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CHARACTERIZATION OF STRIPLINE CROSSING BY TRANSVERSE RESONANCE ANALYSIS

T. Uwano,* R. Sorrentino,** T. Itoh***

*Matsushita Electric Ind. Co., Ltd, Osaka, Japan
**Dept. of Electrical Engineering, University of Rome Tor Vergata, Roma, Italy
***Dept. of Electrical Engineering, University of Texas, Austin, Texas, USA

ABSTRACT

A method of analysis is described for characterizing the discontinuities made of orthogonally crossed two striplines on a suspended structure. The method of analysis is based on generalized transverse resonant technique extended here to 4-port configurations. The technique is used for determination of resonant structure at a given frequency and subsequently the equivalent circuit parameters of the discontinuities.

INTRODUCTION

In advanced planar microwave and millimeter-wave integrated circuits, transmission lines are often created on both sides of the substrate (1). These lines may be in parallel or orthogonal to each other. The present paper reports characterization of the discontinuities of orthogonally crossed two striplines as shown in Fig. 1. It is made up of the crossing between two orthogonal striplines on opposite sides of the same substrate and the auxiliary conducting planes to enclose the substrate. The auxiliary walls are used for field analysis purpose. They permit the structure to be analyzed as a rectangular waveguide discontinuity problem.

The method for analysis is based on the "generalized transverse resonance technique" introduced for finline step discontinuity problem (2). The technique is extended here to a 4-port configuration treated in this paper. The method consists of two parts. First, the resonant structure created by auxiliary walls is described in terms of network representation containing a reactive 4-port. For a fixed resonant frequency, we try to find as many resonator sizes as required for extraction of 4-port matrix elements. The second part of analysis is a full-wave electromagnetic field analysis in which the resonant frequency is found as an eigenvalue problem.

CIRCUIT REPRESENTATION

A procedure for a 2-port resonance method (3)is extended to a 4-port configuration as follows. The crossing between the two suspended striplines can be represented as a 4-port network at some reference planes sufficiently far from the discontinuity region. Each port is terminated with a reactance corresponding to the line section between the reference plane and the auxiliary wall as shown in Fig. 2. The network equations for the entire circuit are expressed in matrix form as

$$[\,[Z] + \mathrm{diag}\,[Z_i]\,]\,[I] = 0$$

where [Z] is the normalized impedance matrix of 4-port network, Z_i (i = 1,2,3,4) are the normalized terminal impedances : $Z_i = j \tan \beta_i l_i$, and l_i (i = 1,2,3,4) are the stripline lengths between the reference planes and the auxiliary walls. [I] is the vector of the currents I_i (i = 1,2,3,4) as shown in Fig. 2. In the absence of losses, [Z] is imaginary. The resonant frequency is obtained from the condition that, in the absence of sources, the voltages and currents are non-trivial, and hence, that the determinant of the matrix equals zero. The impedance matrix of a reciprocal 4-port lossless network possesses in general 10 independent imaginary parameters. In the present case, however, because of the symmetry of the structure, only five parameters are needed to characterize the Z matrix. Now, by properly choosing the terminal impedance Zi's, namely $Z_1=Z_2$ and $Z_3=Z_4$ (thus $l_1=l_2$, $l_3=l_4$), the resonant condition is simplified so that the problem is solved analytically. If these conditions are applied, the equation for the resonant condition can be factorized so that it can be written in the form:

$$Z_{11} + Z_1 - Z_{12} = 0$$

or

$$Z_{33} + Z_3 - Z_{34} = 0$$

or

$$(Z_{11} + Z_{12} + Z_1)(Z_{33} + Z_{34} + Z_3) - 4 Z_{13}^2 = 0 \quad .$$

Each equation corresponds to a different resonant behavior as shown by the corresponding [I] eigenvector. For the first equation, the eigenvector is $I_1=-I_2$ and $I_3=I_4=0$. This condition corresponds to an odd resonance of the structure shown in Fig. 3(a). The structure behaves as if an electric wall is placed symmetrically along at the center of stripline 2. For the given resonant frequency, the required resonance condition provides the quantity $Z_{11}-Z_{12}$ from the value of Z_1. Similary, the condition for the second equation is shown in Fig. 3(b). For the last equation, the eigenvector for an even resonance is obtained : $I_1=I_2$ and $I_3=I_4$. Substitution of these conditions into an original 4-port matrix yields the 2-port network matrix equation

$$\begin{bmatrix} V_1 \\ V_3 \end{bmatrix} = \begin{bmatrix} Z_{11} + Z_{12} & 2 Z_{13} \\ 2 Z_{13} & Z_{33} + Z_{34} \end{bmatrix} \begin{bmatrix} I_1 \\ I_3 \end{bmatrix}$$

