The Effect of Complex Modes at Finline Discontinuities

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Abstract —**The effect of ignoring complex modes on the solution of finline discontinuity problems is investigated. It is shown that the modal energy distribution at both sides of the discontinuity may be greatly affected by overlooking complex modes, even if they are not strongly excited. It is also shown that disregarding only one mode of a pair of complex modes, while taking the other into account, results in a contradiction to the principle of complex power continuity across the discontinuity plane. Comparison to measured data is also given to justify the validity of the numerical results.**

I. Introduction

THE ANALYSIS of discontinuities between planar transmission lines, in particular microstrip lines and finlines, has received increasing interest [1]–[7]. A proper modeling of such discontinuities is of fundamental importance for any successful printed-circuit design. Only finline discontinuity problems will be considered here. Extending the discussion to other planar structures is, however, straightforward.

Two rigorous approaches have been reported for the analysis of finline discontinuity problems. The first one depends on the transverse resonance concept (e.g., [1]–[3]). A determination of high-order finline modes is not needed for this approach. The problem is completely formulated in terms of homogeneously filled rectangular (or parallel-plate) waveguide modes in conjunction with a proper modeling of the tangential field in the metallization plane. This method inherently shows some disadvantages: The effect of the discontinuity is only available with respect to the dominant mode. No information concerning higher order modes can be obtained. A complex discontinuity, which is composed of a number of cascaded simple discontinuities (e.g. steps in the slot width), has then to be analyzed "as a whole." The properties of the individual simple discontinuities cannot, in general, be used to construct an accurate solution of the complex discontinuity due to the lack of information about high-order modes. The "as a whole" analysis of complex discontinuities may need a large number of basis functions to properly model the tangential field in the plane of the fins. This leads to dealing with oversized matrices, which greatly degrades the numerical efficiency of the method.

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The second approach depends on the modal expansion concept (e.g. [4]–[6]). It is an application of the method of moments, in which both the basis and the testing functions are the electromagnetic fields of the normal modes of propagation at both sides of the discontinuity. Another choice of basis and testing functions has been suggested in [7] following the method which has been presented in [8]. In the authors' opinion, modal fields are in general the best choice for basis and testing functions, because they individually satisfy the same equations (Maxwell's equations) and boundary conditions for the expanded field. Continuity of complex power across the discontinuity plane and hence the unitarity of the scattering matrix are also guaranteed in the modal expansion method, as has been shown in [5]. Generalized scattering and/or transmission matrices, which contain all information about the dominant as well as the higher order modes, are obtained for simple discontinuities (e.g. steps). Complex discontinuities can be analyzed by processing the generalized scattering or transmission matrices characterizing the individual simple steps. The main problem in this method, then, is the accurate determination of an approximately complete set of finline modes.

As has already been shown [9], the singular integral equation (SIE) technique is very efficient for determining such a set. It has also been shown that complex modes can be supported by finlines, so that ignoring these modes in constructing an approximately complete set of finline modes may lead to erroneous solutions.

The possibility of complex modes in a circular waveguide containing a coaxial dielectric rod were first predicted in [10]. It has been shown there that the appearance of a backward-wave mode in a certain frequency band is associated with the appearance of complex modes in a lower frequency band. It has also been shown that complex modes can occur under certain conditions even if there is no frequency range in which backward-wave modes can propagate. More theoretical and experimental investigations on complex modes in dielectric-loaded circular waveguides have been reported, e.g. in [11]–[14].

Complex modes in a shielded rectangular dielectric image guide, which can be considered as a rectangular waveguide with a rectangular dielectric insert, have been reported in [15] and [16]. We have recently shown that complex modes can exist in finlines [9], [17].

This paper addresses the effect of overlooking complex modes on the solution of finline discontinuity problems.

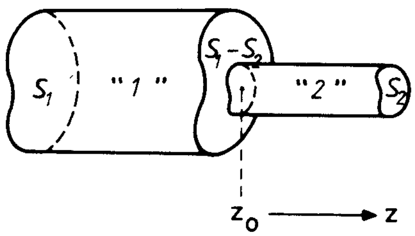

Fig. 1. A boundary-reduction discontinuity between two general waveguides.

