A Dynamic Model for Microstrip--Slotline Transition And Related Structures

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Abstract

An analysis of microstrip to slotline transition is presented. The method of moments is applied to the coupled integral equations. In the formulation, the Green's function for the grounded dielectric substrate is used, which takes into account all the radiation, surface wave and substrate effects. Meanwhile, all the mutual coupling effects are included in the method of moments solution. Certain related structures, such as slotline and microstrip discontinuities, a slot fed by a microstripline and a printed strip dipole fed by a slotline, can also be solved with this analysis. The present approach may find applications to other related transitions in MIC design.

I. Introduction

Quasi-static methods, equivalent waveguide models and equivalent circuit models have been widely used in the modeling of microstrip or slot type discontinuities in microwave and millimeter wave devices in the past [1]. Recently, a more rigorous approach which takes into account all the physical effects including radiation and surface waves has been applied to certain microstrip discontinuities [2]-[5]. In this scheme, the method of moments which determines the current on the strip or electric field on the slot is implemented in the solution of the Pocklington integral equation. The exact Green's function for a grounded dielectric substrate due to either an electric dipole or a magnetic dipole has been used, which includes all the physical effects. Based on this approach, a dynamic model for micostrip-slotline transition and its related structures such as a microstrip fed slot and a slotline fed printed dipole is proposed in this paper. The developed model with some modifications can be applied to other types of transition in MIC and MMIC design. In microstrip-slotline transition, a short-circuit slotline which is etched on one side of the substrate is crossed at a right angle by an open-circuit microstrip on the opposite side. This type of transition makes the two level circuit design possible [1]. Some experimental work has been reported in [6]-[7] and a transmission line circuit model was reported in [8]. In the present approach, the radiation and surface waves due to the cross-junction, the line discontinuities and all the mutual coupling due to the dominant mode as well as higher order modes of each line are included in the method of moments solution. The VSWR and input impedance of the transition, can be determined by the current distribution on the microstripline in conjunction with transmission line theory. One of the main features of the present method is the modeling of two coupled half infinite lines where expansion modes composed of piecewise sinusoidal modes, traveling wave and standing wave modes. In the formulation procedure, certain important problems in MIC, MMIC or printed antennas design can also be solved. When only one of the lines is discussed, the present approach is simplified to the modeling of slotline or microstripline discontinuities as reported in [2]-[5]. If one of the lines is of finite length, then the method presented herein can be applied to the modeling of microstripline fed slot or slotline fed printed strip dipole. The results of short-circuit slotline reactance and the microstripline fed slot input impedance will be given as an illustration.

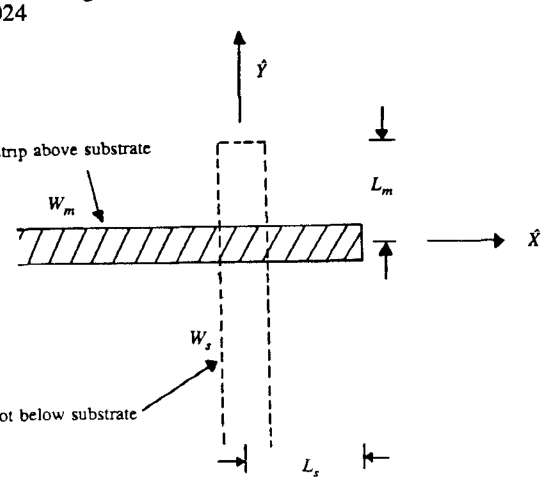

Figure 1 Microstrip- Slotline Transition.

II. Outline of the method

The microstrip-slotline transition is shown in Fig.1, where the lines are extended certain distance from the cross-junction, to act as the tuning stubs. Due to the fact that the width of the lines is much smaller than wavelength, the transverse vector components (J_y and M_{mx}) on the lines are second order effects, and are neglected for simplicity. The coupled integral equations are obtained by the boundary condition that E- field is zero in the microstrip and H- field is continuous across the slotline and they are given by

