Simulation of Complex Coupled PCB Layouts with Non-Linear Digital Device Termination

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ABSTRACT

Performance of digital printed circuit board technology is currently limited by transmission line effects of PCB tracks. This paper presents a numerical method for determining the transient response of arbitrarily complex transmission line systems terminated with digital devices. The technique convolves a time domain Green's function with the voltage/current characteristics of experimentally developed macromodels of terminating digital devices. This paper compares experimental results to analytic results and presents guidelines for design and simulation.

1.0 INTRODUCTION

The limiting factor in improving the performance of digital systems is the interconnection technology. The interconnection technology of interest for this paper is the copper tracking found on multi-layer printed circuit boards (PCB). For high density PCBs with high speed digital devices, the copper tracks appear as complex coupled transmission lines with impedance discontinuities. The digital devices present an additional complexity in that these devices exhibit non-linear resistive and reactive termination characteristics. The purpose of this paper is to present a numerical method for simulating transmission lines on PCBs terminated with digital devices.

The basic circuit layout on a PCB contains discrete integrated circuits interconnected with copper tracks over a dielectric laminante. For most high speed designs, a multi-layer PCB must be used with buried power and ground layers. Thus, for high speed design, the copper tracks act as microstrip (or stripline) transmission lines with multiple impedance discontinuities (bends, vias, etc). For high density boards, there may be non-negligible coupling between adjacent tracks. Thus, a PCB layout can be described as a complex multi-port linear transmission line network.

The complexities involved in simulating actual PCB tracking terminated with digital devices involve both describing the transmission line effects of the tracking as well as describing the termination characteristics of the digital devices. To illustrate the complexities of describing the transmission line effects consider a typical layout of the copper tracking between two devices as shown in figure 1.

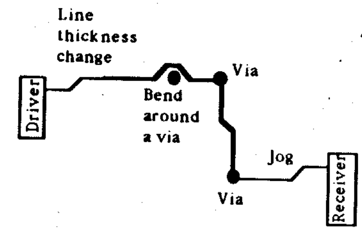

Figure 1: Complex PCB discontinuities

The complex transmission line discontinuities found for a PCB layout include:

- Connectors
- Vias
- Bends
- Bends around a via
- Jogs
- Line thickness changes

In addition, the tracks exhibit lossy or dispersive transmission line behavior. This behavior is due to both finite series resistance of the copper tracks as well as shunt conductance or field losses due to dissipative effects of the dielectric.

Finally, high density PCBs exhibit coupling or crosstalk characteristics between adjacent tracks.

The complexities in describing the digital devices that terminate PCB traces are due to non-linear loading characteristics. This means the resistance and reactance is a function of the device state and the applied voltage. Table 1 illustrates the non-linear termination characteristics of typical TTL digital devices:

	Logic state	Resistance (DC)	Reactance (<100MHz)
Output	low	low	moderately-low capacitive
	high	moderate	moderate capacitive
Input	either*	high	low capacitive

*Most device inputs have undershoot limiting diodes. These diodes exhibit very low resistive impedance and moderate capacitive reactance when forward biased.

TABLE 1: Device Characterization

Once the transmission line effects of the tracks and the termination characteristics of the devices have been described, there is a basic problem with simulating transmission characteristics of PCBs with digital devices. Namely that transmission line effects are best characterized in the frequency domain while a solution to systems with non-linear terminations can only be solved in the time domain.

This paper provides a complete solution to this problem. The solution involves a numerical technique that convolves a time-domain Green's function with the voltage/current characteristics of experimentally developed macromodels of digital devices.

The time domain Green's function is a matrix function with each element being the impulse response of the system at one node. The impulse response is derived from experimentally determine frequency domain scattering (S) parameters.

This method has been implemented on a computer and the analytical results are compared against experimental results.

2.0 BACKGROUND

Other papers have addressed various aspects of the problem of solving the complex transmission line systems and solving transmission lines terminated with non-linear loads.

Shelton [1] simulated a multi-tap backplane using non-linear MSPICE models for drivers and receivers and lumped parameter models (L and C) for transmission lines. While this approach worked well for low speed backplane simulation, lumped parameter models of transmission lines become grossly inaccurate at high frequencies. Furthermore, Shelton did not account for the impedance discontinuities of connectors which have a pronounced effect at high frequencies.

Many authors have addressed the modeling of transmission line discontinuities. For example, Gupta, Garg, and Bahl [2] provided an analytical solutions for microstrips and coupled microstrips, but their work stopped short of providing an analytical solution for vias or coupled lines with discontinuities.

