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EMPIRICAL MODELLING OF A HIGH-SPEED BACKPLANE CONNECTOR

by

Ka Lun Pang, B.Eng.

A thesis submitted to the Faculty of Graduate
Studies and Research in partial fulfilment
of the requirements for the degree of

Master of Engineering

Ottawa Carleton Institute for Electrical Engineering
Department of Electronics
Faculty of Engineering
Carleton University
Ottawa, Ontario
Sept, 1991

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submitted by Ka Lun Pang, B.Eng.
in partial fulfilment of the requirements for
the degree of Master of Engineering.

Prof. B.A. Syrett
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October, 1991

ABSTRACT

This thesis presents an empirical approach to the modelling of a high speed backplane connector. The connector chosen for study belongs to the multi-pin press-fit family of connectors (Teradyne HD+) designed for high speed signals. Modelling is performed in both the time and frequency domains using measured reflection and transmission data. Both the transmission/reflection and crosstalk models developed provide satisfactory results in terms of S-parameters in the frequency domain and pulse response in the time domain. Although it does not provide an accurate result with respect to the rise-time performance, the transmission/reflection model does follow the general trend of the measured data. However, the crosstalk model shows very close approximation of the near-end and far-end crosstalk levels. The fact that the crosstalk model consists of the combination two transmission/reflection models demonstrates the validity of the transmission/reflection model.

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To

My Parents

CHAPTER 1

INTRODUCTION

1.1 Motivation

Due to the rapid development of semiconductor devices, the speed of integrated circuits is steadily increasing. As far as a microprocessor-based system is concerned, a higher clock rate means that the signals have a wider bandwidth. As a result, the performance of many digital systems is no longer limited by the speed of individual components but by the interconnects. In other words, the bottle-neck of the whole system becomes the bus connectors instead of the operating speed limits of the components. Recently, various studies have been conducted on high-speed signal propagation. An example is in [1] which deals with lossy transmission lines.

The primary motives of this work were to understand the limitations of electrical bus connectors at high frequencies, and to predict the effects of such parameters as geometry, materials and shielding on the performance of these connectors.

1.2 Objectives

The objective of this thesis was to characterize and model a typical high-speed backplane connector empirically. Measurements and modelling were conducted in both the time and frequency domains. A requirement for the connector model was that it should be based on the connector's physical geometry. The goal was to match the performance of a connector model of a particular grounding pattern and geometry to its corresponding measurement. The model was then modified for the variations in the pattern and geometry for the rest of the measurements. The final goal was to be able to obtain the crosstalk model of the connector by combining two single line transmission models with minimum modifications in the model parameters.

1.3 Outline

In Chapter 2, a general overview of high speed connectors is presented along with an introduction to the particular type of connector chosen in this thesis (Teradyne HD+ series). Some aspects of two of the most frequently used numerical methods for electromagnetic field analysis are also discussed in this chapter. This is followed by the introduction of the Finite Element Method software package from the Lawrence Livermore National Laboratory in Chapter 3. Also contained in Chapter 3 is the application of this software to determine the characteristic impedance of a section of the

connector as one of the parameters of the model. Chapter 4 contains a description of the actual measurement technique and modelling of the connector. The construction of the test jig for measurements, the de-embedding process and discussion of the computer-aided testing software ANACAT are included in Chapter 4 as well. The developed models are compared with measured results in Chapter 5. Finally, conclusions are drawn in Chapter 6.

CHAPTER 2

CHARACTERIZATION AND MODELLING

OF HIGH SPEED INTERCONNECTS

2.1 High Speed Interconnects

In the past, when the speed of operation of microprocessor- based systems was relatively low, a connection from a signal line on a motherboard to a signal line on a daughterboard through a backplane connector was treated as a short circuit. This was a fairly safe assumption until advances in semi-conductor technology led to the dramatic increase in the speed of integrated circuits. Higher speed is associated with high clock rates and short rise-times. Invoking the Fourier theorem, a short rise-time waveform contains more high-frequency components with greater amplitude than a longer rise-time waveform. In the frequency domain, each frequency component can be considered as independent and separate. When the highest frequency of interest attains a value such that the wavelength becomes comparable to the connector dimensions, travelling wave voltages must be considered instead of terminal voltages.

Today, transmission line theory is employed for high-speed connectors. Problems such as reflections, ringing, degradation in rise-time, propagation delay and crosstalk are

always considered. As a matter of fact, techniques like shielding and matching that were once used solely for high-frequency analog circuits are now being widely employed in the design of high-speed computer systems.

There are two basic characteristics that are of interest when dealing with high-speed connectors: the signal transmission and crosstalk. Transmission performance deals with the desired signal transfer in terms of degradation in rise-time, presence and duration of ringing, and propagation delay in the time domain and reflection and cutoff frequency in the frequency domain. To model a single transmission line network, elements such as capacitors, inductors and transmission lines are usually chosen. Crosstalk performance on the other hand focuses mainly on the level of unwanted signal induced on neighbouring signal lines. Crosstalk models usually include extra elements such as coupled inductors and coupled transmission lines.

Crosstalk is one of the most important issues for microprocessor-based systems due to the density of interconnects. A signal line may be triggered falsely if the crosstalk level is high. For a pair of transmission lines, i.e., a four-port network, crosstalk is usually studied under two categories: near-end crosstalk and far-end crosstalk. Referring to Fig.2.1, near-end crosstalk is the measure of unwanted energy coupled from port 1 to port 3, and far-end crosstalk is the coupling from port 1 to port 4.

Fig. 2.1 Near-end and far-end crosstalk

A common way to suppress crosstalk is through shielding. Shielding is usually done by placing conducting plates connected to ground close to the conductor pins in a connector module or by grounding some conductors in the vicinity of the signal lines. A patterning scheme is often used to specify the grounding of connectors. For example, a 25% ground pattern means a quarter of the total conductor pins of the connector module are grounded in a roughly uniform distribution pattern. This method obviously reduces the number of useful signal lines available and the percentage of grounding is always a trade-off between the crosstalk level and the signal line density of the system.

It should be pointed out that although it was not studied in this thesis, "ground bounce" is another problem that high speed systems often encounter. Ground lines or planes in a high speed system are always connected to a primary ground reference and due to differences in lengths of these connections, different inductances will appear between each line/plane and the primary reference. The result is the ground is not well defined on these lines/planes and short rise-time pulses will cause ringing at the output.

2.1.1 Teradyne HD+ Interconnection System

In this thesis, the Teradyne HD+ interconnection system (trademark of Teradyne Inc) is studied. The Teradyne High Density system (HD+) is a family of modular backplane and daughterboard connectors. Standard HD+ modules contain four rows of contacts on a continuous 0.100" x 0.100" grid. The shortest contacts are in ROW A; the longest contacts are in ROW D. The backplane connector inserts into plated-through holes in the motherboard with compliant press-fit contacts of length 0.125" or 0.685". The daughterboard connector pins are soldered to plated-through holes in the daughterboard. Interconnection is achieved by insertion of the backplane connector contact pins into a spring clip contact in the daughterboard connector.

In order to reduce the level of crosstalk, there are two extra types of modules available: the HD+1 and HD+2. The +1 and +2 indicate the number of extra ground

contact rows in the module. Connected to the ground contacts are sections of copper sheet which provides shielding and low-inductance ground in close proximity to the signal pins. Fig. 2.2 shows the construction of the HD+ family and Fig. 2.3 is the cross-section of the HD+2 module.

Fig. 2.2 Construction of the Teradyne Connectors [from Teradyne High Density Plus Brochure]

- (a) Motherboard connector
- (b) Daughterboard connector

High Density Plus Two Cross Section

Fig. 2.3 Cross-section of the Teradyne HD+2 connectors.

2.2 Characterization

Characterization of multi-port networks can be done in either the time domain or the frequency domain. These measurements are related by the Fourier transform. In other words, one can obtain the frequency response by taking the Fourier Transform of the time domain measurement and vice versa. In this project, both kinds of measurements are taken in order to ensure accuracy.

2.2.1 Characterization in the Frequency Domain

The scattering parameters (or S-parameters) were used in the frequency domain characterization process. S-parameters are ideal for characterizing networks at high frequencies since measurements can be performed easily, requiring only matched terminations. In the high frequency (GHz) range, open- and short circuits that are needed in ABCD- or Y- matrix measurements are very difficult to realize and may cause oscillations of an active device under test (DUT).

S-parameters also show the reflection and transmission characteristics of the measured network as a function of frequency. This property is very useful in the modelling process of the connector since resonances and frequency-dependent losses can easily be identified.

The high-frequency CAD package TOUCHSTONE (trademark of EEsof Inc.) was used to model the connectors. TOUCHSTONE is a circuit simulator in the frequency domain that can be used for circuit analysis and optimization. Elements used in microwave or RF circuits including microstrip, stripline and waveguide elements are all available in TOUCHSTONE. For modelling purposes, the S-parameter measurements can be included in a TOUCHSTONE circuit file so that the performance of a circuit model can be compared with the measurements on one display format directly. The available display formats are the Smith chart, polar display, x-y grid and tabled data. One feature in TOUCHSTONE that is very useful in matching the characteristics of the circuit model and the measurements is called tuning. This feature allows interactive alterations of one of the model's elements and displays the result without requiring re-simulation of the entire circuit. The difficulties in modelling are greatly reduced by this single feature.

2.2.2 Characterization in the Time Domain

Time domain reflectometry (TDR) is employed for time domain characterization. In the time domain, the main concerns are the degradation in rise-time of the signal pulses, and the presence of ringing and overshoots in the signal waveform which are caused by low pass filtering, reflections and mismatches.

TDR is a special form of pulse test in which the signal viewed on the oscilloscope

display of the TDR set-up is the time series of reflections produced by a DUT, as for instance, by imperfections in a transmission line. The basic elements of a TDR measurement system are a fast-rise-time ($\approx 50\text{ps}$) generator emitting a voltage step synchronized with the start of the oscilloscope, and wideband detectors to sample the incident and reflected voltage waveforms. As the pulse travels down the DUT, reflections will occur wherever changes in characteristic impedance take place or wherever discontinuities are present. The reflected wave is recorded by a sampling oscilloscope. Since different imperfections or discontinuities have different forms of reflections, their locations and natures can easily be obtained. Besides reflection, the rise-time of a voltage step can also be measured before and after the DUT so the deterioration of rise-time of the applied step on transmission through the DUT can be identified.