$$E_x = \int G_{xx} J_x \, ds_m + \int G_{xy} M_{my} \, ds_s \qquad (1a)$$

$$H_y = \int G_{yx} J_x \, ds_m + \int G_{yy} M_{my} \, ds_s \qquad (1b)$$

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where G_{xx} and G_{yx} are the dyadic Green's function components due to an $\hat{x}$- directed infinitesimal electric dipole on the top of the substrate and G_{xy} and G_{yy} are the dyadic Green's function components due to a $\hat{y}$- directed infinitesimal magnetic dipole on the ground plane. J_x is the current on the microstrip s_m, while M_{my} is the magnetic current (electric field) on the slotline s_s. In the method of moments procedure, the J_x and M_{my} are expanded in terms of a set of known functions. For the transition under consideration, the modelling of two half infinite lines is necessary. The most convenient way is to use piecewise sinusoidal modes near the cross-junction and half infinite travelling wave on the lines as described in [4]-[5]. The transverse dependence of J_x and M_{my} are assumed to be Maxwell's distribution, such that the edge condition is satisfied. When Galerkin's method is performed in Equations (1a) and (1b) in spectral domain, it is necessary to compute seven different double integrations, which involve self reaction and mutual coupling of different expansion modes. Pole extraction and asymptotic extraction techniques are also applied to speed up convergence in numerical integrations. It can be seen that if the second term in the right hand side of equation (1a) and first term in the right hand side of equation (1b) are deleted, the problem reduces to the determination of end effects of the microstrip or slotline. Also if one of the lines is terminated in a finite length, the problem is reduced to finding the dipole or slot input impedance.

III. Numerical results

(A) Microstrip- slotline transition

The results of microstrip-slotline transition are obtained based on the developed algorithm. The numerical analysis is performed on the IBM 3090 system. Typically, for each data it takes about one minute and thirty seconds of computer time in contrast to half second to obtain the propagation constant k_m, although a lot of effort has been made to reduce the computer cost. An example of a 50Ω microstripline to a 80Ω sloline transition is given. The results of the VSWR are shown in Fig. 2. The results are first checked by interchanging the feed line and parasitic line. The differences in $|\Gamma|$ are within 2% which is consistent with the property of low loss two port networks. The VSWR obtained by the transmission line circuit model and the measurement [7] are also shown in Fig. 2 to provide a comparison. In the transmission line circuit model the stub length is assumed measured from the center of each line and the propagation constants k_m and k_s and the excess length are obtained in the current analysis. It is seen from Fig. 2 that the present method agrees very well with the circuit model. The measurements reported in [7] show wider bandwidth than that of either the circuit model or the present analysis. It is believed that the accuracy of the device parameters, the nonideal match load and especially, the coaxial to microstripline transition will more or less affect the frequency dependent results in the measurement.

(B) Slotline discontinuities

Since the results of open end microstrip discontinuities have been reported with the present approach in [3] and [4], they are not repeated here. Experimental results of shorted end slotline discontinuities were reported in [8]. A spectral domain approach (SDA) of this problem has been reported in [9], where closed coplannar waveguide is used and surface wave and radiation effects are not taken into account. The shorted end reactance of a slotline are shown in Fig. 3 as a function of various device parameters. The normalized reactance obtained here is also compared with measurements reported in [8] and the SDA method reported in [9]. The comparison shows very good agreement with the SDA method.

(C) Microstripline fed slot

The microstripline fed slot has been suggested for use in monolithic antennas where the slot is either used in aperture coupled microstrip antennas or as a radiating element [10]. The input impedance of a slot with a tuning stub is obtained from the information of the reflection coefficient in the microstrip line. The normalized input impedance of a slot fed by a microstrip as a function of frequency is plotted on the Smith chart as shown in Fig. 4. It is observed that the resonant frequency in this case occurs at about 2.73 GHz and the bandwidth is about 2%. It is seen that the bandwidth is mainly determined by the tuning stub since the impedance follows approximatedly $\bar{R} = 1$ circle as frequency changes. Therefore to increase the bandwidth, the stub length should be chosen such that at resonant frequency the change of stub impedance with frequency is as small as possible. Another way of increasing bandwidth is to control the device parameters such that resonance occurs even without the tuning stub.

VI. Conclusion.

A full wave analysis is proposed to develop a generalized dynamic model for microstrip-slotline transition, microstrip-slot, slotline-microstrip dipole as well as slotline and microstrip discontinuities. The present approach may find applications in other transitions in MIC or MMIC design.