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The use of symmetry, therefore, has reduced the 4-port network problem to that of a 2-port, whose impedance matrix is solved by a 2-port resonance method. Combining the results with those for the two structures in Fig. 3, we obtain all five Z parameters.

A possible equivalent circuit respresentation which takes into account the symmetry properties of the structure of Fig. 1 is shown in Fig. 4. Zc is used to represent the coupling capacitance between the two strips. Zp+Zp and Zq+Zq represent the inductances associated with two stripline sections, while Za and Zb represent the two strip-to-ground capacitances.

FIELD ANALYSIS

The resonant structure of Fig. 1 can be subdivided into three homogeneous regions as shown in Fig. 5. In terms of the TE-to-z and TM-to-z representations, the hybrid fields in each homogeneous region in the resonator structure are represented by the sets of orthogonal functions with unknown coefficients. The boundary conditions at each interface are then imposed. The magnetic field discontinuity between the opposite sides of each metallic strip Si (i = 1,2) is expanded in terms of a set of known orthogonal vector functions $\chi_v^{(i)}$ defined over Si :

$$\Delta \mathbf{H}_t^{(i)} = \sum_{v} P_v^{(i)} \chi_v^{(i)} \quad i = 1, 2$$

For a faster numerical convergence, the singular behavior of the magnetic field component normal to the stripline edges can be incorporated in the above expression by a proper choice of the basis functions. The boundary conditions lead to a homogeneous system of equations in terms of the unknown coefficients Pv's. The condition for non-trivial solution determines the characteristic equation of the given structure. This equation may be regarded as a function of ω, $l1$, $l3$ equated to zero :

$$f(\omega, l1, l3) = 0$$

For given value of $\omega = \omega_r$, this can be solved to evaluate the different pairs of $l1$ and $l3$ giving rise to the same resonant frequency ω_r. These values of $l1$ and $l3$ can be used for computing the discontinuity parameters discussed in the previous section.

COMPUTED RESULTS

In the computations, the substrate was placed symmetrically between the top and the bottom planes, with both strips having the same widths. Hence, the impedance matrix representation of the discontinuities has Z11=Z33 and Z12=Z34. The element values of the equivalent circuit calculated at three different frequencies are quoted in Table 1 along with the structural parameters used in the computations. Reference planes were placed at the strip edges. Note that the elements for Za and Zb (= Za) are capacitors with negative value. This is acceptable because they compensate the parallel distributed capacitance for the isolated stripline in the absence of the other stripline. Fig. 6 shows the corresponding S parameters of the discontinuities.

Fig. 7(a)(b) show the longitudinal and transverse components of the current densities at the center of the strip at the three frequencies. The number of terms MN of the field expressions in the homogeneous regions of Fig. 5 as well as the number v of basis functions are also given. In Fig. 7(a), it was observed that each figure was of perturbated cosine form. The longitudinal current, hence, may be represented by the combination of a cosine and an additional polynomial functions (4) so that the computation time could be reduced.