Time domain measurements follow closer to intuition than the frequency domain analysis since the physical locations of discontinuities in the DUT can be determined by knowing the propagation velocity of the wave or by experimental probing of the DUT. One can easily determine if the measurement is a legitimate one or not by relating the measurement to the physical structure of the DUT.

2.3 Numerical Methods in Electromagnetics

The rapid development of digital computer processing speed and memory capabilities has greatly enhanced the ability to solve electromagnetic field problems using numerical techniques. In the past, many field problems were solved by analytical methods. However, most problems cannot be solved analytically without making simplifying assumptions. Although numerical methods were also available, the solutions to complicated problems were often prohibited by the memory size of computers. Also the popularity was limited by the speed of the microprocessors since many iterations are always required in numerical methods for accurate solution.

Currently, it is more convenient to use computer time to reduce the analytical effort and to solve problems intractable to analysis. The accuracy can be improved by increasing the number of iterations (while still limited by round-off errors). As a matter of fact for many problems, a common personal computer can provide the accuracy which is compatible to the analytical solution in a reasonably short period of time.

The computer-aided engineering (CAE) packages now available provide electrical engineers with new insights into old problems by providing visualization of the electric and magnetic fields. By using these packages, engineers can analyze and "see" the electromagnetic properties of a product without having to make measurements on a physical prototype. The time and cost for the development of products can be reduced

significantly.

Perhaps the most encouraging progress made in recent years in CAE tools for electromagnetics is the graphic interface. In the past, input and output data of numerical methods were mostly in the form of tables of numbers. Experience and imagination were required to analyze a problem correctly. With the help of graphic interfaces and advances in algorithms, one needs only to define the geometry of the problem and the software will do the required pre-processing and subsequent calculation automatically. The final result can be displayed on a graphic monitor. For example, a coaxial transmission line to rectangular waveguide adapter was recently analyzed and the electric and magnetic fields were displayed on a colour graphic monitor in three dimensions [2].

The most common numerical methods currently used are the method of moments (MOM) and finite element method (FEM). They are described briefly in the following sections.

2.3.1 Finite Element Method (FEM)

The development of FEM was initially motivated by mechanical and civil engineers' need to analyze heat transfer and stress problems. Since there is an analogy between the equations involved in electromagnetic field analysis and in heat transfer/stress

analysis, FEM has now become an invaluable tool for electrical engineers for various applications. For example, it can be used to find the capacitance of a microstrip line, the distribution of the magnetic fields around the armature of a motor, the characteristic impedance of wave guides, etc.

The approach used in the FEM is to create an approximate expression for the stored energy $W(u)$ in a region (x,y,z) associated with the unknown potential $u(x,y,z)$ by assuming the potential u to be given by a combination of suitably chosen, simple functions with as yet undetermined coefficients. Since a static or equilibrium system always adjusts itself so that the energy content is at its lowest level, minimization of the energy then determines the coefficients, and thereby implicitly determines an approximation to the potential distribution. The following is an example for the two dimensional static field problem.

2.3.1.1 Two Dimensional Static Field Problems

Static field problems always involve solution of the Laplace equation:

$$\nabla^2 u = 0 \quad (2.1)$$

An example is the TEM mode analysis of a coaxial cable with rectangular conductors in which the electric potential distribution in the interconductor space is sought

(see Fig. 2.4). Due to symmetry, only one-quarter of the actual problem needs to be considered. Two kinds of boundary conditions are involved in this static field analysis: Dirichlet (forced) conditions on the conductor surfaces, and Neumann (natural) boundary conditions along the planes of symmetry. These boundary conditions, along with Eqn. (2.1), provide a unique solution to the problem. The minimum-energy principle[3] states that the potential distribution of a system would adjust itself so that the potential energy is minimized. For the rectangular coaxial transmission line, this energy, to within a constant multiplier, is given by [3]

$$W(u) = \frac{1}{2} \int_S |\nabla u|^2 dS \quad (2.2)$$

where the integral is performed over the whole interconductor region, S . Eqn. (2.2) is then minimized in order to find the potential distribution $u(x,y)$.

Fig. 2.4 One quarter of the rectangular coaxial line.

2.3.1.2 First-Order Elements

As the name "Finite Element Method" implies, the solution region is subdivided into a number of elements. In this example, the triangular element is used. However, rectangular elements are also common.

Fig. 2.5 Typical finite element in two dimension.

The first step of FEM is to formulate the basic elements in a standard fashion and then to interrelate all of them so that a continuous potential distribution can be obtained. Fig. 2.5 shows a typical triangular element. Assume that the potential within this element can be approximated by the expression

$$U = a + bx + cy \quad (2.3)$$

Eqn. (2.3) provides a first order piecewise-planar function for the potential distribution and it is continuous everywhere in the solution region. This can be seen by

the fact that the potential along any edge of the triangular element is the linear interpolation between its two vertex values. Higher-order polynomials can also be used but the complexities involved would soon become not viable. Existing FEM software packages seldom use elements higher than second-order.

In addition to Eqn. (2.3), letting the potential at the vertices of the element be U_1 , U_2 and U_3 leads to

$$\begin{bmatrix} U_1 \\ U_2 \\ U_3 \end{bmatrix} = \begin{bmatrix} 1 & x_1 & y_1 \\ 1 & x_2 & y_2 \\ 1 & x_3 & y_3 \end{bmatrix} \begin{bmatrix} a \\ b \\ c \end{bmatrix} \quad (2.4)$$

where the x 's and y 's are the co-ordinates of the vertices. Solving Eqn. (2.4) and substituting into Eqn. (2.3) yields

$$U = [1 \ x \ y] \begin{bmatrix} 1 & x_1 & y_1 \\ 1 & x_2 & y_2 \\ 1 & x_3 & y_3 \end{bmatrix}^{-1} \begin{bmatrix} U_1 \\ U_2 \\ U_3 \end{bmatrix} \quad (2.5)$$

The above equation can be written as

$$U = \sum_{i=1}^3 U_i \alpha_i(x, y) \quad (2.6)$$

where

$$\alpha_1 = \frac{1}{2A} \{(x_2 y_3 - x_3 y_2) + (y_2 - y_3)x + (x_3 - x_2)y\} \quad (2.7)$$

is a linear function of co-ordinates only. A is the area of the triangular element on the x - y plane. The other two functions α_2 and α_3 can be obtained by cyclic interchanges of the subscripts in Eqn. (2.7).

Eqn. (2.2) can now be used to find the energy within one element. Taking the potential gradient of Eqn. (2.6)

$$\nabla U = \sum_{i=1}^3 U_i \nabla \alpha_i \quad (2.8)$$

so the element energy becomes

$$W^{(e)} = \frac{1}{2} \int |\nabla U|^2 dS \quad (2.9)$$

or, from Eqn. (2.8),

$$W^{(e)} = \frac{1}{2} \sum_{i=1}^3 \sum_{j=1}^3 U_i \int \nabla \alpha_i \cdot \nabla \alpha_j dS U_j \quad (2.10)$$

For the sake of clarity, if we define a matrix S which has elements

$$S_{ij}^{(e)} = \int \nabla \alpha_i \cdot \nabla \alpha_j dS, \quad (2.11)$$

then Eqn. (2.10) can be written as the matrix quadratic form

$$W^{(e)} = \frac{1}{2} U^T S^{(e)} U \quad (2.12)$$

where U is the column vector of vertex values of potential and the superscript T denotes the transpose of the matrix.

For any given triangle, S can easily be found by substituting Eqn. (2.7) into Eqn. (2.11).

2.3.1.3 Element Assembly and The Final Solution

After the energy of each single element is found, the total energy is found by summing over all the elements in the problem region. In order to assemble the individual elements, a matrix C is defined to reallocate the potentials of disjoint elements to the conjoint set of elements:

$$U_{dis} = C U_{con} \quad (2.13)$$

Details of the element assembly will not be discussed in this work but can be found in [3]. The energy of the conjoint system can now be written as

$$W = \frac{1}{2} U_{con}^T S U_{con} \quad (2.14)$$

where $\mathbf{S}$ is the assembled coefficient matrix of the connected problem.

As the final step, the energy of the whole system is minimized, i.e.,

$$\frac{\partial W}{\partial U_k} = 0 \quad (2.15)$$

where the index k refers to entries in the connected potential vector or the node numbers in the connected model.

The minimization in Eqn. (2.15) would have a trivial solution with all potentials equal to zero if no boundary conditions are stated. In the coaxial transmission line case, both the conductors have fixed potentials and this is reflected in the $\mathbf{S}$ and $\mathbf{U}$ matrices in which there exist subsets of S_{ij} and U_{ij} which correspond to the fixed potentials. Using the subscripts f and p for the free and prescribed potentials, respectively, Eqn. (2.15) can be written with the matrices in partitioned form,

$$\frac{\partial W}{\partial U_k} = \frac{\partial}{\partial [U_f]_k} [U_f^T U_p^T] \begin{bmatrix} S_{ff} & S_{fp} \\ S_{pf} & S_{pp} \end{bmatrix} \begin{bmatrix} U_f \\ U_p \end{bmatrix} = 0 \quad (2.16)$$

Note that only differentiation with respect to the free nodes is taken since taking the differentiation with respect to prescribed potentials is meaningless. The result of the above equation is

$$[S_f \ S_p] \begin{bmatrix} U_f \\ U_p \end{bmatrix} = 0 \quad (2.17)$$

or,

$$U = \begin{bmatrix} -S_f^{-1} S_p U_p \\ U_p \end{bmatrix} \quad (2.18)$$

Although the final result of the potential for every point is in the form of a column vector, the solution is actually defined everywhere in the problem region according to Eqn. (2.3).

Since the principle of FEM is relatively simple and the mathematics involved are not intensive, FEM is being employed in a number of commercial electromagnetic software packages [2]. One drawback for FEM is that definition of the mesh is often tedious. As one would imagine, this step requires a substantial amount of effort for complicated three-dimensional problems. Automatic mesh generation is currently a popular topic of research in FEM [4,5] and some software packages have already included this feature.