Several authors have addressed the problem of simulating transmission lines with non-linear termination. Schutt-Aine and Mittra [3] provided a time domain solution to a single conductor terminated with digital devices. This work is based on a scattering parameter flow graph approach and addressed only uncoupled lines. This work used simplistic models for digital devices.

Djordjevic, Sarkar, and Harrington [4] developed a technique based on the Green's function for multi-conductor lines. The technique developed by Djordjevic is extended in this paper. Djordjevic's paper is limited because it did not address termination of transmission lines with reactive devices. Both Djordjevic and Schutt-Aine's work did not address the problems of numerical convergence or of numerical calculus.

3.0 DEVELOPMENT OF METHOD

A complex multi-port transmission line network composed of microstrip transmission lines with impedance discontinuities can be fully characterized in the frequency domain using scattering parameters. Scattering parameters define the relative amplitude and phase of the forward and backward travelling waves at each frequency with reference to a known impedance.

Figure 2 illustrates an N-port system terminated with digital devices.

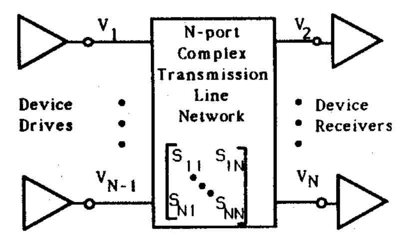

Figure 2: Complex Transmission line system terminated with digital devices

The digital devices that terminate the multi-port transmission line network have non-linear termination characteristics.

The basic method for solving a complex non-linear transmission line systems terminated with digital devices involve the following steps:

1) Characterizing the digital device terminations.
2) Developing the scattering matrix for the N-port system in the frequency domain.
3) Developing a time domain Green's function from the scattering parameters.
4) Removing the effects of the reference impedance associated with the scattering parameters.
5) Developing a system of non-linear termination equations and convolution equations.
6) Solving the system of equations using a numeric technique.

3.1 Termination Models of Digital Devices

Most digital devices can be modelled by an EMF source, a series non-linear resistive element, a shunt non-linear capacitive element, and a series linear inductor as shown in figure 3. The same basic model is used for both the input and the output port of most TTL devices; however, the functions that describe the non-linear characteristics is port and technology dependent.

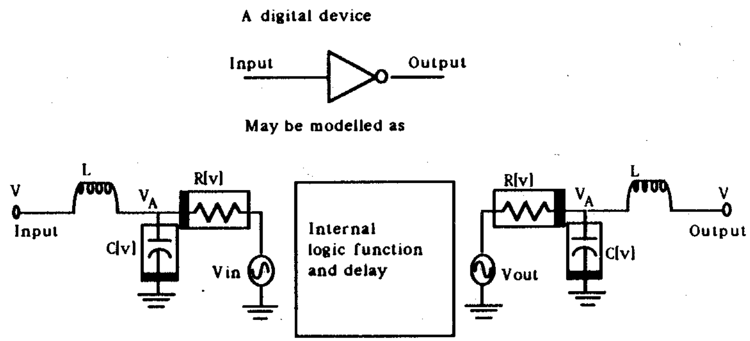

Figure 3: Models of digital devices

The symbol R[v] indicates a nonlinear relationship of voltage across the resistive element to current through the element. This function is obtained from either the IV curves found in the data books or by experimentation. For resistive elements the following constitutive relationship is used [5] for each port i.

(1) $$R[V_{in_i} - V_{A_i}] - I_{Ri} = 0$$

R[v] must be a continuous and differentiable function. Figure 4 and 5 show the output IV characteristics for typical high speed TTL devices.

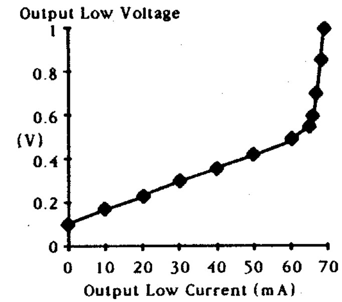

Figure 4: Output Low V/I Curve

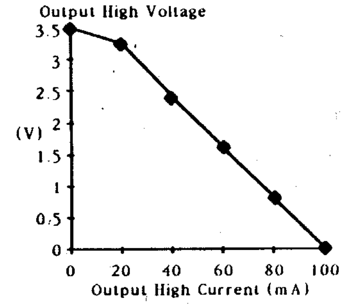

Figure 5: Output High I/V Curve

Figure 6 shows the input IV characteristics for a typical high speed TTL device.

