K_a and for mode b as K_b. Then the transverse field components $\boldsymbol{H}_{ta}$ of mode a and $\boldsymbol{H}_{tb}$ of mode b can be represented as

$$\boldsymbol{H}_{ta} = A(K_a \boldsymbol{h}_{t1} + \boldsymbol{h}_{t2}) \exp[j(\omega t - \beta_a z)] \tag{19}$$

$$\boldsymbol{H}_{tb} = B(K_b \boldsymbol{h}_{t1} + \boldsymbol{h}_{t2}) \exp[j(\omega t - \beta_b z)] \tag{20}$$

where the constants A and B are determined by the initial or excitation conditions. From the above two equations, the transverse field components $\boldsymbol{H}_{t1}$ of line 1 and $\boldsymbol{H}_{t2}$ of line 2 can be obtained as

$$\boldsymbol{H}_{t1} = \boldsymbol{h}_{t1}[AK_a \exp(-j\beta_a z) + BK_b \exp(-j\beta_b z)] \cdot \exp(j\omega t) \tag{21}$$

$$\boldsymbol{H}_{t2} = \boldsymbol{h}_{t2}[A \exp(-j\beta_a z) + B \exp(-j\beta_b z)] \exp(j\omega t). \tag{22}$$

Let the normalized complex amplitudes of the waves traveling along line 1 and line 2 be $a_1(z)$ and $a_2(z)$, respectively. From (21) and (22), the complex amplitudes $a_1(z)$ and $a_2(z)$ can be expressed in terms of the complex amplitudes of the incoming waves $a_1(0)$ and $a_2(0)$ at the beginning of the coupling portion $z = 0$ as follows:

$$\begin{bmatrix} a_1(z) \\ a_2(z) \end{bmatrix} = \exp(-j\beta_0 z) \cdot \begin{bmatrix} \cos cz - j\Delta \sin cz & -j(1-\Delta^2)^{1/2} \sin cz \\ -j(1-\Delta^2)^{1/2} \sin cz & \cos cz + j\Delta \sin cz \end{bmatrix} \begin{bmatrix} a_1(0) \\ a_2(0) \end{bmatrix}. \tag{23}$$

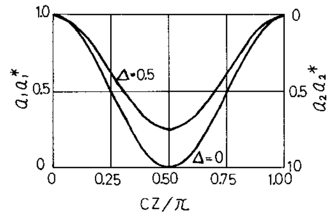

Fig. 3. Power transfer between line 1 and line 2 due to the coupling.

Equation (23) provides the necessary information about the coupling phenomenon between line 1 and line 2. For example, let us consider the typical case where $a_1(0) = 1$ and $a_2(0) = 0$. In this case, (23) becomes

$$\begin{aligned} a_1(z) &= (\cos cz - j\Delta \sin cz) \exp(-j\beta_0 z) \\ a_2(z) &= -j(1-\Delta^2)^{1/2} \sin cz \exp(-j\beta_0 z) \end{aligned} \tag{24}$$

from which we can evaluate the power exchange between two coupled lines. The results are shown in Fig. 3. As we can see from Fig. 3, the power fed into line 1 at $z = 0$ is transferred completely to line 2 at the particular coupling length

$$z = \frac{\pi}{2c} \tag{25}$$

provided that $\Delta = 0$ (that is, $\beta_1 = \beta_2$).

IV. APPLICATIONS TO OPTICAL CIRCUITS

To illustrate the application of the theory, let us consider the coupling between two lines (Selfocs) consisting of lens-like dielectric media whose permittivity distribution is

$$\epsilon = \epsilon(0)\{1 - g^2(x^2 + y^2)\} \tag{26}$$

where $\epsilon(0)$ is a maximum value of the permittivity on the center axis of the line, and g is a characteristic parameter determining the rate of change of permittivity variation in the transverse directions x and y (see Fig. 4). For simplicity, we assume the two dielectric lines are identical, that is, the values of $\epsilon(0)$ and g are the same for two lines. The field of the fundamental mode of each line, polarized in the y direction and propagating in the z direction, is given as [8]

$$h_y = \exp\{-\tfrac{1}{2}gk(x^2 + y^2)\} \tag{27}$$

where $k = \omega[\epsilon(0)\mu]^{1/2}$.

