Session 10B7

Microstrip Discontinuity Modeling

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Abstract

In this report we develop an efficient modeling technique for microstrip discontinuities. The technique obtains closed form expressions for the lumped element equivalent circuits which are used to model these discontinuities.

To describe our technique, we focus on the bend discontinuity and report the inductance, and capacitance of the equivalent circuit as calculated by an expression developed with this technique.

Introduction

Definition of problem

Due to layout necessities, an electromagnetic wave that propagates down a microstrip line may encounter discontinuities such as bends, tees, and vias. These discontinuities cause disturbances in the electric and magnetic fields which affect the integrity of the signal. It is the prediction and control of the these disturbances that face the Printed Circuit Board (PCB) designer and have made CAD capabilities essential to the engineering process. Engineers must be able to simulate and test their printed circuit designs before major expenses are committed to PCB construction. In order for the CAD process to be an interactive one, the calculation of the circuit characteristics must be relatively fast. Also, the accuracy of these calculations is equally important since realistic simulations are essential.

In general, models for the transmission line effects of microstrip transmissions lines have been both fast and relatively accurate. However, most models of microstrip discontinuities are numerically cumbersome or do not maintain accuracy as frequency increases. Therefore, the heart of the simulation problem is to develop discontinuity models that are numerically efficient and maintain reliability even at relatively high frequencies.

Background

Many papers have addressed the modeling of microstrip discontinuities in terms of lumped element equivalent circuits. In these circuits the disturbances in the electric and magnetic fields are represented as an equivalent capacitance and an equivalent inductance respectively. In the past a quasi-static approach has been used to arrive at values for these equivalent capacitances and inductances [6]; however, there are several disadvantages to this method.

One disadvantage in quasi-static techniques is that extensive computation times are often required to arrive at a value for the desired parameter. As indicated by Anders and Arndt, in their report on the moment method, it can take several minutes per data point to calculate the capacitances and inductances of some discontinuities [1]. Similar computation times are also reported for the matrix inversion method [3]. With these long computation times an interactive design session could not be realized.

Another problem that exists with quasi-static methods is that they cannot take into account the higher order modes, and as a result, they begin to show errors at high frequencies. To solve this problem, many have turned to the planar waveguide model, presented in [4], to assist in the calculations of the scattering (S) parameters, of several discontinuities. These calculated S-parameters have been shown to maintain reasonable accuracy up to approximately 12 GHz.

Approach

Sylvester and Benedek pointed out that the problems of computation time in their technique could be overcome by fitting empirical equations to the curves which the technique generates [2]. The resulting expressions are faster, but their accuracy depends on the accuracy of the original results. If an accepted level of accuracy exists for the original data, a reliable expression can be developed to represent the data. Our approach is to depend on measurements to give us a collection of accurate data and the use of curve-fitting techniques to obtain empirical expressions to represent the data.

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We are focusing this paper on a bend discontinuity where the angle of the bend is allowed to vary from 22.5 to 90 degrees. Although other physical parameters, such as the width of the trace, the dielectric constant, and the thickness of the dielectric, are important to the characteristics of the bend, only the angle of the bend will be allowed to vary. However, the techniques of curve-fitting can be applied to include all of the important physical parameters. The only reason for limiting the number of variables is to simplify the presentation of the technique. As a result, the final closed form equation will be a function of only frequency and the angle of the bend.

Theory

Definitions

Several network parameters are used by this technique. We begin with measured S-parameters and use T-parameters and Z-parameters during the development of the models. All of these parameters represent certain characteristics of a network.

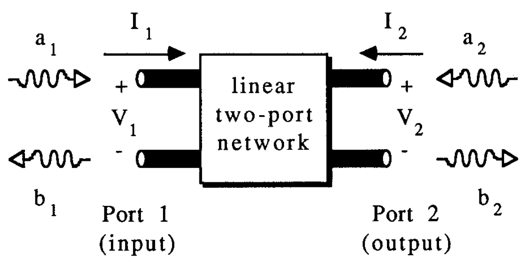


Figure 1: A linear two-port network and its associated input and output quantities.