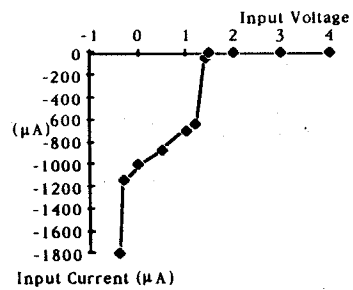

Figure 6: Input V/I Curve

The symbol C[v] in figure 3 indicates a non-linear relationship between the charge on the capacitive element and the applied voltage. A function that can be used to describe the non-linear capacitive characteristics of TTL devices is obtained from the reverse biased capacitance of a graded PN junction.

For capacitive elements, it is best to introduce charge as an additional variable to separate the constitutive relationship from the differential equation. Thus:

(2a) $$C[V_{A_i}] - Q_{c_i} = 0$$

(2b) $$I_{ci} = \frac{dQ_{c_i}}{dt}$$

The function **C[v]** is a continuous and differentiable function of charge in terms of voltage at each port **i**.

The inductor, in figure 3, is linear for digital devices and due mainly to packaging technology. The current through the inductor is given by the following formula at each port **i**:

(3) $$I_L = \frac{1}{L_i}\int_0^t (V_{A_i} - V_i)\,dt$$

3.2 Development of Scattering Matrix

In order to incorporate arbitrarily complex layouts, this paper proposes the development of a library of measure scattering parameters for PCB sub-structures (lines, bends, vias, coupled lines, etc.). Sub-structures are then cascaded using transmission line circuit theory to find the S-parameters for a complex layout.

To find the scattering parameters for PCB sub-structures we used a network analyzer and measured actual structures. The measurements were de-embedded to remove the effect of the coax-to-microstrip connectors [11],[12]. Other suitable methods to find the scattering matrix for sub-structures may involve field theory analysis.

3.3 Development of Time Domain Green's Function.

For any linear system, the response at an observation point can be found by the method of Green's function. The theory associated with the Green's function states that the total response of a system for an arbitrary source can be determined by convolving the source with a Green's function over the variable of interest. The Green's function is the response of the system at the observation point to a unit delta source. Figure 7 illustrates a transmission line network with known reference impedances and arbitrary sources.

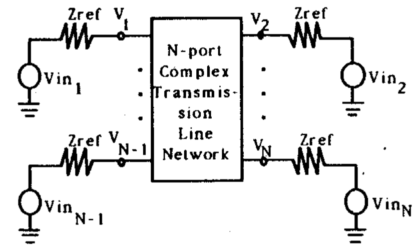

Figure 7: N-port system with known Reference impedance and arbitrary sources.

For transmission line systems the variable of interest is time and the convolution is over a given period of time. The Green's function is the time domain impulse response at the port of interest.

The Green's function or impulse response is found from the system scattering parameters according to the following formula:

(4) $$g_{ij}(t) = \frac{1}{\Delta t}F^{-1}\left[\begin{cases} 0.5(1+S_{ij}(\omega))H(\omega) & i = j \\ 0.5(S_{ij}(\omega))H(\omega) & i \neq j \end{cases}\right]$$

The term **$g_{ij}(t)$** is the time domain impulse response at port **j** and at time t to a unit delta source at port **i** and time zero. The function **$F^{-1}[\ ]$** is the inverse Fourier transform, $S_{ij}(\omega)$ is the frequency domain scattering parameters at radian frequency ω. **$H(\omega)$** is a low pass filter transfer function at radian frequency ω. Finally, Δt is the time step used in the numerical convolution. The factor of 0.5 is obtained from flow graph techniques.

The low pass filter is required to band limit the scattering parameters. In general, low loss systems do not exhibit a rolloff characteristic of the S-parameters. The LPF forces a rolloff in the frequency domain in order to reduce aliasing errors. Provided that the band-stop frequency is chosen high enough, the error introduced by the filter is small compared with the error introduced by aliasing. Aliasing errors can be particularly acute for low loss systems in which the source has a short rise or fall time.

Once the Green's function is obtained from the system S-parameters, the response to arbitrary sources can be found using the convolution equation.

(5) $$V_j(t) = \sum_{i=1}^{Nport} \int_{-\infty}^{t} G_{ij}(t-\tau)\,Vin_i(\tau)\,d\tau = \sum_{i=1}^{Nport} G_{ij} * Vin_i$$

The symbol $*$ denotes a convolution, **Nport** refers to the number of ports in the system, and $Vin_i()$ is the source at node i.