Table 1 Element values of equivalent circuit

	0.5 GHz	1 GHz	2 GHz
Lp=Lq	0.331nH	0.331nH	0.329nH
Ca=Cb	-0.0885pF	-0.103pF	-0.101pF
Cc	0.249pF	0.272pF	0.258pF

$$Lp = Lq = \frac{Zo\,Zp}{j\,\omega}$$

$$Ca = Cb = \frac{1}{j\,\omega\,Zo\,Za}$$

$$Cc = \frac{1}{j\,\omega\,Zo\,Zc}$$

Zp, Zq, Za, Zb, Zc : Normalized impedance

Zo : Stripline characteristic impedance
(159.2, 159.1, 158.3Ω at 0.5, 1, 2 GHz)

h1 = h2 = 5 mm, t = 1 mm

w1 = w2 = 1 mm, $\varepsilon r = 3.8$

CONCLUSIONS

A method of analysis has been described for characterizing the discontinuities of two crossed striplines. The method is based on a generalized transverse resonance technique for computing the resonant frequency of a resonator created by enclosing the crossing with auxiliary perfectly conducting walls. This resonator problem is analyzed as the waveguide scattering for waves traveling in the direction normal to the substrate surface. For a specified frequency, resonant structures are found by adjusting the lengths of the strips and hence the resonator size. These structures are used for deriving the equivalent circuit parameters characterizing the discontinuity.

This method can also be applied for the characterization of stripline-slotline transition.

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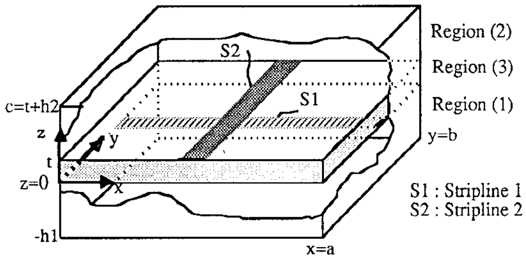