II. Basic Formulation

The formulation of the modal expansion method is well documented in the literature and can be found in, e.g., [18] and [19] for homogeneously filled waveguides and in [5] for inhomogeneously filled waveguides. The describing equations can be equally obtained as a result of minimizing certain error quantities, as will be shown in the following.

Consider the boundary reduction discontinuity which is shown in Fig. 1. Waveguides 1 and 2 are assumed to be quite general, except for the restriction that both have discrete modal spectra; this simplifies the discussion to some extent, because all field expansions include in this case only summations and no integrations. Expanding the transverse fields of guides 1 and 2 at the discontinuity plane $z = z_0$ with respect to N-dimensional and M-dimensional sets, respectively, of their modes results in

$$\boldsymbol{E}^{(1)} = \sum_{n=1}^{N} V_n^{(1)} \boldsymbol{e}_n^{(1)}$$

$$\boldsymbol{H}^{(1)} = \sum_{n=1}^{N} I_n^{(1)} \boldsymbol{h}_n^{(1)} \tag{1}$$

$$\boldsymbol{E}^{(2)} = \sum_{m=1}^{M} V_m^{(2)} \boldsymbol{e}_m^{(2)}$$

$$\boldsymbol{H}^{(2)} = \sum_{m=1}^{M} I_m^{(2)} \boldsymbol{h}_m^{(2)} \tag{2}$$

where $\boldsymbol{e}_n^{(1)}(\boldsymbol{h}_n^{(1)})$ and $\boldsymbol{e}_m^{(2)}(\boldsymbol{h}_m^{(2)})$ are the transverse electric (magnetic)-field vectors of the nth mode in guide 1 and the mth mode in guide 2, respectively, and $V_n^{(1)}$, $I_n^{(1)}$, $V_m^{(2)}$, and $I_m^{(2)}$ are expansion coefficients. Looking from side 1, the errors $\boldsymbol{R}_e^{(1)}$ and $\boldsymbol{R}_h^{(1)}$ in the electric and magnetic fields, respectively, are given by

$$\boldsymbol{R}_e^{(1)} = \begin{cases} \boldsymbol{E}^{(1)} - \boldsymbol{E}^{(2)} & \text{on } (S_2) \\ \boldsymbol{E}^{(1)} & \text{on } (S_1 - S_2) \end{cases}$$

$$\boldsymbol{R}_h^{(1)} = \begin{cases} \boldsymbol{H}^{(1)} - \boldsymbol{H}^{(2)} & \text{on } (S_2) \\ \boldsymbol{H}^{(1)} - \boldsymbol{J}_s & \text{on } (S_1 - S_2) \end{cases} \tag{3}$$

where S_1 and S_2 are the cross sections of guides 1 and 2, respectively, and $\boldsymbol{J}_s$ is the induced surface current at $(S_1 - S_2)$. Two error quantities $F_e^{(1)}$ and $F_h^{(1)}$ can now be defined as

$$F_e^{(1)} = \int_{S_1} \left(\boldsymbol{R}_e^{(1)} \times \boldsymbol{H}^{(1)} \right) \cdot \boldsymbol{ds}$$

$$F_h^{(1)} = \int_{S_1} \left(\boldsymbol{E}^{(1)} \times \boldsymbol{R}_h^{(1)} \right) \cdot \boldsymbol{ds}. \tag{4}$$

The error quantity $F_h^{(1)}$ contains the unknown surface current $\boldsymbol{J}_s$, and hence will not be used, in order not to increase the number of unknowns. The error quantity $F_e^{(1)}$ can be regarded as a functional of $\boldsymbol{H}^{(1)}$ which must be minimized. This means that the expansion coefficients $I_n^{(1)}$ of $\boldsymbol{H}^{(1)}$ must be adjusted in order to minimize $F_e^{(1)}$. Substituting (1) and (3) into (4) and carrying out the necessary integrations, one obtains

$$F_e^{(1)} = \sum_{n=1}^{N} I_n^{(1)} \left(P_n V_n^{(1)} - \sum_{m=1}^{M} A_{nm} V_m^{(2)} \right) \tag{5}$$