3.4 Removing the Effect of the Reference Impedance

To remove the reference impedance from the system a negative reference impedance can be placed in series between the digital device and the reference impedance. This creates a virtual short circuit between the load and the transmission network as shown in figure 8.

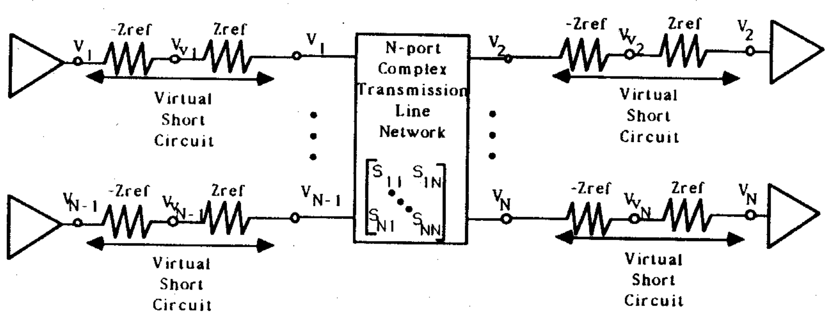

Figure 8: N-port transmission line network with reference impedances removed.

By using the compensation theorem, the circuit shown in figure 8 can be separated into a set of convolution equations and non-linear termination equations as shown in figure 9. The termination of the devices are explicitly shown in figure 9.

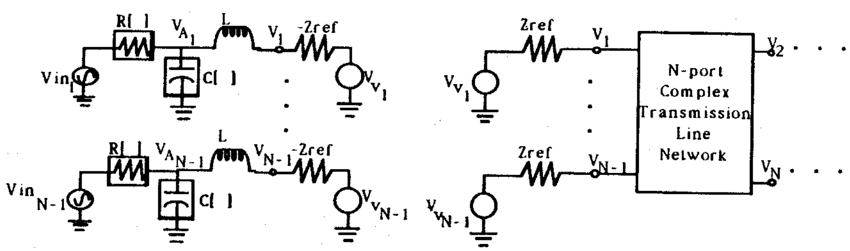

Figure 9: Separation of non-linear termination from transmission line network

The compensation theorem states that any impedance in a network may be replaced by a generator of zero internal impedance, whose generated voltage at every instant is equal to the instantaneous potential difference that exists across the impedance because of current flowing though it. In figure 9, the zero impedance generator is designated as Vv_i (designating a virtual source).

3.5 Development of a System of Equations

To solve the N-port system we use basic circuit theory to set-up the equations at each port for the non-linear termination. The transmission line system is solved at each port using a convolution equation. Thus, the system of equations at each port i are:

(6a) $$Rl[V_{in_i} - V_{A_i}] - I_{Ri} = 0$$

(6b) $$C[V_{A_i}] - Q_{c_i} = 0$$

(6c) $$I_{Ci} = \frac{dQ_{ci}}{dt}$$

(6d) $$I_{Ri} - I_{Ci} - \frac{1}{L_i}\int_{-\infty}^{t}(V_{Ai} - V_i)dt = 0$$

(6e) $$\frac{1}{L_i}\int_{-\infty}^{t}(V_{Ai} - V_i)dt + \frac{1}{Zref}(V_i - V_{vi}) = 0$$

(6f) $$V_i - \sum_{j=0}^{Nport} G_{ji} * V_{vj} = 0$$

3.6 Solving the System of Equations using a Numerical Technique

Since the system of equations at each port contain non-linear equations that do not have a close form solution, a numerical method must be employed to find a solution.

The derivative in equation 6c can be estimated using a finite difference method:

(7) $$\frac{dQ_{ci}}{dt} \approx \frac{Qci(t) - Qci(t-1)}{\Delta t}$$

The integration of equation 6d and 6e can be estimated using the trapezoidal rule.

$$(8)\quad \int_{-\infty}^{t} [V_{A\,i}(t)-V_i(t)]dt \approx \frac{1}{2}\Delta t \sum_{p=0}^{q} [V_{A\,i}(p)-V_{A\,i}(p+1)-V_i(p)+V_i(p+1)]$$

Finally, the convolution of equation 6f can be estimated using Simpson's rule:

$$(9)\quad G_{ji} * V_{vj} = \int_{-\infty}^{t} G_{ji}(t-\tau)V_{vj}(\tau)\,d\tau \approx \sum_{p=0}^{q} G_{ji}(q-p)V_{vj}(p)\Delta t$$

In general the initial value of a digital system is not zero at all nodes. Therefore to obtain an accurate solution with a non-zero initial value a circular convolution is used. Thus,

$$(10)\quad G_{ji} * V_{vj} \approx \sum_{p=0}^{q} G_{ji}(q-p)V_{vj}(p)\Delta t + \sum_{p=q+1}^{NOT} G_{ji}(p)V_{vj}(0)\Delta t$$

where $V_{vj}(0)$ is the initial value of the virtual source at port j and NOT is the number of time points in the time period of interest.