2.3.2 Method of Moments (MOM)

The method of moments is a method of solving linear integrodifferential equations [6]. Consider the inhomogeneous equation

$$L(f) = g \quad (2.19)$$

where L is a linear operator, g could be the source or excitation of the system and is known, and f may be the field or response which is to be determined. The first step in MOM is expansion, in which f is represented by a series of pre-determined expansion functions f_i in the domain of L , as

$$f = \sum_n \alpha_n f_n \quad (2.20)$$

where the α_n are unknown constants associated with each f_n . The number n is arbitrary but for an exact solution, an infinite sum is usually required. Since L is a linear operator, putting Eqn. (2.20) into Eqn. (2.19) yields

$$\sum_n \alpha_n L(f_n) = g \quad (2.21)$$

The second step is called testing. Testing requires a set of functions called testing functions, or weighting functions. It is assumed that a suitable inner product $\langle f, g \rangle$ has been determined for the problem and the testing functions $w_1, w_2, \dots$ are in the range of

L. Taking the inner product of Eqn. (2.20) with each of the testing functions,

$$\sum_n \alpha_n \langle w_m, Lf_n \rangle = \langle w_m, g \rangle \quad (2.22)$$

The above equation can be written in matrix form as

$$[l_{mn}][\alpha_n] = [g_m] \quad (2.23)$$

where

$$[l_{mn}] = \begin{bmatrix} \langle w_1, Lf_1 \rangle & \langle w_1, Lf_2 \rangle & \dots \\ \langle w_2, Lf_1 \rangle & \langle w_2, Lf_2 \rangle & \dots \\ \dots & \dots & \dots \end{bmatrix} \quad (2.24)$$

$$[\alpha_n] = \begin{bmatrix} \alpha_1 \\ \alpha_2 \\ \cdot \\ \cdot \\ \cdot \end{bmatrix} \quad [g_m] = \begin{bmatrix} \langle w_1, g \rangle \\ \langle w_2, g \rangle \\ \cdot \\ \cdot \\ \cdot \end{bmatrix} \quad (2.25)$$

If the matrix $[l]$ is non-singular and its inverse exists, α_n are determined by

$$[\alpha_n] = [l_{mn}^{-1}][g_m] \quad (2.26)$$

Then the solution for f can be obtained by the substitution of the α_n into Eqn. (2.20). The solution may be exact or approximate, depending upon the choice of the f_n and w_n . If the same function is used for expansion and testing, it is called the Galerkin's method. Two

other methods frequently used are the point-matching method and the method of subsections. Point-matching employs the Dirac delta functions as testing functions and the method of subsections involves the use of basis functions f_n each of which exists only over subsections of the domain of f . Both methods are designed to simplify the evaluation of the matrix $[I]$ and they are sometimes used together in one problem.

To relate the method of moments to the coaxial cable problem in the previous section: the linear operator is the inverse of the Laplacian operator; the unknown function is the charge distribution on the conductors' boundaries; the known function is the set of potentials at the boundaries. In MOM, the solution is the charge distribution at the boundaries whereas in FEM it is the potential distribution.

2.3.3 Comparison of the FEM and MOM

The most obvious difference between the FEM and MOM is the definition of elements. FEM requires elements in the entire problem region while the MOM has the elements on the fixed potential boundaries and interfaces of different dielectrics only. The number of elements in MOM is usually far less than FEM since the dimension of the elements is always one less than the dimension of the problem. In other words, MOM requires one dimensional elements in 2-D problems, and 2-D elements in 3-D problems.

In general, FEM is more flexible once the code is written for a particular problem. For example the static field case, any changes in the geometry do not change the formulation; only a re-definition of the finite element mesh is needed. On the other hand, MOM often calls for different "*Green's functions*"¹ for variations of a particular problem. A number of publications have studied different cases of multi-dielectric, multi-conductor, planar transmission lines using the MOM [7,8,9]. Although the effort of finding the suitable Green's function for a particular problem can sometimes be great, once it is found, the computation is fairly straight forward. Since the number of unknown in the MOM is smaller than the FEM, computation is usually faster. Thus there is a trade off between speed, formulation effort and flexibility for the choice of which method to use. It should also be mentioned that for unbounded problems, MOM used to have an advantage since no elements are needed to fill up the infinite space while the FEM always required an approximation of the field at some distance far away from the region of interest. However, a modification of the FEM called the *infinite element* was introduced to overcome this weak point of FEM [10]. Infinite elements were put on the edge of the finite element mesh to simulate the unbounded effect and satisfactory results were reported.

In addition, a hybrid method that combines both FEM and MOM was also reported for antenna problems[11]. In this method, the solution region is divided into two

¹Green's function is the kernel of the inverse of the operator L in Eqn. 2.19. Here, it represents the relationship of charge and potential on conductors with fixed potential.

regions. The region that includes the source of radiation is approximated by finite elements and a transition is made at the boundary of the two regions. Outside the finite element region, there are no elements but the effect of radiation into infinite space is taken into account by the elements at the boundaries.

CHAPTER 3

CHARACTERISTIC IMPEDANCE FROM NUMERICAL METHODS

In this project, the characteristic impedances of transmission line components in the model for the Teradyne HD+ connector were determined by a FEM software package and the odd- and even- mode impedances of the coupled transmission line were obtained by a software package designed primarily for transmission line problems using MOM [12]. The following is a brief introduction to the FEM package.

3.1 FEM Software from LLNL

FEM software capable of solving two-dimensional and three-dimensional problems was obtained from the Lawrence Livermore National Laboratory (LLNL). These packages are not commercial products and are still under development. Each package is divided into three parts. The first part is the pre-processing, the second part is the system equation solver and the last part is the post-processor. The two-dimensional package is composed of MAZE, TOPAZ2D and ORION. The three-dimensional package is composed of INGRID, TOPAZ3D and TAURUS. The UNIX version of these software packages has been used, which can be run in the X-window environment.

3.1.1 MAZE

MAZE is an interactive program that has been developed as an input generator for TOPAZ2D and some other finite element codes at LLNL.

MAZE consists of three stages. In the first stage, lines and parts are defined. Elements are defined at the same time as the parts. The user has control over the number of elements in each part. The shape of elements in MAZE is rectangular. In order to improve the efficiency on element counts, a feature called variable zoning is available. Variable zoning means that one can have fine elements close to the boundaries and coarse elements far away from the boundary. The first stage ends with auto-numbering of the elements.

The second stage is the boundary condition definition. By means of the zooming feature of MAZE, the boundary nodes can be read directly from the graphic display and definitions can be made interactively. However, since the boundary conditions are defined by node numbers, any changes made to the mesh means that the definitions have to be altered since the numbers at the boundaries would be changed.

The third phase is the definition of material properties such as the dielectric constant and polarization. Upon termination of MAZE, an ASCII output file is created with sufficient information for the system equation solver TOPAZ2D.

Two examples will illustrate mesh definition. In Fig. 3.1, the finite element mesh for the two coupled rectangular lines is shown[4]. The bottom boundary at $y=0$ is the ground plane. An interface of dielectric of the lower layer and air of the top layer is located at $y = 5$. The boundaries at $y = 25$, $x = -10$ and $x = 50$ satisfy the natural boundary condition. The regions of greatest interest (the most dense regions) are the surface of the lines since integration was performed in order to find the capacitance matrix.

Fig. 3.2 shows the mesh definition for a truncated cross-section of the Teradyne connector. The rectangular areas having areas of 10×40 represents the cross-sections of the conductor pins. The highest density region centers at $(50, 250)$ is the signal pin. The hollow ones are the ground pins which a fixed potential of 0 can be imposed on their boundaries. The pins that have subdivisions inside are floating lines. The subdivisions are actually material which has a high electrical conductivity. The top and bottom plate at $y = -50$ and $y = 450$ are ground plates where the left and right boundaries at $x = -300$ and $x = 400$ are the natural boundary conditions.

The mesh in both figures were designed in such a way that the number of long and narrow elements are at the minimum for stability of the result.

Fig. 3.1 Finite element mesh of the cross-section of two coupled transmission lines.

Fig. 3.2 Finite element mesh of the Teradyne connector.

3.1.2 TOPAZ2D

TOPAZ2D was originally developed for two-dimensional heat transfer analysis. However, the system of equations involved is analogous to the system of equations for electrostatic and magnetostatic problems. Thus TOPAZ2D was modified for electrostatic field analysis. It either reads the output file from MAZE which defines the element coordinates, material properties, etc. or the user can define the problem in TOPAZ2D as the first stage. The outputs of TOPAZ2D are the potential and flux distributions in the electrostatic or magnetostatic cases. TOPAZ2D also produces a "plot file" for the post-processor ORION. Some aspects of TOPAZ2D are listed in Appendix A.

3.1.3 ORION

ORION is a 2-D interactive colour post-processor. It takes the output file TOPAZ2D and gives plots of potential and flux. For the potential distribution, ORION plots constant potential lines while it gives constant strength vector plots with directions for the vector plots. Some examples of ORION are shown in section 4.2.1.

3.2 C- and L-matrices and Characteristic Impedance

Numerical methods in electrical engineering usually give the potential, flux, or charge distribution in the final solution. In order to obtain more meaningful and useful information, post-processing is always needed. Among the most common interests are the C-matrix, the L-matrix and the characteristic impedance for a system of two or more conductors, ie. a transmission line.

At low to medium frequencies, static field analysis is sufficient to determine the C- and L-matrix of the transmission line.

The output of TOPAZ2D provides the potentials at every node of the mesh and according to Gauss' law:

$$\int_{surf.} \vec{D} \cdot d\vec{s} = Q_{encl} \quad (3.1)$$

$$and, \quad \vec{D} = \epsilon \vec{E} = -\epsilon \nabla \phi \quad (3.2)$$

$$so, \quad Q_{encl} = -\epsilon \int_{surf.} \nabla \phi \cdot d\vec{s} \quad (3.3)$$

where $\vec{D}$ = electric flux density

$\vec{E}$ = electric field intensity

ϵ = permittivity

ϕ = potential

Q_{encl} = charge enclosed

Eqn. (3.3) reveals that the charge enclosed by a Gaussian surface can be determined if the potential distribution is known. A brief discussion is given in the following section.

3.2.1 Finding the Charge Enclosed Inside a Gaussian Surface

As an example, let us look at an infinitely long conductor which has a rectangular cross-section, as shown in Fig. 3.3.