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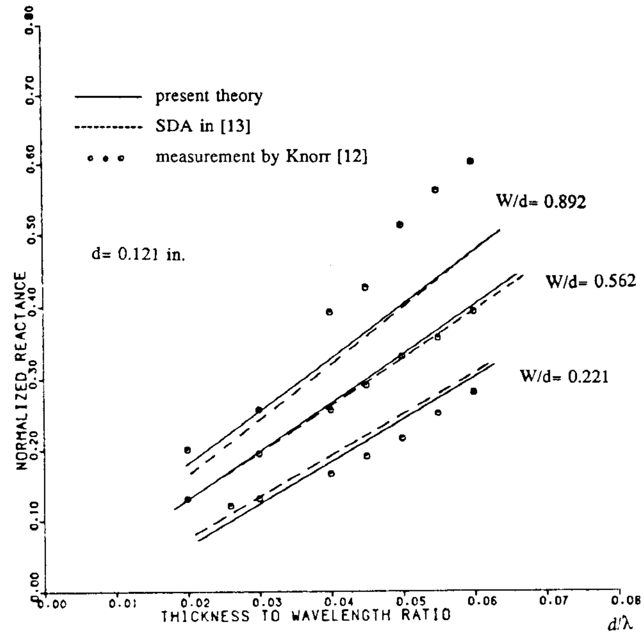

Figure 3. Normalized reactance of a shorted slot, $\varepsilon_r = 12$.

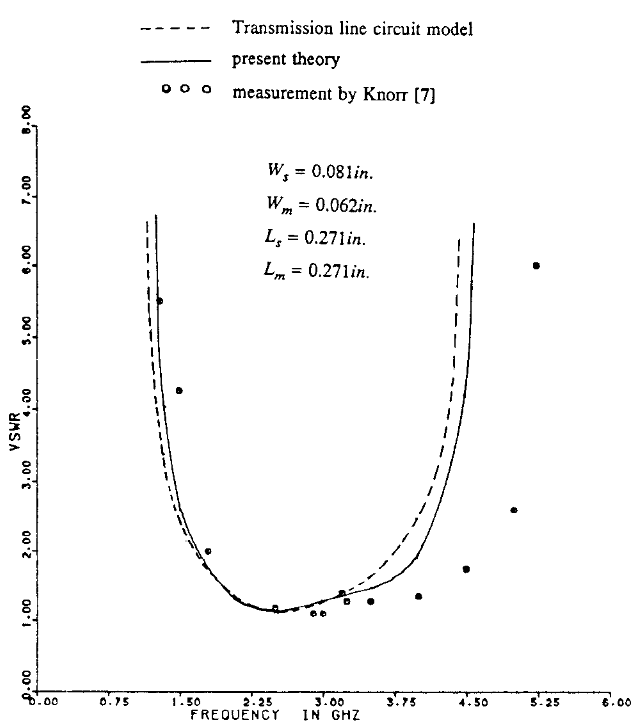

Figure 2. VSWR versus frequency for microstrip-slot transition. $\varepsilon_r = 20$, $d = 0.125$ *in.*

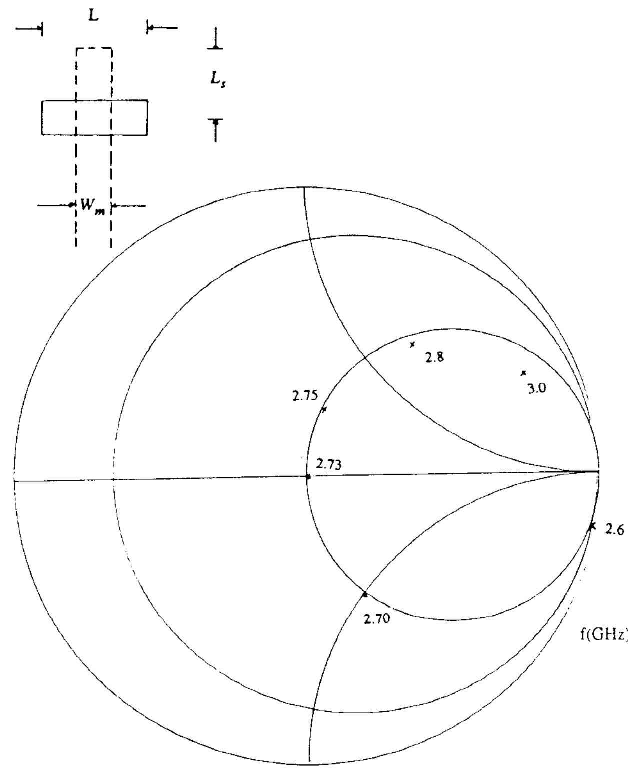

Figure 4. Input admittance of a stub-tuned slot fed by a microstripline. $\varepsilon_r = 2.56$, d = 0.1578 cm, $W_m = 0.44$ cm, L = 4.0 cm, $W_s = 0.6$ cm and $L_s = 3.94$ cm.