Figure 1 shows a two-port network with both the input and output parameters displayed. From this network the S-parameters are obtained as follows:

$$b_1 = S_{11}(a_1) + S_{12}(a_2)$$

$$b_2 = S_{21}(a_1) + S_{22}(a_2) \qquad (1)$$

where a and b are the incident and reflected waves of the network. These equations can be represented in matrix form as

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}. \qquad (2)$$

The S-parameters give us a way of representing a network's transmission and reflection coefficients, and they are the only parameters that can be measured at high frequencies. They are, however, non-cascadable. When we need to consider a cascade of S-parameters we must convert them to T-parameters according to the conversion expressions given in [7] and summarized in Table 1. In fact, the T-parameters are sometimes refered to as cascadable S-parameters

Table 1: [S] ¤ [T] Conversions

[S] to [T]	[T] to [S]
$t_{11} = \frac{1}{s_{21}}$	$s_{11} = \frac{t_{21}}{t_{11}}$
$t_{12} = -\frac{s_{22}}{s_{21}}$	$s_{12} = t_{22} - \frac{t_{21}t_{12}}{t_{11}}$
$t_{21} = \frac{s_{11}}{s_{21}}$	$s_{21} = \frac{1}{t_{11}}$
$t_{22} = s_{12} - \frac{s_{11}s_{22}}{s_{21}}$	$s_{22} = -\frac{t_{12}}{t_{11}}$

Table 2: [S], [Z] Conversions

$z'_{11} = \frac{(1+s_{11})(1-s_{22})+s_{12}s_{21}}{\Delta_2}$	$z'_{12} = \frac{2\,s_{12}}{\Delta_2}$
$z'_{22} = \frac{(1-s_{11})(1+s_{22})+s_{12}s_{21}}{\Delta_2}$	$z'_{21} = \frac{2\,s_{21}}{\Delta_2}$
$s_{11} = \frac{(z'_{11}-1)(z'_{22}+1)-z'_{12}z'_{21}}{\Delta_1}$	$s_{12} = \frac{2\,z'_{12}}{\Delta_1}$
$s_{22} = \frac{(z'_{11}+1)(z'_{22}-1)-z'_{12}z'_{21}}{\Delta_1}$	$s_{21} = \frac{2\,z'_{21}}{\Delta_1}$
$\Delta_1 = (z'_{11}+1)(z'_{22}+1) - z'_{12}z'_{21}$	
$\Delta_2 = (1 - s_{11})(1 - s_{22}) - s_{12}s_{21}$	
$z'_{11} = z_{11}/Z_o$; $z'_{12} = z_{12}/Z_o$;	
$z'_{21} = z_{21}/Z_o$; $z'_{22} = z_{22}/Z_o$;	

The Z-parameters represent the impedance characteristics of a network. Their development is similiar to that of the S-parameters and conversions between Z-parameters and S-parameters are accomplished through the use of the conversion expressions of Table 2 [7]. The Z-parameters are represented in matrix form as follows:

$$\begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} \\ Z_{21} & Z_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} \qquad (3)$$

where

$$V_1 = a_1 + b_1$$

$$V_2 = a_2 + b_2 \qquad (4)$$

with I_1 and I_2 being the corresponding currents.

Removing Unwanted Effects by De-embedding and Stripping

Figure 2a shows the coax-to-microstrip test fixture we used while developing our expressions. The test fixture consists of five sub-structures; two connectors, two lengths of line that the connectors are attached to, and a "point discontinuity" where we used the concept of a point discontinuity to describe the perturbations in the fields due to the effects of the discontinuity. Each sub-structure of this fixture has its own characteristics and can be represented by a set of S-parameters (Figure 2b). To isolate the characteristics of the bend, the effects of all the sub-structures must be removed. The de-embedding techniques outlined in [5] can be used to remove the effects of the connectors. We developed a method for removing the effects of the lengths of line that link the connectors to the bend.