The finite element mesh was so designed that a fine mesh is present in the vicinity of the boundaries. To find the charge enclosed, one can put a Gaussian surface between the closest nodes and the conductor surface.

As shown in the Fig. 3.3, the normal vector $d\vec{s}$ is always parallel to the x- or y-axis. In order to find the charge enclosed, the difference in potential between each node

Fig. 3.3 cross-section of a conductor.

and the conductor is taken and multiplied by the width of the associated division and the dielectric constant. These values are then summed over the Gaussian surface.

A Fortran program CHARGE (Appendix B) was written to calculate the enclosed charge. The program requires the node number at the corners of the cross-section in Fig. 3.3 and a few geometrical parameters. Then the output file from TOPAZ2D is read in which contains the potential distribution. The output of this program is the enclosed charge and the capacitance per unit length.

3.2.2 Finding the C-matrix

As mentioned previously, the capacitance matrix or C-matrix is an important property of the multi-conductor system. In the single conductor case, it degenerates to the capacitance per unit length of the transmission line. The C-matrix is used to find the characteristic impedance for the single conductor system and the even- and odd- mode impedances for coupled conductor systems in this project.

For a single conductor configuration, the capacitance per unit length can be found using the fundamental relationship between charge, voltage and capacitance, i.e.,

$$Q = CV \quad (3.4)$$

Since V is defined on the surface of the conductor and Q can be found according to Eqn. (3.3), C can be obtained without any difficulties.

In a multi-conductor system, the relationship between charges and voltages on each conductor as given by [13] is

$$\begin{bmatrix} Q_1 \\ Q_2 \\ \vdots \\ Q_N \end{bmatrix} = \begin{bmatrix} C_{11} & C_{12} & \cdots & C_{1N} \\ C_{21} & C_{22} & \cdots & C_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ C_{N1} & C_{N2} & \cdots & C_{NN} \end{bmatrix} \begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_N \end{bmatrix} \quad (3.5)$$

The diagonal elements of $[C_{ij}]$ are called coefficients of capacitance or self capacitances. The off-diagonal elements are called the coefficients of induction or mutual capacitances. To find the $[C_{ij}]$ matrix, a fixed potential is placed on each conductor in turn while the rest are grounded. For example,

$$Q_1 = C_{11} V_1 \bigg|_{V_2, \dots, V_N = 0} \quad (3.6)$$

$$\text{so,} \quad C_{11} = \frac{Q_1}{V_1} \bigg|_{V_2, \dots, V_N = 0} \quad (3.7)$$

and it is obvious that

$$C_{12} = \frac{Q_1}{V_2} \bigg|_{V_1, V_3, \dots, V_N = 0} \quad (3.8)$$

The C-matrix also has the following properties [13],

$$C_{ij} = C_{ji} \quad i, j = 1, 2, \dots, N \quad (3.9)$$

$$C_{ii} > 0 \quad i = 1, 2, \dots, N \quad (3.10)$$

$$C_{ij} < 0 \quad i \neq j = 1, 2, \dots, N \quad (3.11)$$

$$\sum_{j=1}^N C_{ij} > 0 \quad i = 1, 2, \dots, N \quad (3.12)$$

3.2.3 The L-matrix

Once the C-matrix is found, the inductance matrix or L-matrix can be obtained by the relationship given by [13]

$$L = \mu_0 \epsilon_0 [C_0]^{-1} \quad (3.13)$$

where C_0 is the capacitance matrix which would result if the transmission line were surrounded by free space. This method assumes that the mode of propagation is pure transverse electromagnetic (TEM).

3.2.4 Characteristic Impedance

Characteristic impedance is often used in high-frequency circuits when knowledge of reflections, matching and power transfer is required. Thus it is of great interest to find the characteristic impedance by a numerical method. In this project, both single and coupled transmission lines are studied and methods of finding the characteristic impedances for these two systems are reviewed.

For a single transmission line in a homogeneous medium, the characteristic impedance Z_o is given by

$$Z_o = \frac{1}{vC} \quad (3.14)$$

where v is the velocity of TEM-mode wave propagation in the medium surrounding the line and is equal to the speed of light c when the transmission line is placed in free space. C is the capacitance per unit length of the transmission line. The above equation is valid for a homogeneous medium only and cannot be used, for example, in microstrip problems. For non-homogeneous media, the characteristic impedance can be found by

$$Z_o = \frac{1}{c\sqrt{C_a C_d}} \quad (3.15)$$

where C_d is the capacitance value obtained with the dielectric in place and C_a is the

capacitance with the dielectric medium replaced by free space.

In the case of coupled lines which involve only 2 lines, there are two modes of operation: the even and odd modes. With a ground reference established, the even mode corresponds to $V_1 = V_2$, where V_1 and V_2 are the voltages on line 1 and line 2, respectively. The odd mode, on the other hand, corresponds to $V_1 = -V_2$.

Hence there exists two characteristic impedances, Z_{oe} and Z_{oo} which are given by [13] as the even-mode impedance

$$Z_{oe} = \frac{1}{v (C_{11} + C_{12})} \quad (3.16)$$

and the odd-mode impedance

$$Z_{oo} = \frac{1}{v |C_{11} - C_{12}|} \quad (3.17)$$

C_{11} and C_{12} are the elements of the 2×2 C-matrix of the system.

3.3 Verification of the FEM software package

Since the FEM package from LLNL does not have automatic mesh generation capability, the user has to define the problem region into finite element partitions manually. Due to the fact that poorly defined elements can cause unstable results [2], a large number of elements does not imply high accuracy. In order to become familiar with and to verify the software package TOPAZ2D, some electrostatic problems with known results were first solved.

3.3.1 Characteristic Impedance of Microstrip Transmission Line

The characteristic impedance of a microstrip transmission line was sought and the result was compared with the one obtained from the software package SUPERCOMPACT (Trade mark of Compact Software Inc). The potential plot of the microstrip is shown in Fig. 3.4. The width of the microstrip is 90 mils and the height of the dielectric substrate is 30 mils, i.e. the W/H ratio is 3. The dielectric constant of the substrate is 10. Also a grounded cover plate is located 120 mils over the top of the substrate. From SUPERCOMPACT, the characteristic impedance is equal to 25.48 Ω . The characteristic impedance obtained using TOPAZ2D and the program CHARGE was 26.15 Ω . Other geometries were also examined with good results.

Fig. 3.4 The potential plot of a microstrip transmission line.

3.3.2 Capacitances of coupled transmission lines

Figures 3.5(a) and (b) are the potential distributions of a pair of coupled transmission lines. The capacitance matrix was found in [4]. Fig 3.5(a) is for the case that a 100 V potential was put on the upper conductor with the lower conductor grounded. Fig. 3.5(b) is for the case that a 100 V potential was put on the lower conductor with the upper conductor grounded. Calling the upper conductor "conductor 1" and the lower conductor "conductor 2", the C matrix can be defined and the values obtained from the FEM package are compared with the ones in [4] in Table 3.1.

Table 3.1

Comparison of the C matrix obtained from the FEM package and reference [4].

C-matrix elements	FEM in [4] F/m	FEM from LLNL F/m
C_{11}	0.3654E-10	0.3402E-10
C_{12}	-0.1538E-10	-0.1450E-10
C_{21}	-0.1599E-10	-0.1444E-10
C_{22}	0.2146E-09	0.1946E-09

Fig. 3.5(a) Potential distribution of coupled transmission lines with the upper conductor at 100 V and the lower conductor at 0 V.

couple lines
time= 0.10000E+01 contours of temperature
dsf = 0.10000E+01

min(-) = 0.75E-30
max(+) = 0.10E+03
contour levels

Fig. 3.5(b) Potential plot of coupled transmission lines with the lower conductor at 100 V and the upper conductor at 0 V.

The differences in the C matrix obtained from the FEM package from LLNL and [4] in Table 3.1 are less than 10%. The difference is believed to be caused by the employment of infinite elements in [4]. In section 3.3.1, the result of the microstrip transmission line is very close to the one obtained by SUPERCOMPACT since it is a semi-bounded problem. i.e., the top and bottom of the solution region are ground plates.

CHAPTER 4

MEASUREMENT AND MODELLING

The Teradyne HD+ high speed connector was measured and modelled in both the time domain and the frequency domain in this project. The goal was to model the connector in one particular case, i.e., with a certain grounding pattern and number of extra ground lines. This model was then modified to fit the measurement of the various other configurations. For the crosstalk model, two transmission models were put side-by-side with coupling added between them. The results are discussed in Chapter 5.

4.1 Measurement

Assuming the device measured was terminated with a linear load and the device itself is non-dispersive, the measurements in time and frequency domain should be related by the Fourier transform. Theoretically, measurement in either domain will suffice. However, to ensure that the data is valid and to provide more accurate modelling, both measurements were taken.

4.1.1 Description of the Test Jigs

Test jigs were built in this work in order to provide low loss adaption between the Teradyne connector and the measuring instruments.

Fig 4.1 Cross-section of the test jig for the Teradyne connector measurements.

The test jigs are actually laminated printed circuit boards. They are called the motherboard and the daughterboard. A cross-section of the boards is given in Fig. 4.1. An RT/Duroid (trademark of Rogers Co.) substrate of dielectric constant 6.0 and 0.025" thickness with copper ground plane was laminated to G10 epoxy glass substrate having thickness 0.0625" or 0.031" for the motherboard or daughterboard, respectively. Before lamination, the ground plane side of the RT/Duroid substrate was patterned to allow 0.085" diameter clearance from the ground plane around all signal holes. After lamination the signal and ground holes were drilled to 0.048" diameter. The holes were then plated through to about 1 mil thickness. The microstrip lines on the bottom side of the PCB were then etched. Also, on the top side of the PCB, a torus of inside diameter 0.065" and outside diameter 0.085" was etched around all signal holes. Thus the G10 substrate had a perforated continuous ground plane.

When laminating the two substrates, the RT/Duroid substrate was allowed to overhang by 0.25". This overhang permitted the soldering of a 0.125" x 0.500" x 0.000" brass stiffener along the length of test board. The stiffener was pre-drilled with the hole pattern required for mounting broadband SMA/microstrip launchers (type OSM-204-8232A). The centre pins of the microstrip launchers were soldered to the 50 Ω microstrip transmission lines.