The system of equations can be solved using a Newton-Rapson method expanded to multi-dimensions [6]. The Jacobian matrix used in this method is generally sparse. An iterative solution to the Jacobian matrix produces the next iteration faster than a direct method (such as Gaussian elimination and backward substitution) for systems of six or more ports. For systems of few ports a direct method is faster.

The initial iteration value used for this method can be obtained from the linear interpolation of the previous two values.

4.0 RESULTS

To test the accuracy of this method we compared the analytic results against measured results for two-port structures. We also present four-port simulations.

For the two port setup we measured the S-parameters with a network analyzer across a card-backplane-card configuration. The S-parameter readings were then de-embedded in order to remove the effect of the coax-to-microstrip connectors. These connectors are required for connecting the network analyzer to the structure.

Results were then simulated for a 50% duty cycle clock pulse generated by F buffer/drivers. The analytic results were compared against experimental results. The experimental results consisted of a 74F244 on each card of the card-backplane-card setup. Figure 10 shows the analytic results. Figure 11 shows the measured driver response and figure 12 shows the measure receiver response.

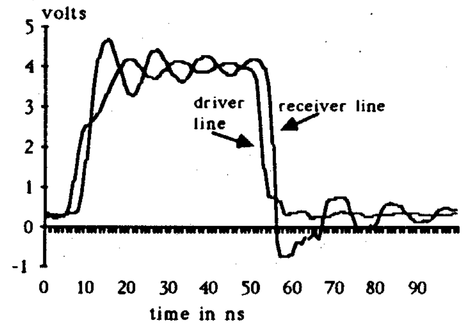

Figure 10. Analytic results for 74F244 at either end of a card-backplane-card setup.

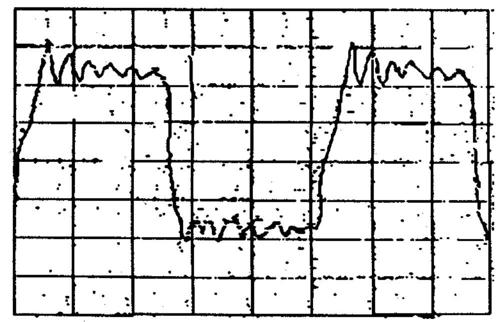

Figure 11. Driver side measured response

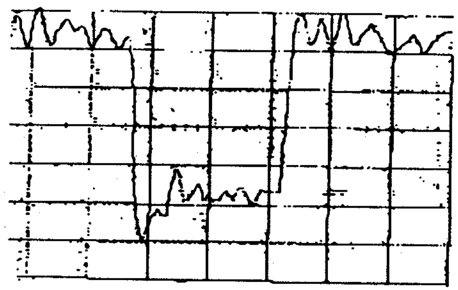

Figure 12. Receiver side measured response.

Similarly accurate results were obtained with a four-port setup. Figures 13 and 14 capture the transmission line effects of two 12 inch coupled low loss lines terminated with F buffer/drivers. A device at port one outputs a 20MHz clock with a 50% duty cycle. The subsequent ringing and crosstalk are displayed.

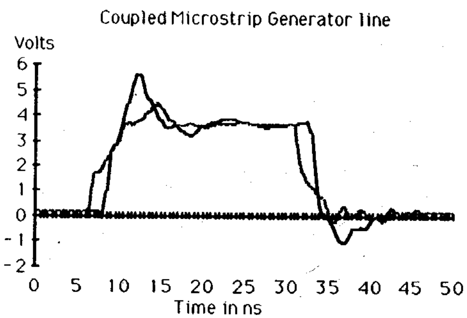

Figure 13. Generator Line Response

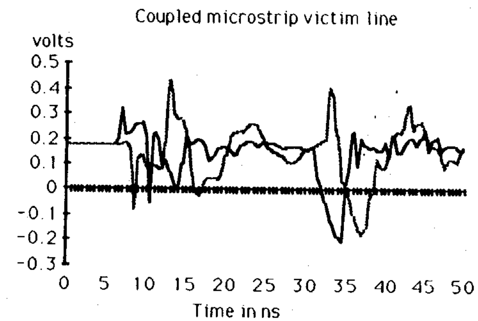

Figure 14. Crosstalk Line Response.