where

$$\int_{S_1} \left(\boldsymbol{e}_n^{(1)} \times \boldsymbol{h}_m^{(1)} \right) \cdot \boldsymbol{ds} = P_n \delta_{nm} \tag{6}$$

$$\int_{S_2} \left(\boldsymbol{e}_m^{(2)} \times \boldsymbol{h}_n^{(1)} \right) \cdot \boldsymbol{ds} = A_{nm} \tag{7}$$

and δ_{nm} is the Kronecker delta. Performing the usual Fourier expansion procedure by differentiating $F_e^{(1)}$ with respect to $I_n^{(1)}$ $(n = 1, 2, \cdots, N)$ and equating the result to zero, one obtains

$$[\Lambda_p] \boldsymbol{V}^{(1)} = [A] \boldsymbol{V}^{(2)} \tag{8}$$

where $[\Lambda_p]$ is an $(N \times N)$ diagonal matrix with elements P_n; $[A]$ is an $(N \times M)$ matrix with elements A_{nm}; and $\boldsymbol{V}^{(1)}$ $(\boldsymbol{V}^{(2)})$ is an $N(M)$-dimensional column vector with elements $V_n^{(1)}(V_m^{(2)})$.

Looking from side 2, the errors in the electric and magnetic fields are given by

$$\boldsymbol{R}_e^{(2)} = \boldsymbol{E}^{(2)} - \boldsymbol{E}^{(1)}$$

$$\boldsymbol{R}_h^{(2)} = \boldsymbol{H}^{(2)} - \boldsymbol{H}^{(1)}. \tag{9}$$

Again, two error quantities can be defined, namely

$$F_e^{(2)} = \int_{S_2} \left(\boldsymbol{R}_e^{(2)} \times \boldsymbol{H}^{(2)} \right) \cdot \boldsymbol{ds}$$

$$F_h^{(2)} = \int_{S_2} \left(\boldsymbol{E}^{(2)} \times \boldsymbol{R}_h^{(2)} \right) \cdot \boldsymbol{ds}. \tag{10}$$

The error quantities $F_e^{(2)}$ and $F_h^{(2)}$ can be regarded as functionals of $\boldsymbol{H}^{(2)}$ and $\boldsymbol{E}^{(2)}$, respectively, which must be minimized. Due to the completeness properties of the normal modes, it can be proven that in the limiting case when N, M tend to infinity, minimizing $F_e^{(2)}$ with respect to the expansion coefficients $I_m^{(2)}$ of $\boldsymbol{H}^{(2)}$ results in a matrix equation, which is equivalent to (8). Hence, $F_h^{(2)}$ must be minimized with respect to the expansion coefficients $V_m^{(2)}$ of $\boldsymbol{E}^{(2)}$. $F_h^{(2)}$ is readily proved to be given by

$$F_h^{(2)} = \sum_{m=1}^{M} V_m^{(2)} \left(Q_m I_m^{(2)} - \sum_{n=1}^{N} A_{nm} I_n^{(1)} \right) \tag{11}$$

where

$$\int_{S_2} \left(\boldsymbol{e}_n^{(2)} \times \boldsymbol{h}_m^{(2)} \right) \cdot \boldsymbol{ds} = Q_m \delta_{mn}. \tag{12}$$

Differentiating $F_h^{(2)}$ with respect to $V_m^{(2)}$ $(m = 1, 2, \cdots, M)$ and equating the result to zero, one obtains

$$[\Lambda_Q] \boldsymbol{I}^{(2)} = [A]' \boldsymbol{I}^{(1)} \tag{13}$$

where $[\Lambda_Q]$ is an $(M \times M)$ diagonal matrix with elements Q_m, and $\boldsymbol{I}^{(1)}(\boldsymbol{I}^{(2)})$ is an $N(M)$-dimensional column vector with elements $I_n^{(1)}(I_m^{(2)})$. Equations (8) and (13) are the characteristic equations of the modal expansion method. They are equivalent to, e.g., [5, eq. (3)], which has been derived using another approach.