Microstrip lines were used in the test boards due to their low loss at high frequencies and to permit de-embedding using a wide variety of discontinuity models in

TOUCHSTONE.

The test board approximates an actual G10 epoxy 0.091" thick motherboard or 0.0625" thick daughterboard with a single buried ground layer. The connectors are mounted from the back side of the mother- and daughterboards so that the connectors are characterized with maximum feed-through pin length in the interconnection.

4.1.2 Signal/Ground Pin Configurations

The manufacturer of Teradyne connectors has suggested that different ground patterns may be used in order to reduce the crosstalk level. Three ground patterns are studied in this project: 25%, 50% and 75% grounding. The patterns are shown in Fig. 4.2.

The top and bottom rows for each ground pattern are the additional ground contacts which are made of copper sheets.

It is apparent that the higher the percentage of grounding, the lower the crosstalk level and the lesser the number of usable signal lines. So there is always a trade off between the number of signal lines and the percentage of grounding. In the next two sections, all the possible cases that can be measured are given.

Fig 4.2 Signal/ground pin configurations

(a) 25% ground

(b) 50% ground

(c) 75% ground

4.1.2.1 Transmission (Single Pin) Test Configurations

A complete analysis of the single pin transmission characteristics includes sixteen different configurations for each case of the HD+, HD+1 and HD+2 connectors. These configurations are shown in Fig. 4.3.

Note that in Fig. 4.3, only parts of the connector modules are shown due to the periodic nature of them.

4.1.2.2 Crosstalk (Coupled Pins) Test Configurations

Two adjacent active pins are considered to be coupled pins and there are seventeen different configurations as given in Fig. 4.4.

Fig 4.3 Different ground pin patterns for single active pin.

Fig 4.4 Different ground pin patterns for coupled pair of active pins.

4.1.3 Test Board Patterns

In order to cover all the possible cases shown in the last two sections, a total of fifteen sets of test boards were designed: Six sets for the 25% ground pattern; six sets for the 50% ground pattern and 3 sets for the 75% ground pattern. The layouts of the boards are shown in Fig. 4.5. Note that only one layout is shown for 0, 1 and 2 extra ground rows since they look exactly the same. The numbering system for the patterns can be explained by:

First two digits (MB/DB) - mother board or daughter board.

Third digit - percentage of grounding, 2 stands for 25%, 5 stands for 50% and 7 stands for 75%.

Fourth digit (A/B) - pattern A or pattern B.

Last three digits - used to identify the number of extra ground rows during the etching and lamination process.

In Fig. 4.5, the solid circles are the ground lines and the boards shown are the HD+ only. For the HD+1 connectors, there is one additional row next to the shortest signal line, and for HD+2 connectors, there are two additional rows.

Fig. 4.5(a) Test boards A for 25% ground pattern.

Fig. 4.5(b) Test boards B for 25% ground pattern.

Fig. 4.5(c) Test boards A for 50% ground pattern.

Fig. 4.5(d) Test boards B for 50% ground pattern.

Fig. 4.5(e) Test boards for 75% ground pattern.

4.1.4 Measurement in the Frequency Domain

The scattering parameters or S-parameters were used in this project to characterize the Teradyne connectors. The actual measurements were made on a HP 8720A network analyzer controlled by a 80-386 based personal computer through the GPIB bus for a frequency range from 200 MHz to 20 GHz. The software package ANACAT by EEsof INC. was used to automate the measurement process.

4.1.4.1 CAD Software package ANACAT

ANACAT(trademark of EEsof Inc.) is a computer-aided test program for automatic network analyzers (ANAs). By executing commands from the personal computer, the user can calibrate the ANA, apply bias, take measurements, de-embed measurement data, and sweep data through the GPIB bus that connects the ANA and the personal computer. The data storage in ANACAT is in a database format; comments or description of the measurement can be inserted in the data. This feature is very important when a large amount of data is taken. ANACAT can also convert the measurements in the database into the TOUCHSTONE data file format. In this thesis, the de-embedding feature in ANACAT was not used but instead de-embedding was done in TOUCHSTONE, due to the complicated nature of the embedded measurements which were beyond ANACAT's capability. ANACAT provides a direct interface with the ANA through its graphic interface, allowing the measurement process to be carried out independent of the type of

ANA used.

4.1.4.2 De-embedding with TOUCHSTONE

De-embedding is a very important procedure for the measurement of S-parameters. All S-parameter measurements must include calibration which establishes a reference plane at each port of the device under test. In cases where extra networks are inserted between the ANA and DUT, the desired measurements will be masked by the effect of these extra networks. Examples of extra networks are: adapters for connection purposes, test jigs for physical support and/or biasing purposes, protection purposes, etc. In all cases, the effect is undesired and it is necessary to remove the extra networks in order to isolate the behaviour of the embedded DUT. Test jigs may sometimes cause undesired resonances even though their structures may be very simple. In the Teradyne connector measurements, test jigs with microstrip lines were used to realize the connection of the mother board and the daughter board of a digital system and to provide launcher interfaces between the network analyzer and the connector.

De-embedding can be done in ANACAT but was performed instead in the high frequency CAD program TOUCHSTONE. Firstly, a TOUCHSTONE circuit file defining the microstrip sections of the test jig was created. In this file, the S-parameter file from ANACAT is referred to and TOUCHSTONE was able to "subtract" the effect of the

microstrip circuit of the test jig from the data. The circuit file is shown in Appendix C.1 and the S-parameters for one of the Teradyne connector transmission/reflection measurements before and after de-embedding are shown in Fig 4.6 in the grid format.

Fig. 4.6 S-parameters before and after de-embedding
(a) S_{11} before and after de-embedding.
(b) S_{22} before and after de-embedding.

Fig 4.6 (cont'd)
(c) S_{21} before and after de-embedding.

4.1.5 Measurement in the Time Domain

The Time Domain Reflectometry (TDR) method was used in order to measure the Teradyne connector's time domain performance.

The measurements were performed by DEC-Canada personnel on a TEKTRONICS 11801 digital sampling oscilloscope.

These time domain measurements were used as a means to confirm the S-parameter measurements. An FFT program was used to convert the S-parameter measurements to the time domain response so that a comparison could be made.

4.2 Modelling

An experimental approach was taken in this project in order to model the high speed connector. There is also an alternate approach which involves the full wave electromagnetic analysis.

Full wave analysis usually employs numerical methods to model the device as a whole. For example, the integral equation method has been used to model back plane connector pins [14]. Patches were defined on the connector pins' surfaces. The result was

the C-matrix of the connector as a whole. This method has no intermediate step, i.e., one would not know if the result is legitimate until it is compared with the measurement. Also, the simple capacitor model predicts only a low-frequency filter response. The experimental approach taken on the other hand characterizes the connector as a distributed/lumped elements network in which each element represents a physical part of the connector. Complicated behaviours over the frequency range of interest can thus be modelled.

Modelling of the Teradyne connector was done in two parts, i.e. transmission/reflection and crosstalk. As the first step, the skeleton of the model related to the physical geometry is created as a TOUCHSTONE circuit file. For example, the values of input capacitance and pin inductance were estimated and used as the values of the parameters in the circuit file. TOUCHSTONE has the capability of importing four S-parameter files in one circuit file so the model can be compared with the de-embedded S-parameter measurement file. The frequency response can be shown on a S-parameter versus frequency grid, Smith chart or table format. The grid format is used, in general. The Smith chart which simultaneously display magnitude and phase information is useful when one wants to match resonances in the model to measured resonances. Modelling was the most time consuming stage in this project since a number of iterations were usually required before a reasonable model could be obtained.

After the frequency domain characteristics were matched sufficiently, the model's

response was viewed in the time domain by use of an FFT program. It was often the case that the models' responses in the time domain did not match satisfactorily the TDR data and the parameters of the models had to be modified. Fortunately, time domain responses could always provide considerable insight for the model's response. For instance, one could determine whether more or less capacitance or inductance is needed in the model. Iterations in modelling of both the time and frequency domain responses were made until the best model was obtained.

4.2.1 Transmission/Reflection Model

A cross-section of the Teradyne connector is shown in Fig. 4.7(a) and the corresponding model is in Fig. 4.7(b). The elements of the model are:

L_1 - the inductance of the section of pin that goes from side 1-1' of the motherboard, through the motherboard connector, to the plane where the motherboard connector meets the daughter board connector.

C_1 - the fringing capacitance between the pin that extends in the air at side 1-1' and the ground layer.

C_2 - the fringing capacitance between the section of pin inside the motherboard connector and the buried ground layer.

d_1 - the transmission line representation of the section of conductor pin from the motherboard-daughterboard connector interface to the spring contact inside the daughterboard connector. This transmission line has a characteristic impedance Z_1 and dielectric constant K_1 .

L_2 & C_3 - a LC network to represent the spring contact.

d_2 - represents the section of signal pin inside the plastic housing at the motherboard end of the daughterboard connector.

C_4 - the fringing capacitance caused by the discontinuity of dielectrics inside the daughterboard connector.

d_3 & d_4 - another section of signal pin inside the daughterboard connector having characteristic impedances Z_3 and Z_4 and dielectric constant K_3 and K_4 .

L_3 , L_4 & C_5 - an LC network representation of the bend of the signal line.

C_6 & C_7 - analogous to C_2 and C_1 .

L_5 - the section of signal line passes through the daughterboard connector plastic housing and the daughterboard card.

Fig 4.7 (a) Single line representation of Teradyne HD+ connector.
 (b) Model of the connector corresponds to structure of (a).

Some of the values of the parameters are fixed for the various cases of the connectors and some are allowed to vary accordingly. The ones that are allowed to vary are the lengths of d_1 and d_2 and the characteristic impedances of the transmission lines. The lengths differ according to the different rows of conductor pins, and the characteristic impedances of the transmission lines change according to the ground patterns and the number of additional ground lines. The lengths of the different conductor pins are given in Table 4.1. The lengths of d_1 and d_2 are 100 mils and 350 mils respectively and remain constant. All the transmission line elements in the model are assumed to have an attenuation of 0.001 dB/m at 1000 MHz. Table 4.2 shows the characteristic impedances of the various signal/ground pin configurations calculated using the FEM software from LLNL. Only the values for Z_1 were shown which is surrounded by air. The values of Z_4 are the same and the values for Z_1 and Z_2 are scaled by K_1^{-1} .