4.1 DISCUSSION

For the four-port simulation, the ringing and cross-talk due to the HIGH-to-LOW transition is larger than the effect due to the LOW-to-HIGH transition. This is due to the fact that impedances are more closely matched when devices are in the high state then when devices are in the low state.

For the two port simulation, the difference between the analytic and experimental results are very small, some of the discrepancies between the measured and analytic results are due to the following factors:

- Bandlimiting of the S-parameters to 3.0 GHz because of equipment limitations. This bandlimit removed high frequency components from the final result.
- Low pass filtering with a pass frequency of 1.5 GHz. This reduced errors due to aliasing but also reduced the high frequency components.
- Ringing due to oscilloscope probes. The probes of an oscilloscope introduce the greatest amount of error in a time domain measurement. This ringing cannot be de-embedded
- Device models are only approximation to actual transistor termination characteristics.
- Numerical errors due to approximation of integrals and derivatives.

In general the problems of numerical mathematics fall into two general categories: convergence and accuracy.

Provided the step size (Δt) is chosen small enough, the solution will generally converge. The point were the solution may not converge is around the point where the negative voltage limiting diode of the input devices becomes forward biased. A further reduction of the step size generally improves convergence around this point.

However, there are accuracy problems associated with a small values of Δt which include:

- Use of double precision numbers for calculations
- Long execution time
- Round off errors due to very small numbers
- Large memory requirements
- Increased cumulative errors

In general the derivative in equation 7 is the most error prone portion of the numerical method. Reducing the value of Δt generally improves the accuracy of the derivative but may reduce the accuracy of the overall solution.

Finally, the inverse Fourier transform required to produce the time domain Green's function generally introduces aliasing errors due to finite bandwidth of the S-parameters. Increasing the bandwidth of the S-parameters reduces the aliasing errors. Increasing the pass band of the low-pass filter increases the bandwidth of the final results but can also increase the aliasing errors.

5.0 CONCLUSION

A numerical technique to solve arbitrarily complex transmission line system terminated with digital devices is presented. This technique convolves the time domain Green's function with the voltage/current characteristics of macromodels of digital devices. The time domain Green's function is derived from measured scattering parameters. The macromodels of digital devices are derived from measured data. Good agreement was obtained between simulated and experimental results.

6.0 REFERENCES

[1] R. Shelton, Sequent Computer Systems, Inc. "VLSI Design Technology, Backplane Simulation". ESD:, May 1988 pp 87-95.

[2] K.C. Gupta, R. Garg, I:J. Bahl. Microstrip Lines and Slotlines. Artech House,1979.

[3] J.E. Schutt-Aine, R. Mittra. "Scattering Parameter Transient Analysis of Transmission Lines Loaded with Nonlinear Termination". IEEE Transactions on Microwave Theory and Techniques, Vol. 36, No. 3, March 1988.

[4] A.R. Djordjevic, T.K. Sarkar, R.F. Harrington. "Analysis of Lossy Transmission Lines with Arbitrary Nonlinear Termination Networks" IEEE Transactions on Microwave Theory and Techniques, Vol. 34, No. 6, June 1986.

[5] J. Vlach, K.Singhal. Computer Methods for circuit analysis and design. Van Nostrand Reinhold Co.,1983.

[6] G. Dahlquist, A. Bjorck. Numerical Methods. Prentice-Hall, 1974.

[7] S. Akhtarzad, T.R. Rowbotham, P.B. Johns. "The Design of Coupled Microstrip Lines". IEEE Transactions on Microwave Theory and Techniques, Vol. 23, No. 6, June 1975.

[8] L.S. Napoli, J.J. Hughes. "Characteristics of Coupled Microstrip Lines". RCA Review, Sept 1970.

[9] G. Gonzalez, Microwave Transistor Amplifiers. Analysis and Design. Prentice-Hall,1984

[10] R.K. Keenan. Decoupling and Layout of Digital Printed Circuits. The Keenan Corporation, 1985.

[11] J. Kasten, M. Steer. De-embedding of Two-Port Networks With Symmetric Fixtures. To be published IEEE MTT.

[12] J. Kasten, M. Steer, R. Pomerleau. Through Symmetric Fixture: A Two-Port S-parameter Calibration Technique. Submitted for publication RF Expo East 88.