From the above discussion, it is easily seen that the expansion coefficients of the fields at both sides of the discontinuity are so adjusted that the error quantities $F_e^{(1)}$ and $F_h^{(2)}$ are minimized. This can be viewed as "a similarity balance" process, in which the fields at both sides of the discontinuity are "similar" with respect to minimum error quantities $F_e^{(1)}$ and $F_h^{(2)}$. The mode coupling coefficient A_{nm} defined by (7) represents a measure of the degree of similarity between the nth mode of guide 1 and the mth mode of guide 2. Two different modes in either guide 1 or guide 2 are then completely "dissimilar" due to the orthogonality relations (6) and (12).

III. Effect of Ignoring Modes at Either Side of the Discontinuity

According to the similarity balance concept discussed above, the nth mode excited in guide 1 (which will be called mode (a)) is balanced by exciting a similar field in guide 2. This similar field is, in general, composed of a superposition of all the M modes of guide 2, the magnitude of each depending on its degree of similarity to mode (a). In particular, the magnitude of a mode with a high degree of similarity will dominate the magnitudes of the other, less similar modes. This similarity balance holds for each of the N modes of guide 1.

Let us assume now that the mth mode of guide 2 (which will be called mode (b)) has the largest degree of similarity to mode (a). Omitting mode (b) from the M modes of guide 2 can only be compensated by increasing the magnitudes of modes that are less similar to mode (a), in order to restore the similarity balance. This will disturb the modal distributions (and hence the stored energy) at both sides of the discontinuity. It is important to note that this disturbance does not necessarily depend on how strongly mode (a) is excited, because balancing a weakly excited mode in guide 1 may require strongly excited modes in guide 2 which have a very weak degree of similarity to that mode. Omitting, however, both mode (a) and mode (b) will have a much smaller effect on the modal distributions, in particular, if both are just weakly excited.

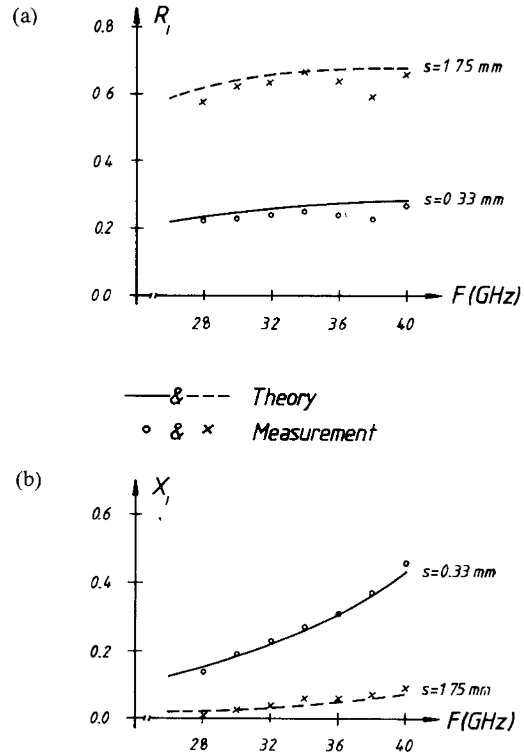

Fig. 2. Frequency response of the normalized input impedance of the waveguide–finline junction shown in Fig. 3. (a) Normalized input resistance. (b) Normalized input reactance.

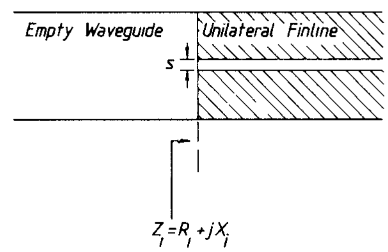

Fig. 3. A waveguide–finline discontinuity. Parameters: WR-28 housing; substrate thickness = 0.254 mm; substrate dielectric constant = 2.22.