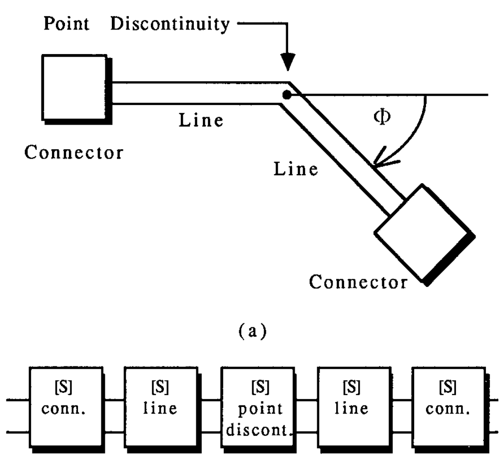


Figure 2 a) Test fixture for the bend discontinuity
b) Sub-structures represented by S-parameters.

After the de-embedding is performed, we strip the remaining structure of the effects of the microstrip lines on either side of the discontinuity, by a method of vector multiplication on a T-parameter network. By de-embedding the test fixture of Figure 2a, we were left with a set of S-parameters that represent the two lengths of line, and the point discontinuity (see Figure 3). From these S-parameters we obtained a cascade of T-parameters according to the conversion expressions summarized in Table 1. These T-parameters are represented mathematically as

$$[T_{tot}] = [T_{11}][T_{pd}][T_{12}] \qquad (6)$$

where $[T_{11}]$ and $[T_{12}]$ are the T matrices of the lines, and $[T_{pd}]$ is the T matrix of the point discontinuity.

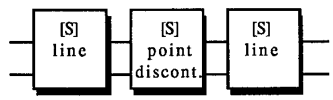


Figure 3: S-parameter representation of test fixture after de-embedding.

In order to obtain the T-parameters of the point discontinuity from Equation 6, we first needed to find the T-parameters of the lines. A test structure similar to that of Figure 2a was used for this purpose. The bend was removed from the structure leaving a straight microstrip line of known length. The S-parameters of the length of line were measured, and de-embedding was performed to remove the effects of the connectors. The result was a set of S-parameters that represented only the line. Again the conversions of Table 1 were used to convert to T-parameters.

The only quantity in Equation 6 remaining unknown to us at this point was the T-parameters of the point discontinuity. Equation 6 was solved for these parameters to give

$$[T_{pd}] = [T_{11}]^{-1}[T_{tot}][T_{12}]^{-1}. \qquad (7)$$

Bend Equivalent Circuit

The equivalent circuit that was used by Gupta [6] for the 90 degree bend is shown in Figure 4a. This circuit can be expanded by using impedance and admittance elements, as shown in Figure 4b, instead of only capacitors and inductors. It is shown in [8] that if the network is assumed to be symmetric, the impedance elements of a two-port tee network can be represented as follows:

$$Z_a = Z_{11} - Z_{12}$$

$$Z_b = Z_{22} - Z_{12} \qquad (8)$$

$$Y_c = 1/Z_{12} \quad .$$

The values of the impedances of Figure 4b can be obtained, according to Equations 8, after the Z-parameter representation of the network is determined. The Z-parameters are obtained from measured S-parameters by using the conversion expressions of Table 2.

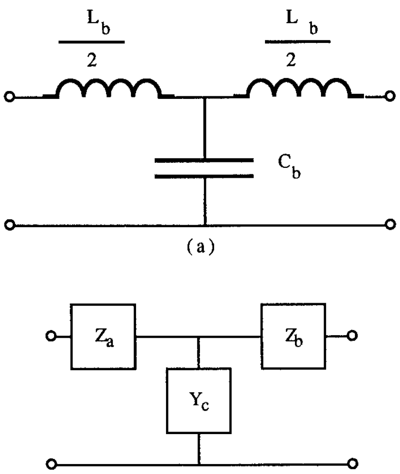


Figure 4 a) Equivalent circuit for a 90 degree bend.
b)Enhanced equivalent circuit.

Curve Fitting Technique

If an accurate set of Z-parameters is available, curve-fitting techniques can be used to arrive at expressions for the impedances of Figure 4b. Many procedures are available for fitting empirical expressions to measured data. Among these are multi-dimensional regression for the development of polynomial fits, and fits which are based on physically justified functionals such as tanh(x). The first allows for interpolation between data points, and functional fits supply a means for obtaining extrapolation capabilities.