Table 4.1

Lengths of transmission lines in the Teradyne model.

ROW	LENGTH d_1 (mils)	LENGTH d_2 (mils)
A	75	52*
B	175	32
C	275	132
D	375	232

* As in Fig. 2.3, ROW A does not go through the plastic housing section. L_5 in the model is assigned to be 3 nH to take this into account (see Table 4.3).

Table 4.2
Characteristic impedances in the transmission/reflection
model of the Teradyne connector.

ROW		A	B	C	D
25%(A)	HD+	158.4	144.3	144.3	158.4
	HD+1	128.6	136.5	139.6	157.7
	HD+2	128.5	136.0	136.0	128.5
25%(B)	HD+	210.1	160.0	160.0	210.1
	HD+1	137.6	150.0	152.5	186.3
	HD+2	137.5	146.6	146.6	137.5
50%	HD+	130.3	127.1	127.1	130.3
	HD+1	117.9	126.5	127.1	130.2
	HD+2	117.9	126.5	126.5	117.9
75%	HD+	128.6	124.6	124.6	128.4
	HD+1	117.1	124.6	124.6	128.4
	HD+2	117.1	124.6	124.6	117.1

The rest of the elements in the model are the inductors and capacitors. Their values were determined by matching the model's response to the measured response. The actual values are listed in Table 4.3.

Table 4.3

Fixed element values in the transmission/reflection model.

Element	Value
C_1	0.6 pF
C_2	0.3 pF
C_3	4.0 pF
C_4	0.2 pF
C_5	0.8 pF
C_6	0.6 pF
C_7	0.6 pF
L_1	3.0 nH
L_2	1.5 nH
L_3	0.6 nH
L_4	0.6 nH
L_5	4.0 nH

Fig. 4.8 shows four examples of the potential distribution of different pin configurations of the Teradyne connector by ORION. The signal pins are at 100 volts and the solid pins are grounded while the rest are floating pins. The top and bottom plates are ground while the sides have the natural boundary conditions. These plots are useful to verify the accuracy of the solution since the contour lines of potential should always be parallel to the conductors (equal potential surfaces). Note that in the figures the contours of temperature are shown but actually they are contours of electrical potential.

Fig. 4.8(a) Teradyne connector, 25% ground pattern, type 1.

Fig. 4.8(b) Teradyne connector, 25% ground pattern, type 2.

Fig. 4.8(c) Teradyne connector, 25% ground pattern, type 3.

Fig. 4.8(d) Teradyne connector, 25% ground pattern, type 4.

4.2.2 Crosstalk model

The crosstalk model for coupled signal lines is based on the transmission/reflection model. Coupled capacitances, mutual inductance and coupled transmission lines are added to two side-by-side transmission models to account for the signal coupling. The model is in Fig. 4.9.

2

of/de

2

PM-1 3½"x4" PHOTOGRAPHIC MICROCOPY TARGET
NBS 1010a ANSI/ISO #2 EQUIVALENT

Fig 4.9 Equivalent circuit or model for the Teradyne connector's near-end and far-end crosstalk performance.

In Fig. 4.9, C_8 and C_9 account for the coupling capacitances between the small sections that go beyond the boards at both ends. The inductors L_1 , L_2 , and L_3 were modified to have mutual inductances M_1 , M_2 and M_3 . Note that no mutual inductance was added to L_3 and L_4 since they represent the bends only. The transmission line sections in the single pin model were replaced by coupled transmission lines with the even- and odd-mode impedances Z_{ie} and Z_{io} , where $i = 1, 2, 3, 4$. These impedances were determined by the MOM software package and their values are shown in Table 4.4.

Table 4.4

Even- and odd-mode impedances and relative dielectric constant used in the Teradyne crosstalk model.

Element	Characteristic Impedance (Ω)
Z_{1e}	105.4
Z_{1o}	71.2
Z_{2e}	105.4
Z_{2o}	71.2
Z_{3e}	149.1
Z_{3o}	100.7
Z_{4e}	149.1
Z_{4o}	100.7
K_1	2.0
K_2	2.0
K_3	1.0
K_4	1.0

The capacitances with respect to the ground are in general decreased in value

compared to the single line model. This is to account for the fact that some of the electric field lines are now terminated on the other signal line rather than on ground. The capacitors C_8 and C_9 represent this coupling. The values of the rest of the elements remain unchanged.

Due to imperfect alignment in the fabrication process, only one configuration was available for the crosstalk measurement. The values of all the elements are listed in Table 4.5.

Table 4.5

Fixed element values of the Teradyne crosstalk model.

Element	Value
C_1	0.4 pF
C_2	0.2 pF
C_3	4.0 pF
C_4	0.1 pF
C_5	0.6 pF
C_6	0.4 pF
C_7	0.4 pF
C_8	0.1 pF
C_9	0.1 pF
L_1	3.0 nH
L_2	1.5 nH
L_3	0.6 nH
L_4	0.6 nH
L_5	4.0 nH
M_1	0.3 nH
M_2	0.2 nH
M_3	0.3 nH
d_3	75 mils
d_4	52 mils

CHAPTER 5

COMPARISON OF MODELS AND MEASUREMENTS

5.1 Comparison of the Transmission/Reflection (Single Pin) Model

The comparison of the transmission/reflection (single pin) models and measurements were made in the cases of the longest and shortest pins, for all the ground patterns and for zero, one and two additional ground rows.

Table 5.1

The modelled cases.

Case	Description
1	Row A, 25%, HD+
2	Row A, 50%, HD+
3	Row A, 75%, HD+
4	Row A, 25%, HD+1
5	Row A, 50%, HD+1
6	Row A, 75%, HD+2
7	Row D, 25%, HD+
8	Row D, 50%, HD+
9	Row D, 75%, HD+
10	Row D, 25%, HD+2
11	Row D, 50%, HD+2
12	Row D, 75%, HD+2

Note that not all conceivable cases were modelled. The reason being that the extra ground row for the HD+1 connector is adjacent to Row A so the second extra ground row of HD+2 adjacent to Row D would have very little effect on Row A's performance (see Fig. 2.2). Hence only HD+ and HD+1 were considered with Row A and for the same reason, only HD+ and HD+2 were considered with Row D.

Due to the large amount of data, only two cases are presented in this report. They are the best and worst cases. The worst case is case 7 in Table 5.1 (Fig.5.1). The best case is case 6 in Table 5.1. However, the alignment problem in the fabrication process has prohibited the measurement of this case so case 3 was used instead (Fig. 5.2).

Fig. 5.1 and Fig. 5.2 each has 11 sub-figures, Figs. 5.1/2(a), (b) and (c) are the TDR measurements. Figs. 5.1/2(d) and (e) are the comparison of S-parameters S_{11} and S_{21} of the transmission/reflection model and the measurements. Figs. 5.1/2(f) to Fig. 5.1/2(k) are the comparison in the time domain by converting the S-parameters of the model and the measurements with the FFT program.

In Figs. 5.1(a) and (b), the TDR measurement of the Teradyne connector from the motherboard and daughterboard sides are shown. The little dots on the traces mark the reflection caused by the Teradyne conductor pin. The waveform to the left of the left dot indicates the coax to microstrip launcher and the microstrip section of the test jig. On the right of the right dot is the multiple-time reflection which carries no important

information. Fig. 5.1(c) is the measurement of the transmission pulse. Transmission measurements are used to identify the degradation of rise-time only.

Figs. 5.1(d) and (e) are the comparison in the frequency domain. In Fig. 5.1(d), the model's input reflection (S_{11}) is compared with the measured data. "TERTX" stands for "Teradyne Transmission model" and "RD25P0" stands for "Row D, 25% ground with 0 extra ground row". The rest of the figures also follow this nomenclature. It can be seen that the modelling was done mainly up to 10 GHz and the roll-off after 10 GHz in Fig. 5.1(d) was not modelled. The high frequency roll-off is believed to be caused by high order parasitic effects in the Teradyne connector and attempting to model these effects by LC networks will make the model much more complicated. In addition, the energy content after 10 GHz is usually small and insignificant. Fig. 5.1(e) is the comparison of the S_{21} data and it is obvious that the model does not match well with the measurement after 9 GHz. However, the model does behave closely to the Teradyne connector from d.c. to around 8 GHz. The general trend of the peaks and troughs were closely approximated. Fine tuning the model could improve the match of the measurement and the model but then the match in the time domain was affected.

Figs. 5.1 (f) and (g) are the comparison of the model and the measurement in the time domain. Fig. 5.1(f) was obtained by feeding the de-embedded S-parameter measurement through the FFT program and Fig. 5.1(g) was obtained by taking the TOUCHSTONE S-parameter file of the model and feeding through the same FFT

program. Figs. 5.1(f) and (g) do not agree with each other closely and is considered as the compromise made in order to match the model and measurement in both time and frequency domains. However, the output reflection (S_{22}) and forward transmission (S_{21}) were closely approximated. They are shown in Figs. 5.1(h) and (i) and Fig. 5.1(j) and (k), respectively.

The results of the best case (case 3) in Fig. 5.2 are not as satisfactory as the worst case (case 7) in Fig. 5.1. The general shape of the response in both the time and frequency domains is still followed.

In addition to the previous comparison, the rise-times for the 12 cases are summarized in the following tables. Table 5.2 list the rise-time measurements and predictions from the model for a 40 ps rise-time input.

(vertical: 20mV/div, horizontal: 200ps/div)

(a)

(b)

Fig. 5.1 Performance measured and predicted, case 7 (worst case).

(a) TDR measurement from the motherboard side.

(b) TDR measurement from the daughterboard side.

Fig. 5.1 (cont'd)

(c) Time domain measurement of the transmission pulse.

(d)

(e)

Fig. 5.1 (cont'd)

(d) Comparison of S_{11} for the model and measurement.
 (e) Comparison of S_{21} for the model and measurement.

Fig. 5.1 (cont'd)

(f) Calculated input reflection from the model.

(g) Calculated input reflection from the measurement.

(h)

(i)

Fig. 5.1 (cont'd)

(h) Calculated output reflection from the model.

(i) Calculated output reflection from the measurement.