IV. Effect of Overlooking Complex Modes on Finline Discontinuities

As has already been shown in [9], finline modes change their nature as any of the finline parameters changes. A pair comprising an inductive and a capacitive evanescent mode may become a complex pair, and vice versa, as, e.g., the slot width changes. If we would analyze a discontinuity using only usual (noncomplex) modes, it can happen that a pair (or more) of modes at one side of the discontinuity is noncomplex, while the corresponding pair at the other side, which has the largest degree of similarity, is a complex one. Both the modal distributions and the stored energy at both sides of the discontinuity would then be greatly affected, even if the former pair is not strongly excited. The situation is much more favorable if both pairs are complex, so that both would be ignored in the matching process.

Another interesting effect is that of disregarding only one mode of a pair of complex modes, while taking the

TABLE I
MODAL ENERGY DISTRIBUTION OF THE DISCONTINUITY SHOWN IN FIG. 4 WITH AND WITHOUT TAKING COMPLEX MODES INTO ACCOUNT

(a) at finline 1

mode order	1	2	3	4	5	6	7	8	9	10
ß 1/(mm)	0.4853	-j0.6209	-j1.1519	-j1.6490	-j1.6513	-j1.7001	-j1.7043	-j1.8729	-j1.8770	-j2.0855
energy with C.M.	+0.2554	+0.0046	+0.1409	+0.0001	+0.0043	-0.1749	+0.0976	+0.0009	-0.0009	-0.0632
energy without	-0.0003	+0.0044	+0.1204	+0.0001	+0.0033	-0.2510	+0.1377	+0.0009	-0.0008	-0.0911

mode order	11	12	13	14	15	16	17	18	19	20
ß 1/(mm)	-j2.1089	-j2.1094	-j2.4184	-j2.4409	-j2.5746	-j2.6949	-j2.7514	-j3.0220	-j3.1225	-j3.1641
energy with C.M.	+0.0610	+0.0192	+0.0005	-0.0007	+0.0065	-0.0279	+0.0081	+0.0724	+0.0030	-0.0012
energy without	+0.0000	+0.0000	+0.0006	-0.0011	+0.0037	-0.0458	+0.0061	+0.0264	+0.0005	-0.0392

(b) at finline 2

mode order	1	2	3	4	5	6	7	8	9	10
ß 1/(mm)	0.7024	-j0.6037	-j0.7271	-j1.5945	-j1.6488	-j1.6772	-j1.7427	0.0073 -j1.8699	-0.0073 -j1.8699	-j1.8886
energy with C.M.	-0.8219	+0.0630	+0.2718	+0.0381	+0.0000	+0.0054	+0.0375	-0.0008	-0.0008	-0.0016
energy without	-0.6629	+0.0879	+0.3971	+0.1023	+0.0000	+0.0198	+0.1278	-------	-------	-0.0000

mode order	11	12	13	14	15	16	17	18	19	20
ß 1/(mm)	-j1.9680	-j2.3999	-j2.4095	-j2.4667	-j2.4745	-j2.5318	-j2.6976	-j3.0679	-j3.1139	-j3.2065
energy with C.M.	+0.0006	+0.0164	-0.0174	-0.0035	+0.0181	+0.0002	+0.0175	+0.0035	-0.0167	-0.0153
energy without	+0.0221	+0.0058	-0.0132	-0.0053	+0.0022	+0.0002	+0.0089	+0.0032	-0.0058	-0.0027

Operating frequency = 30 GHz.

other into account. As has been shown in [9], one mode of a pair of complex modes propagates in the same direction in which it is attenuated (let it be called mode (c)), while the other propagates opposite to the direction in which it is damped (let it be called mode (d)). Each carries, by itself, neither active nor reactive power. Let us now assume that both modes have been excited in guide 2. Mode (c) (mode (d)) represents for guide 1 an energy loss (gain) mechanism, which occurs in guide 2. Guide 1 does not "know" that each of these modes carries, by itself, no power. If mode (d) (mode (c)) is disregarded while mode (c) (mode (d)) is retained in the matching process, the amplitudes of the different modes in guide 1 are adjusted to account for the energy loss (gain), which occurs in guide 2. In other words, guide 2 appears lossy (active) if it is looked at from guide 1. The complex power, which is calculated in guide 1, takes this energy loss (gain) into account. On the other hand, the complex power, which is calculated in guide 2, does not "feel" any energy loss (gain) due to the absence of the power carried by mode (c) (mode (d)). This consequently results in a discontinuity in the complex power across the junction. This effect cannot be compared to the truncation effect (i.e., matching finite number of modes at both sides of the discontinuity), because continuity of complex power is independent of the number of modes which enter the matching process, as has been shown in [5].