Fig. 1 Structure for the problem

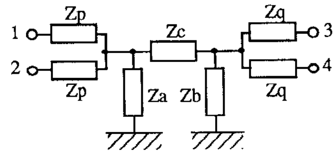

Fig. 4 Equivalent circuit for the problem

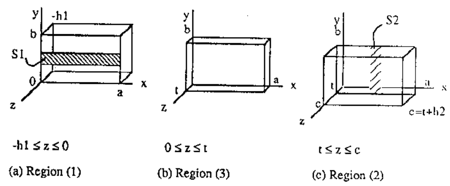

Fig. 5 Subregions for field analysis

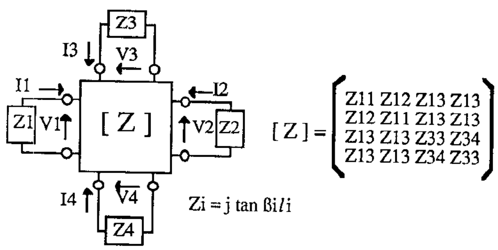

Fig. 2 4-port network for the problem

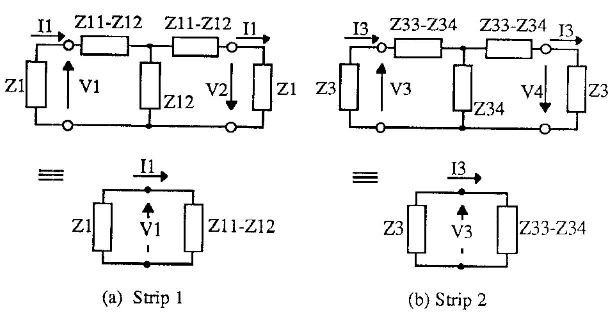

Fig. 3 Equivalent circuits for an odd resonance

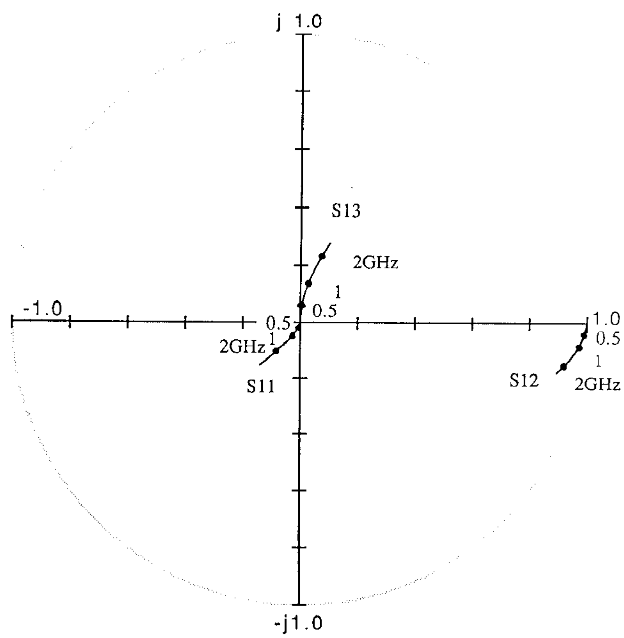

Fig. 6 S parameters of the discontinuites of the structure

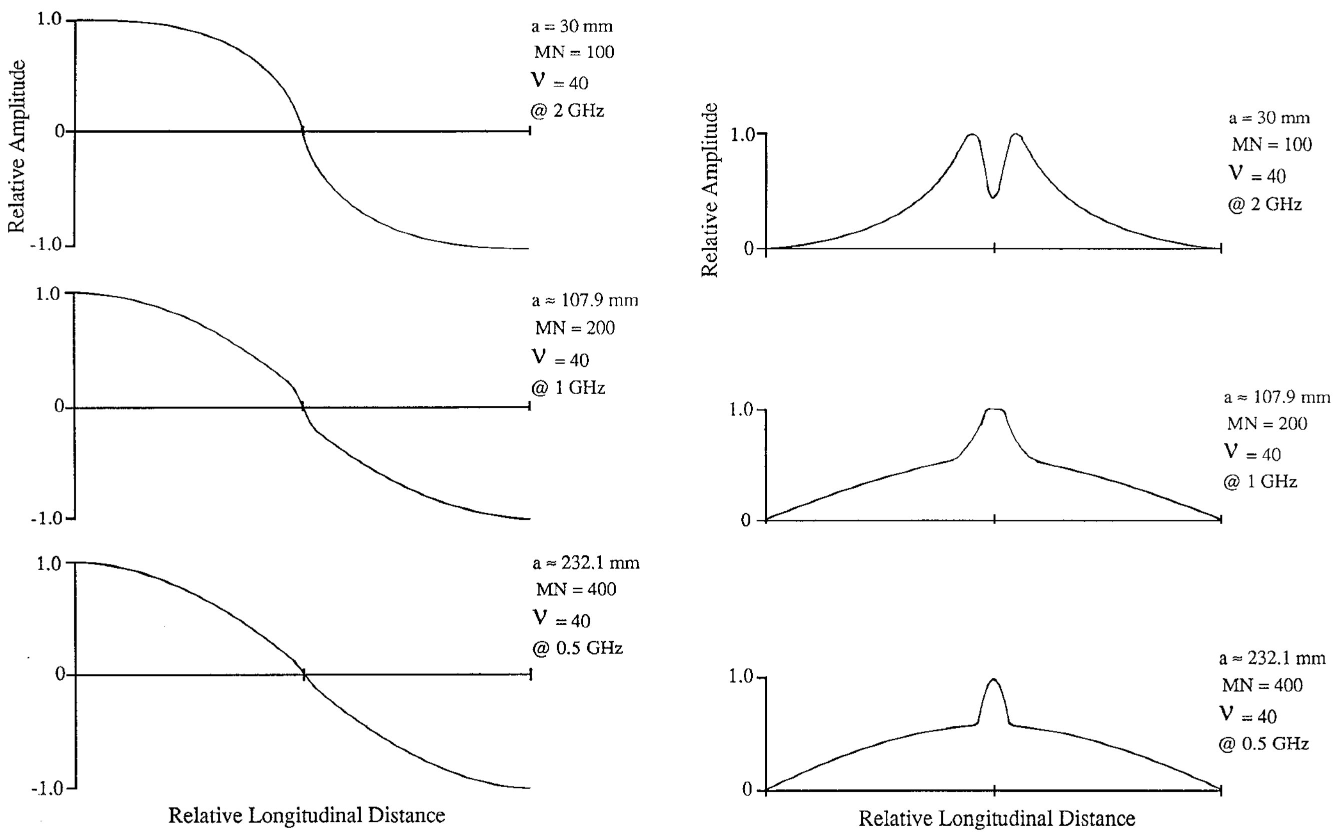

Fig. 7 Current distribution on the stripline (a) longitudinal current (b) transverse current