(j)

(k)

Fig. 5.1 (cont'd)

(j) Calculated pulse transmission from the model.

(k) Calculated pulse transmission from the measurement.

(vertical: 20mV/div, horizontal: 200ps/div)

(a)

(b)

Fig. 5.2 Performance measured and predicted, case 3 (best case).

(a) TDR measurement from the motherboard side.

(b) TDR measurement from the daughterboard side.

Fig 5.2 (cont'd)
(c) Time domain measurement of the transmission pulse.

(d)

(e)

Fig. 5.2 (cont'd)

(d) Comparison of S_{11} for the model and measurement.
(e) Comparison of S_{21} for the model and measurement.

(f)

(g)

Fig. 5.2 (cont'd)

(f) Calculated input reflection from the model.

(g) Calculated input reflection from the measurement.

(h)

(i)

Fig 5.2 (cont'd)

(h) Calculated output reflection from the model.

(i) Calculated output reflection from the measurement.

(j)

(k)

Fig. 5.2 (cont'd)

(j) Calculated pulse transmission from the model.

(k) Calculated pulse transmission from the measurement.

Table 5.2

Measured and predicted rise-time for the Teradyne connector (ps).

CASE	MEASURED RISE-TIME	PREDICTED RISE-TIME
1	387*	313
2	196	298
3	115	298
4	193	298
5	117	293
6	---	293
7	343	391
8	276	352
9	132	352
10	160	352
11	134	342
12	134	342

* excessive ripple in leading edge of waveform

Rise-time comparison does not show too much similarity between the measurement and the model's prediction. However, it should be pointed out that the modelling was mostly concentrated on the difference in the connector's performance for different grounding patterns, pin lengths and extra ground lines. The predicted rise-times in Table 5.2 do follow the correct trend, i.e., the rise-time decreases as ground pattern is increased from 25% to 75% and/or if extra rows are added.

5.2 Comparison of the Crosstalk (Coupled Pins) Model

Only one set of measurements was available for the crosstalk model due to an alignment problem encountered in the fabrication process of laminating the boards and the ground plate. The available set is the shortest line of the Teradyne HD+1 connector with 25% ground pattern (Fig. 5.13).

Fig. 5.14 shows the comparison of the input reflection, forward transmission, near-end crosstalk and far-end crosstalk between the model and measurements. "TER4P" stands for the four-port S-parameter file of the model and "DUT" is the measurement of the Teradyne connector. Fig. 5.15 shows the TDR measurements of the far-end and near-end crosstalk compared with the model's output converted by the FFT program.

It should be noticed that the model fits extremely well with all the crosstalk measurements. Considering the fact that the crosstalk model was not created separately but by combining two transmission models together, the validity of the transmission model is further confirmed.

Fig 5.3 Balanced signal lines for crosstalk measurement on the Teradyne connector with 25% ground pattern.

(a)

(b)

Fig. 5.4 A comparison of crosstalk performance in the frequency domain.

(a) Near-end crosstalk (line 13 - line 14).

(b) Far-end crosstalk (line 13 - line N).

(c)

(d)

Fig. 5.4 (cont'd)

(c) Input reflection on motherboard (line 13).

(d) forward transmission (line 13 - line M).

Fig. 5.5 A comparison of crosstalk performance in the time domain.
(a) Measured near-end crosstalk. (vertical: 10mV/div, horizontal: 300ps/div)
(b) Near-end crosstalk predicted by the model.

(c)

(d)

Fig. 5.5 (cont'd)

(c) Measured far-end crosstalk. (vertical: 10mV/div, horizontal: 300ps/div)
 (d) Far-end crosstalk predicted by the model.

CHAPTER 6

CONCLUSIONS

Based on the results of Chapter 5, the following conclusions can be drawn.

The transmission/reflection characteristics of the Teradyne connector were reasonably modelled. As shown in Fig. 5.1 and Fig. 5.12, the general trends of the S-parameter behaviour were approximated up to about 10 GHz. The energy content of the signals above 10 GHz is usually very small and the fact that the S-parameters were measured up to 20 GHz was for the need of the FFT program. Small ripples in the frequency domain measurements caused by high order parasitic effects were not modelled since a very complicated model will result otherwise. Also too many elements in a model would make the modelling process very difficult, if viable at all. Although the rise-time performance of the model does not seem to be close to the measurement, the important point is the model correlates to both the S-parameter and the TDR measurements at an adequate level.

The crosstalk model in the other hand demonstrated an excellent match in the near-end and far-end crosstalk performance. Since the crosstalk model was obtained by combining two transmission/reflection models and adding coupling elements between

them, the match in model and measurement of the crosstalk performance implied that the transmission/reflection models are legitimate, too. It was regretful that the alignment problem in the fabrication process prohibited more comparison of the crosstalk performance of the Teradyne connector with respect to the grounding pattern and extra ground rows.

Two numerical methods were employed in this project, one for the transmission/reflection model and one for the crosstalk model. The FEM package from LLNL was used in the transmission/reflection modelling and difficulties were experienced on the work space of memory available to the system equation solver TOPAZ2D. Since the physical structure of the cross-section of the Teradyne connector requires complicated mesh definition, the number of elements involved was usually large and sometimes large enough to make it impossible for TOPAZ2D to run (e.g. Fig. 3.2 has a large number of elements). Analysis of the source code suggested that the limitation of memory work space was caused by the software itself instead of the memory size of the UNIX-based SPARC stations. However, attempts to modify the source code were not successful. To obtain the even- and odd- mode impedances of the crosstalk model which requires many more elements than transmission/reflection model, the MOM package TDTL [12] was used.

Convergence of the FEM software was not analyzed due to the lack of a consistent algorithm for finite element mesh generation. As mentioned previously, a larger number

of elements does not always imply a higher accuracy if long and narrow elements were used. However, the FEM package was shown in section 3.3 to have valid results.

It should be pointed out that small variation in the characteristic impedances of the transmission line elements would not affect the model's performance significantly since their values are large relative to the 50Ω termination in TOUCHSTONE. The different in performance was mainly caused by the change in length of the transmission line elements from row to row.

Theoretically, more than two of the transmission/reflection models can be combined side-by-side to model the crosstalk performance of multiconductor cases. This, however, is limited by the maximum number of ports of the networks supported by TOUCHSTONE.

A drawback of the empirical method used in this project was the enormous amount of time and effort involved in the alternate modelling of the time- and frequency domain responses. As a suggestion to future work on the same kind of problems, the modelling time could be reduced if the time- and frequency domain modelling could be done simultaneously. EEsof Inc. announced recently that software Version 3.0 has the capability of running MICROWAVE SPICE and TOUCHSTONE together in a window environment. Since the circuit files for MICROWAVE SPICE and TOUCHSTONE are very similar, one can see both the time- and frequency domain simulations of the same

circuit on the same screen.

To finalize, the modelling scheme used in this project was found to be feasible since it gave a satisfactory prediction of the transmission/reflection and crosstalk level of the Teradyne connector. With the help of numerical methods, one can gain the insight into the effects of different characteristic impedances of the transmission line elements caused by the signal/ground pin configuration. And it was shown in this work that the characteristic impedance was not the crucial parameter since the values were high and did not change much with respect to ground patterns (see Table 4.2).

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APPENDICES

Appendix A

Some Aspects of TOPAZ2D

A.1 Bandwidth minimization in TOPAZ2D

Since the number of unknowns for the FEM may easily become excessive, efficient handling of large matrix equations is extremely important. The matrix equation in FEM is often sparse, and in TOPAZ2D, An option is available to minimize the bandwidth and profile of the system matrix. As experienced in the microstrip coupled lines cases, this option was necessary to run TOPAZ2D on a Sparc station with 8 MB memory. The bandwidth limitation process also includes a re-numbering step which is transparent to the user. i.e., the numbering scheme used in MAZE stays the same upon termination of TOPAZ2D.

A.2 Boundary Conditions

Boundary conditions can be functions of potential and/or time. In the electrostatic/magnetostatic cases, the potential boundary conditions will suffice since they

are independent of time. As a property of FEM, only forced boundary conditions are needed to be specified since natural boundary condition are always satisfied on the edge of an element.

A.3 Electrostatic Problems

Since TOPAZ2D solves the heat diffusion equation

$$\nabla \cdot k \nabla \theta + q = \rho c \frac{\partial \theta}{\partial t} \quad (\text{A.1})$$

on two dimensional planar and axisymmetric geometries, this equation is applicable to other field problems with the appropriate definition of the physical parameters.

Maxwell's equations for static fields are:

$$\nabla \cdot \bar{D} = \rho \quad (\text{A.2})$$

$$\nabla \times \bar{E} = 0 \quad (\text{A.3})$$

which are connected by the constitutive relation

$$\bar{D} = \epsilon \bar{E} \quad (\text{A.4})$$

where $\bar{D}$ = electric flux density

$\bar{E}$ = electric field intensity

ρ = volume charge density

ϵ = permittivity tensor

Also

$$\bar{E} = -\nabla V \quad (\text{A.5})$$

where V is the electrostatic potential.