V. Numerical Results

In order to check the accuracy of the numerical results, comparison to measured data in *Ka*-band is demonstrated in Fig. 2, which shows the frequency response of the resistive (Fig. 2(a)) and reactive (Fig. 2(b)) parts of the normalized input impedance of the waveguide–finline junction shown in Fig. 3. For the two indicated cases, convergence of the numerical results has been achieved by using only five modes at each side of the discontinuity. For both slot widths ($s = 1.75$ mm, $s = 0.33$ mm), the order of the first complex modes is larger than five; hence, complex modes have negligible influence on the numerical results for these two cases. The reader can refer to [5] for further information about the convergence problem.

Table I shows the stored energy distributions at both sides of the finline discontinuity shown in Fig. 4, computed with and without taking complex modes into consideration. Due to the large slot width ratio ($s_1/s_2 = 35$), convergence had to be achieved by using 20 modes at both sides of the discontinuity. The incident field is the dominant mode of finline 1 carrying unit power. The dominant mode of finline 2 shows a standing wave pattern within the distance l between the discontinuity and the open circuit. It stores capacitive energy because l is slightly larger than half a guide wavelength for this mode ($l = 5.0$ mm, $\lambda_{g2} = 8.945$ mm). If complex modes are omitted, the stored energy is calculated with an error of 19.34 percent. The total energy, on the other hand, which is stored in both the dominant mode and all higher order modes turns out to amount to -0.4059 taking complex modes into account while it is $+0.1272$ if these modes are omitted. The corresponding normalized input impedance at the discontinuity plane is $-j0.1039$ with and $+j0.0001$ without complex modes. The error is larger than 100 percent because the effect of overlooking complex modes has changed the capacitive nature of the structure between the discontinuity and the open circuit into an inductive one. It should be pointed out that this severe error is due to overlooking only one pair of complex modes (namely the eighth and ninth modes of finline 2), which are only weakly excited. The error would be much more disastrous if many complex mode pairs existed on one or both sides of the discontinuity.

The frequency response of the normalized input reactance of the same discontinuity (seen at the discontinuity plane) is plotted in Fig. 5, computed with and without complex modes. The irregularity of the curve representing the computation without complex modes is due to the fact

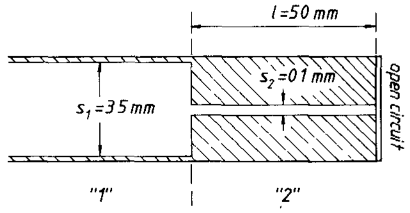

Fig. 4. A unilateral finline discontinuity. Parameters: as in Fig. 3.

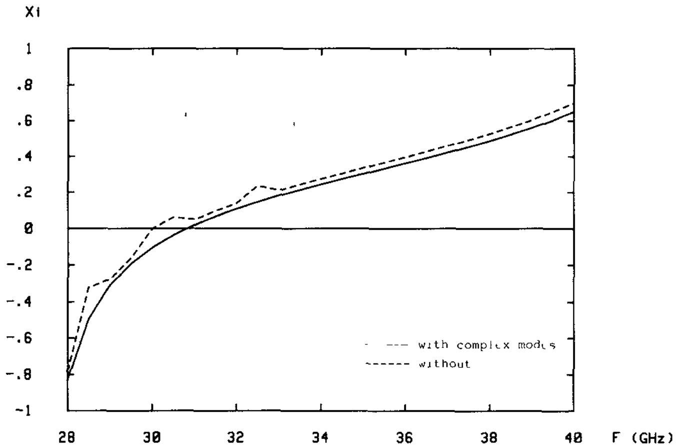

Fig. 5. Frequency response of the normalized input reactance of the finline discontinuity shown in Fig. 4.

that some of the modes change their nature (from complex to noncomplex and vice versa) as the frequency changes. A pair of complex modes which is not taken into account in a certain frequency range may become a noncomplex pair in another frequency range. The reader can refer to [16] for a deeper understanding of this phenomenon. Apart from this irregularity, a deviation of about 2.6 percent in the resonance frequency (at which $X_i = 0$) can be observed. This is a crucial error for any narrow-band application. The deviation in the upper half of the *Ka*-band is at least 8 percent, while the convergence of the results has been achieved within only 1 percent.