The problem of the coupling between these two lines can now easily be solved by applying the theory described in the preceding section. Using the coordinate system shown in Fig. 4, the permittivity distribution of the coupled system is expressed as

$$\epsilon = \begin{cases} \epsilon(0)[1 - g^2\{(x-l)^2 + y^2\}], & (x > 0) \\ \epsilon(0)[1 - g^2\{(x+l)^2 + y^2\}], & (x < 0) \end{cases} \tag{28}$$

and the fields of line 1 and 2 are, respectively,

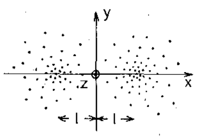

Fig. 4. Coupling between two Selfocs.

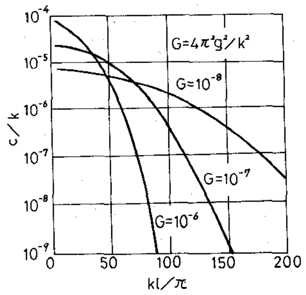

Fig. 5. Coupling coefficient between two Selfocs.

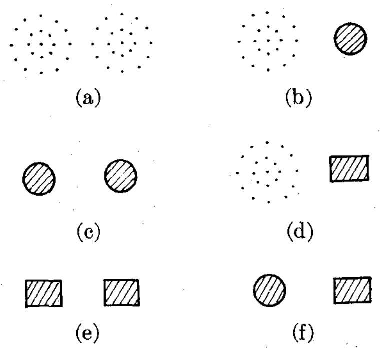

Fig. 6. Examples of coupled optical transmission lines and optical integrated circuits. (a) Coupling between two Selfocs. (b) Coupling between two circular dielectric lines. (c) Coupling between two rectangular dielectric lines. (d) Coupling between Selfoc and circular dielectric line. (e) Coupling between Selfoc and rectangular dielectric line. (f) Coupling between circular and rectangular dielectric lines.

$$\boldsymbol{h}_{t1} = \boldsymbol{i}_y h_{y1} = \boldsymbol{i}_y \exp\left[-\tfrac{1}{2}gk\{(x-l)^2 + y^2\}\right], \quad \text{for line 1}$$

$$\boldsymbol{h}_{t2} = \boldsymbol{i}_y h_{y2} = \boldsymbol{i}_y \exp\left[-\tfrac{1}{2}gk\{(x+l)^2 + y^2\}\right], \quad \text{for line 2.} \tag{29}$$

Substituting both (28) and (29) into (7) and (8) yields N_{ij} and D_{ij}, which in turn determine β in accordance with (10) or (12). The coupling coefficient[3] c can be expressed approximately as

$$c = \frac{g}{\pi^{1/2}} (gkl^2)^{1/2} \exp(-gkl^2) \tag{30}$$

provided that $gkl^2 \gg 1$.

Fig. 5 shows the coupling coefficient c calculated from (30). The parameter $G = 4\pi^2 g^2/k^2$ in Fig. 5 represents the amount of decrease of permittivity at the point which is one wavelength away from the center axis. As we can see from Fig. 5, the coupling coefficient decreases rapidly with increasing separation $2l$ between the two lines.

Fig. 6 shows some typical examples of coupled systems consisting of Selfoc, circular, and/or rectangular dielectric waveguides. These are practically important in connection with applications to optical integrated circuits such as couplers, modulators, power dividers, launchers, etc., in laser engineering.

We have examined, in this section, the coupling between two identical Selfocs [Fig. 6(a)] as an illustrative example, but any arbitrary coupled lines as shown in Fig. 6(b)–(f) can of course be analyzed as well. In fact, the coupling between two identical dielectric rectangular lines [Fig. 6(c)] has been analyzed by means of the proposed method, and the coupling coefficients have been easily derived which coincide exactly with that obtained by Marcatili [3, eq. (56) and (59)] using another method.

[3] c is the "coupling coefficient" defined in Miller's coupled mode theory [9].

V. CONCLUSION

A technique of analyzing the coupled open transmission lines based on the variational method has been proposed. This technique is widely applicable and the procedure of calculation is quite straightforward. We can analyze the widespread variety of any coupled structure in microwave, millimeter wave, and optical wave circuitry by means of the proposed method, as long as we know the electromagnetic fields associated with individual line.