Substituting (A.4) & (A.5) into (A.2),

$$\nabla \cdot \epsilon \nabla V + \rho = 0 \quad (\text{A.6})$$

compare (A.6) with (A.1) leads

$$\epsilon = k \quad (\text{A.7})$$

$$V = \theta \quad (\text{A.8})$$

Thus TOPAZ2D can be used to analyze electrostatic problems with proper substitutions for the variables.


```

      read (14, '(1x,3)') (nground(i), i=1, nground)
      read (14, '(1x,3)') (lground(i), i=1, nground)
      goto 260
cdec of (lground ne 'y') and (lground ne 'n')) then
  print *, 'answers with y or n'
  goto 265
cdec
  print *, 'enter number of jet on the top, bottom, left and right.'
  read *, nump, nmbut, nambut, nambig
  print *, 'enter the y co-ordinates for the top and bottom.'
  read *, top, bottom
  print *, 'enter the x co-ordinates for the left and right.'
  read *, left, right
  print *, 'enter ltr of the top, bottom, left and right.'
  read *, ep, epb, epl, epr
  print *, 'enter the potential of the strip'
  read *, fapex
210 print *, 'enter the ', nambig, ' jets on the top.'
  read *, (lground(i), i=1, nground)
  print *, 'confirmed? [y/n]'
  read *, yesno
  if (yesno eq 'y') goto 220
  if (yesno eq 'n') goto 210
220 print *, 'enter the ', nambut, ' jets on the right side.'
  read *, (nground(i), i=1, nground)
  print *, 'confirmed? [y/n]'
  read *, yesno
  if (yesno eq 'y') goto 230
  if (yesno eq 'n') goto 220
230 print *, 'enter the ', nmbut, ' jets on the bottom.'
  read *, (lground(i), i=1, nground)
  print *, 'confirmed? [y/n]'
  read *, yesno
  if (yesno eq 'y') goto 240
  if (yesno eq 'n') goto 230
240 print *, 'enter the ', nambig, ' jets on the left side.'
  read *, (lground(i), i=1, nground)
  print *, 'confirmed? [y/n]'
  read *, yesno
  if (yesno eq 'y') goto 250
  if (yesno eq 'n') goto 240
250 open (unit=14, file='ave', status='unknown')
  write (14, '(1x,4x3)') nump, nmbut, nambig, nambig
  write (14, '(1x,3x15.5)') ep, epb, epl, epr, fapex
  write (14, '(1x,4x15.5)') top, bottom, left, right
  write (14, '(1x,3)') (lground(i), i=1, nground)
  write (14, '(1x,3)') (nground(i), i=1, nground)
  write (14, '(1x,3)') (lground(i), i=1, nground)
  write (14, '(1x,3)') (lground(i), i=1, nground)
260 do until 'end'
  table : hct : 1
  call trapezable (fapex, fapex, top, nump, nambig, ep, epb, epl, epr, fapex)
  print *, 'sum on top : ', nump
  call trapezable (fapex, fapex, bottom, nmbut, epb, epr, fapex)
  print *, 'sum on bottom : ', nmbut
  call trapezable (fapex, fapex, left, nambig, nambig, epl, epr, fapex)
  print *, 'sum on right : ', nambig

```


[illegible]

```

print *
10 read *, character
if (character.eq.1) then
  call single(cap,dimen)
else if (character.eq.2) then
  call couple
else
  print *, 'enter 1 or 2'
  goto 10
endif
end

.....
subroutine single(cap,dimen)
character*10 dimen
real cex, c, ccap, ccoupl
print *, 'in homogeneous or inhomogeneous medium [uA]'
100 read '(0)', m
if (m.eq.'h') goto 500
if (m.eq.'i') goto 400
if ((m.eq.'h') and (dimen.'i')) then
  print *, 'enter h or i'
  goto 100
endif
400 print *, 'enter the capacitance in pF in air'
read *, ccaple
print *, 'enter the capacitance in pF in air and dielectric'
read *, ccapd
print *, 'pF per m or mil?'
410 read '(0)', dimen
if (dimen.eq.'m') then
  m=1/30.48*sqrt(ccaple*ccapd)
else if (dimen.eq.'mil') then
  m=2.54e-5/30.48*sqrt(ccaple*ccapd)
else
  print *, 'enter m or mil'
  goto 410
endif
goto 600
500 print *
print *, 'enter the self capacitance in F...'
print *
read *, ccap
print *
c=10 read '(0)', dimen
if (dimen.eq.'m') then
  z=(1/30.48*sqrt(c))
if (dimen.eq.'mil') then
  z=(1/30.48*sqrt(c*39370.07874))
else if (dimen.eq.'m') and (dimen.ne.'mil') then
  print *, 'enter m or mil'
  goto 510
else
  print *
endif
600 print *
print *, 'characteristic impedance = z, z, ohms'
print *

```

Appendix C

TOUCHSTONE CIRCUIT FILES

C.1 De-embedding File

```

! Circuit for deembedding two-port S-par. files
! SOURCE: C:\EESOF\TDYNMEAS\
!
! Input files: Txxx1A.s2p
! Output files: DTxxx1A.S2P
!
! April 21, 1990

DIM
RES OH
FREQ MHZ
IND NH
CAP PF
LNG MIL
ANG DEG

VAR
LENG1=80
LENG2=380

CKT
MSUB ER=6.0 H=25 T=0.7 RHO=0.84 RGH=0.0
TAND TAND=0.0
TLINP 1 2 Z=50.0 L=440.0 K=1.0 A=0.0 F=0 !SMA connector
MLIN 2 3 W=36 L=1320
MTAPER 3 4 W1=36 W2=10 L=50
MLIN 4 5 W=10 L=LENG1
MBEND 5 6 W=10 ANG=45 M=0
MLIN 6 7 W=10 L=30
DEF2P 1 7 LAUNCH1
!
TLINP 11 12 Z=50.0 L=440.0 K=1.0 A=0.0 F=0 !SMA connector
MLIN 12 13 W=36 L=1575
MTAPER 13 14 W1=36 W2=10 L=50
MLIN 14 15 W=10 L=LENG2
MBEND 15 16 W=10 ANG=45 M=0
MLIN 16 17 W=10 L=30
DEF2P 11 17 LAUNCH2
!
S2PA 22 23 0 TS11A
DEF2P 22 23 DUT
NEG2 21 22 LAUNCH1
DUT 22 23
NEG2 24 23 LAUNCH2
DEF2P 21 24 DEEMB1 !DTS11A

OUT
DUT DB(S11) GR1
DEEMB1 DB(S11) GR1
DUT DB(S21) GR2
DEEMB1 DB(S21) GR2
DUT DB(S22) GR3
DEEMB1 DB(S22) GR3

FREQ
SWEEP 200 20000 200

GRID
RANGE 0 20000 1000
GR1 -42 0 3
GR2 -42 0 3
GR3 -42 0 3

```


C.2 Transmission Model

```

! MODEL OF THE TERADYNE CONNECTOR TRANSMISSION.

DIM
  FREQ MHZ
  RES OH
  IND NH
  CAP PF
  LNG MIL
  TIME PS
  COND /CM
  ANG DEG

VAR
  ATTEN=.001    !.00035
  FR=1000
  Z2=75

EQN
  Z1=Z2/(2**0.5)

CKT

  IND      1      4      L=3
  TLINP    4      25     Z^Z1      L=100      K=2.0      A*ATTEN F*FR
  IND      25     26     L=1.5
  CAP      25     26     C=.4
  TLINP    26     5      Z^Z1      L=350      K=2.0      A*ATTEN F*FR
  CAP      4       0      C=.3
  TLINP    5       6      Z^Z2      L=75       K=1.0      A*ATTEN F*FR !L=375 for D
                                           !L=275 for C
                                           !L=175 for B
                                           !L= 75 for A

  CAP      5       0      C=.2
  CAP      1       0      C=.6
  CAP      12      0      C=.6
  IND      6       7      L=.6
  CAP      7       0      C=.8
  IND      7       8      L=.6
  TLINP    8      10     Z^Z2      L=52       K=1.0      A*ATTEN F*FR !L=232 for D
                                           !L=132 for C
                                           !L=32  for B
                                           !L=52  for A

  CAP      10      0      C=.6
  IND      10     12     L=3      !L=4 to take the extra section into account
                                           !L=3 for Row A exclusively

  DEF2P 1 12 TERTX
  S2PA 21 22 0 DT5B09I
  DEF2P 21 22 DUT

GUT
  TERTX DB[S11] GR1
  TERTX DB[S21] GR2
  TERTX DB[S22] GR3
  DUT   DB[S11] GR1
  DUT   DB[S21] GR2
  DUT   DB[S22] GR3
  TERTX S21
  DUT   S21

FREQ
  SWEEP 200 20000 200

GRID
  RANGE 0 20000 20%
  GR1 0 -42 3
  GR2 0 -42 3
  GR3 0 -42 3

```

C.3 Crosstalk Model

```

! CROSS TALK MODEL OF THE TERADYNE CONNECTOR TRANSMISSION.

DIM
  FREQ MHZ
  RES OH
  IND NH
  CAP PF
  LNG MIL
  TIME PS
  COND /OH
  ANG DEG

VAR
  ATTEN=.001    !.00035
  FR=1000

CKT
  MUC 1 21 2 22 L1=3 L2=3 M=0.3
  CAP 1 21 C=0.1
  CLINP 2 22 23 3 ZE=105.4 ZO=71.2 L=100 KE=2 KO=2 AE=0 AO=0
  CAP 3 4 C=4
  CAP 23 24 C=#
  MUC 3 23 4 24 L1=1.5 L2=1.5 M=0.2
  CLINP 4 24 25 5 ZE=105.4 ZO=71.2 L=350 KE=2 KO=2 AE=0 AO=0
  CLINP 5 25 26 6 ZE=149.1 ZO=100.7 L=75 KE=1 KO=1 AE=0 AO=0
  IND 6 7 L=0.6
  IND 26 27 L=0.6
  IND 7 8 L=0.6
  IND 27 28 L=0.6
  CLINP 8 28 29 9 ZE=149.1 ZO=100.7 L=52 KE=1 KO=1 AE=0 AO=0
  MUC 9 29 10 30 L1=4 L2=4 M=0.3
  CAP 10 30 C=0.1
  CAP 1 0 C=0.4
  CAP 21 0 C=0.4
  CAP 2 0 C=0.2
  CAP 22 0 C=0.2
  CAP 5 0 C=0.1
  CAP 25 0 C=0.1
  CAP 7 0 C=0.6
  CAP 27 0 C=0.6
  CAP 9 0 C=0.4
  CAP 29 0 C=0.4
  CAP 10 0 C=0.4
  CAP 30 0 C=0.4
  DEF4P 1 21 10 30 TER4P

! DEF2P 1 12 TERTX
  S4PA 31 32 33 34 XT2B113
  DEF4P 31 32 34 33 DUT

OUT
  TER4P DB(S11) GR1
  TER4P DB(S21) GR2
  TER4P DB(S31) GR3
! TER4P DB(S41) GR1
  DUT DB(S11) GR1
  DUT DB(S21) GR2
  DUT DB(S31) GR3
! DUT DB(S41) GR1
! TER4P S21
! DUT S21
FREQ
  SWEEP 200 20000 200

GRID
RANGE 0 20000 2000
GR1 0 -42 3
GR2 0 -42 3
GR3 0 -42 3

```

END

0/2-06-9/2

FIN