Finally, in order to demonstrate the effect of disregarding only one mode of a pair of complex modes while taking the other into account, the structure shown in Fig. 4 (with the distance l being shortened to $l = 0.5$ mm) has been analyzed. The complex powers carried by the first eight modes at both sides of the discontinuity once by disregarding the ninth mode and once by disregarding the eighth mode of finline 2 are tabulated in Tables II and III, respectively. Although the structure has been assumed lossless and passive, the complex power carried by the dominant mode of finline 1 has an active part, which is negative (positive) in Table II (III) in order to account for the disappearance of the ninth (eighth) mode of finline 2. The dominant mode of finline 2 stores capacitive energy due to its standing wave nature between the discontinuity and the nearby open circuit. The complex power carried individually by the eighth and ninth modes of finline 2 vanishes as expected. The total complex powers carried by

TABLE II
COMPLEX POWER DISTRIBUTION OF THE DISCONTINUITY SHOWN IN FIG. 4 WITH THE NINTH MODE OF FINLINE 2 DISREGARDED

(a) finline 1

mode order	1	2	3	4	5	6	7	8	Total
ß 1/(mm)	0.4853	-j0.6209	-j1.1519	-j1.6490	-j1.6513	-j1.7001	-j1.7043	-j1.8729	
complex power	- 0.0230 -j0.5002	+j0.0134	+j0.3870	+j0.0002	+j0.0112	-j0.4652	+j0.2661	+j0.0191	- 0.0230 -j0.2684

(b) finline 2

mode order	1	2	3	4	5	6	7	8	Total
ß 1/(mm)	0.7024	-j0.6037	-j0.7271	-j1.5945	-j1.6488	-j1.6772	-j1.7427	+ 0.0073 -j1.8699	
complex power	-j0.8674	+j0.0909	+j0.4834	+j0.2273	+j0.0000	+j0.0311	+j0.3401	+j0.0000	+ 0.0000 +j0.3054

Distance l shortened to 0.5 mm.

TABLE III
COMPLEX POWER DISTRIBUTION OF THE DISCONTINUITY SHOWN IN FIG. 4 WITH THE EIGHTH MODE OF FINLINE 2 DISREGARDED

(a) finline 1

mode order	1	2	3	4	5	6	7	8	Total
ß 1/(mm)	0.4853	-j0.6209	-j1.1519	-j1.6490	-j1.6513	-j1.7001	-j1.7043	-j1.8729	
complex power	+ 0.0239 -j0.5195	+j0.0139	+j0.4019	+j0.0002	+j0.0116	-j0.4831	+j0.2764	+j0.0198	+ 0.0239 -j0.2788

(b) finline 2

mode order	1	2	3	4	5	6	7	9	Total
ß 1/(mm)	0.7024	-j0.6037	-j0.7271	-j1.5945	-j1.6488	-j1.6772	-j1.7427	- 0.0073 -j1.8699	
complex power	-j0.9008	+j0.0944	+j0.5020	+j0.2360	+j0.0000	+j0.0323	+j0.3532	+j0.0000	+ 0.0000 +j0.3171

Distance l shortened to 0.5 mm.

the modes of finlines 1 and 2 are not equal, which contradicts the principle of complex power continuity.

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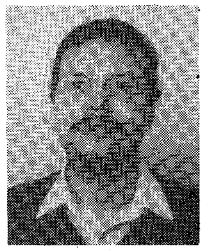

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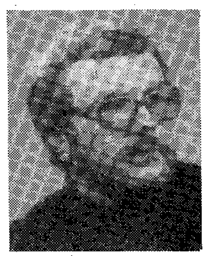

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