APPENDIX

ACCURACY OF THE PROPOSED METHOD

Let us examine the accuracy of the proposed method with the aid of a simple example, a coupled system consisting of two identical dielectric slab waveguides, shown in Fig. 7. We evaluate the coupling between the lowest TM modes of each guide.

The field expressions for these modes can easily be derived by a conventional method. Substituting these field expressions into (7) and (8), we get N_{ij} and D_{ij}. Using N_{ij} and D_{ij} thus obtained, the propagation constants β_a for mode a and β_b for mode b can be yielded either from (10) or (12).

Let $(\beta_b - \beta_a)$ calculated from (10) be $\Delta\beta_1$, and $(\beta_b - \beta_a)$ calculated from (12) be $\Delta\beta_2$ $(=2c)$, whereas $(\beta_b - \beta_a)$ evaluated by solving the rigorous characteristic equation given by Bracey *et al.* [1] is $\Delta\beta_0$. By comparing $\Delta\beta_1$ and $\Delta\beta_2$ with $\Delta\beta_0$ numerically, we can estimate the accuracy of the proposed method.

The results are shown in Fig. 8. K is a ratio of the permittivity of the slab guide and that of the surrounding medium, and k is a free-space propagation constant in the

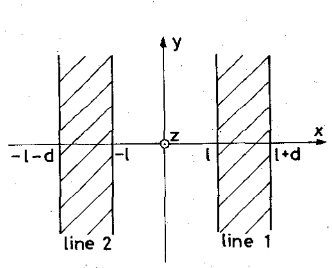

Fig. 7. Coupling of two identical dielectric slab waveguides.

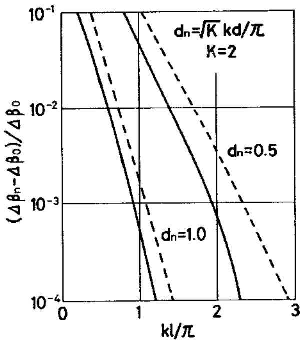

Fig. 8. Accuracy of the proposed method in comparison with a rigorous solution.

surrounding medium. The solid lines show $(\Delta\beta_1 - \Delta\beta_0)/\Delta\beta_0$ while the broken lines show $(\Delta\beta_2 - \Delta\beta_0)/\Delta\beta_0$, both as a function of the normalized frequency (or the normalized separation distance). As we can see from Fig. 8, the accuracy of the proposed method is satisfactorily good. The accuracy becomes excellent as the separation l increases, or the frequency becomes higher (in other words, the amount of the power transmitted within each guide increases).

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An S-Band Radiometer Design with High Absolute Precision

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***Abstract*—A radiometer for the remote measurement of sea surface temperature is described. Two requirements are necessary for the attainment of an absolute accuracy of 1 or 2 K in molecular temperature. Although the first is inappropriate for discussion here, it is clear that corrections must be developed to account for perturbations caused by surface effects (roughness, foaming, and salinity changes) and for atmospheric effects (absorption and scattering). The second requirement, namely, the development of an instrument capable not only of high relative accuracy (i.e., resolution) but also of high absolute precision, is the subject of this paper.**

The concepts underlying the design of an instrument capable of an absolute accuracy of a few tenths degrees Kelvin in the measurement of brightness temperature at *S* band are described. The role of the antenna is discussed and the importance of high ohmic and beam efficiencies is stressed. The hardware itself is fully described, along with an outline on the design of a unique cryogenically cooled termination used to calibrate the whole radiometer, including antenna.

Finally, some test results are presented that show that the design goals for the instrument have been closely approached.

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I. INTRODUCTION AND BACKGROUND

OVER the past 25 years a wide variety of microwave radiometer configurations of varying complexity and sensitivity have been proposed. Although it is common practice in the literature to state the theoretical and experimental temperature resolution of a radiometer, there is almost a complete lack of information on the absolute accuracy achieved. There is a potential need for a spaceborne microwave radiometer to measure remotely the sea surface temperature, in which case it is necessary that the radiometric temperature be determined with a resolution of 0.1 K and with an absolute accuracy approaching ±0.1 K.

Antenna temperature resolutions of 0.1 K are routinely