In [9] we find many solutions to the problems of curve-fitting. The approach taken by most curve-fitting techniques is to define a function ("merit function" [9]) which will decrease as the agreement between the measured data, and an empirical equation, increases. The goal then is to minimize the merit function. This source, [9], discusses the choice of a merit function, techniques by which the merit function can be minimized, and several approaches for dealing with both linear and non-linear data.

Method

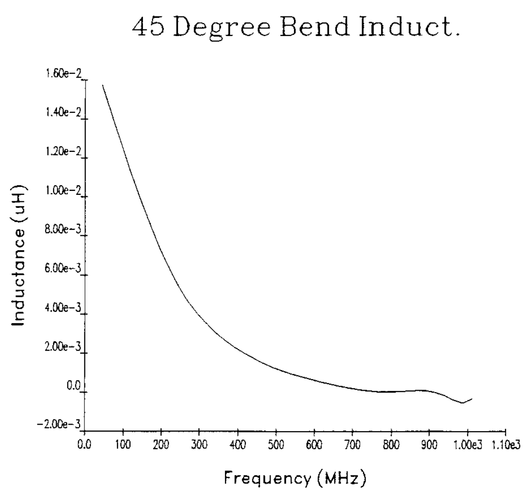
In order to address the angle at which the bend occurs, we considered four structures covering four different bend angles; 22.5, 45, 67.5, and 90 degrees. The S-parameters for each of these structures were measured with a Hewlett-Packard 8510B Automatic Network Analyzer (ANA).

The set of measured S-parameters was then submitted to the de-embedding process where the effects of the connectors were removed. Next, the effects of the lines were removed by the stripping method, according to Equation 7. We were then left with a set of S-parameters that represented the effect of the point discontinuity alone.

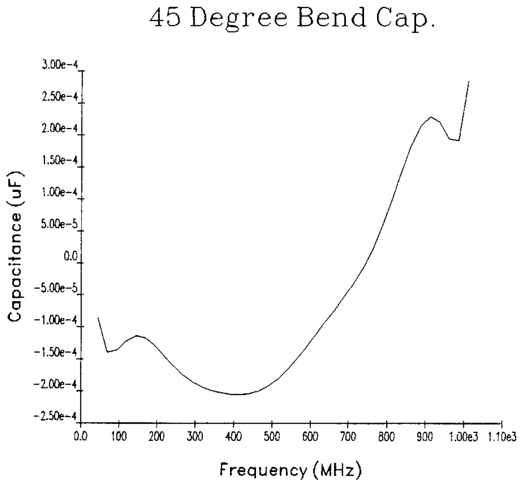
For the purposes of this report the equivalent circuit of Figure 4b was used to represent the bend discontinuity. The S-parameters of the point discontinuity were converted to Z-parameters (Table 2), and these Z-parameters were used to obtain the values of Z_a, Z_b, and Y_c of Figure 4b. The imaginary parts, or reactances, of Z_a, Z_b, and Y_c were then used to obtain values for the inductances and capacitance of Figure 4a. Finally, the curve-fitting techniques were applied to the values of L_b, and C_b, to develop general closed form expressions for both.

Results

We developed a general expression for the lumped elements of Figure 4a. These expressions are a function of frequency and the angle of the bend, and are valid for bends from 22.5 to 90 degrees. Figure 5 shows the values of capacitance and inductance for a 45 degree bend as calculated by the expressions we've developed.



(a)



(b)

Figure 5: a) Calculated Inductance for a 45 degree bend.
b) Calculated Capacitance for a 45 degree bend.

Conclusion

We have described a technique by which fast and accurate expressions can be obtained for the lumped elements of equivalent circuits for microstrip bends. It is based on fitting measured data to empirical expressions. It can be applied to any microstrip discontinuity, and if enough data is available for the curve fitting techniques to properly characterize the discontinuity, all the important parameters can be considered.

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