The Samtec Golden Standard: A Reference Structure for Electrical Simulation and Measurement:

Part II

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Abstract

Modern electromagnetic analysis and design of transmission line structures relies heavily on numerically based EM field solvers incorporated in most commercial system level CAE and CAD tools. These solvers utilize closed form equations (CFE), method of moments (MOM), finite element analysis (FEA), finite difference time domain analysis (FDTD), or transmission line matrix method (TLM) to evaluate the EM fields and transmission line parameters of two dimensional or, as appropriate, three dimensional transmission line structures in either the time or frequency domains.

In this paper, several types of these solvers, i.e. 2D CFE, and MOM, as well as 3D TLM, are used to analyze a lossy FR-4 circuit board structure comprising of two coupled microstrip transmission lines. The CFE method uses a lossless model, whereas the MOM and TLM methods utilize models that incorporate both frequency dependent copper and dielectric losses.

This simple coupled microstrip test board structure presents a relatively rich set of time and frequency domain features such as modal dispersion, and both TEM and distributed directional couplers. These features are used to evaluate the quality of the 2-D and 3-D simulations including the effects of losses, relative to the expected analytical and measured performance of the test board.

1.1 Transmission Line Basics

To understand the following sections, it is essential to fully understand a transmission line and its properties. In broad terms, a transmission line is a uniform system or line consisting of two parallel conductors. This means that the dimensions, materials, and cross-section of the line and its surrounding environment remain constant throughout its entire length. These conductors do not need to be the same material or similar in anyway between each other, but rather they, in and of itself, must remain the same for its length.

A second requirement is that the current flowing through the conductors flow in the direction of the line, and that the instantaneous current in the two conductors are equal in magnitude and opposite in direction.

Another requirement is that the transmission line can be completely described by four electric circuit coefficients whose value per length are constant everywhere along the line. These coefficients are inductance, resistance, capacitance, and leakage conductance. The inductance and resistance are series coefficients, whereas the capacitance and leakage conductance are shunt. Below is a model of a transmission line using these four coefficients.

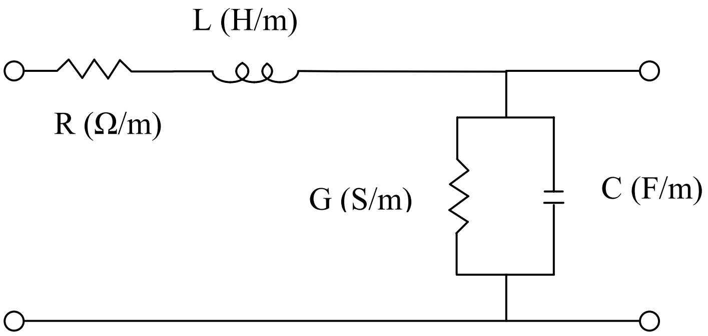


Figure 1: Transmission line model with lumped elements

These parameters do vary with frequency, line voltage, or current and are explicitly determined by the materials and dimensions of the line conductors and their surrounding medium. The materials and physical geometry of the line determines the impedance of the line, and therefore, any physical change made throughout the transmission line effects these parameters and therefore its impedance. These changes in impedance are called discontinuities.

We are familiar with transmission lines, such as TV coaxial cable, being round and therefore contain long, round conductors. Conductors can be round, flat, hollow, solid, or any geometry necessary. In today's world of printed circuit boards, space and cost come at a premium, and therefore, very thin, flat transmission lines are used. These transmission lines are copper traces of varying width and thickness on a non-conducting, flat material.

1.2 Microstrip Basics

A microstrip line is one of the most popular types of planar transmission lines primarily because of its ease of fabrication and its compatibility with other passive and active microwave devices. Figure 2 is a classic example of a transmission line: a copper transmission line of width W and thickness t deposited onto a thin dielectric with relative permittivity ε_r, and a copper ground plane underneath.

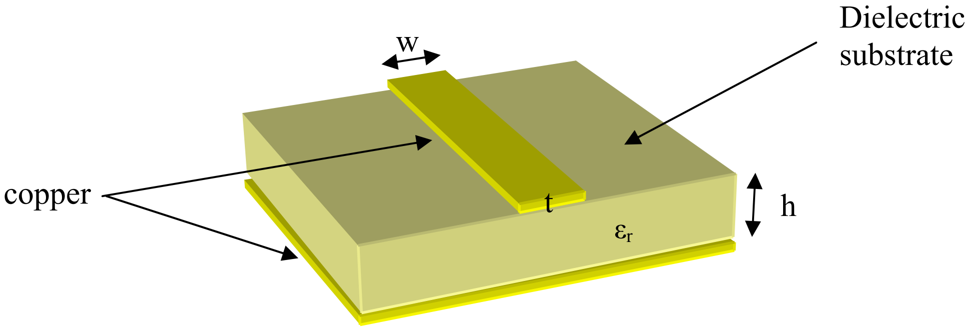


Figure 2: Microstrip on dielectric

Due to the open structure of the microstrip line, the EM field is not confined solely to the dielectric, but rather to both the air and dielectric as can be seen in Figure 3.

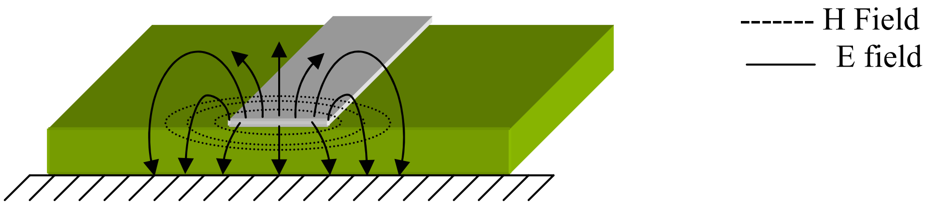


Figure 3: Electric and magnetic field lines

Due to this difference in propagating media, calculations for the impedance are different than those for a stripline where the conductor is completely enveloped within the dielectric. The relative permittivity is no longer valid by itself as the media is discontinuous. The effective permittivity, or ε_{eff}, is used in its place and is calculated by using Equation 1.

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + \frac{10h}{W}\right)^{-\frac{1}{2}}$$

Equation 1: Effective dielectric constant of wide microstrips where W/h>1.3

The effective permittivity will be less than the relative permittivity of the substrate due to fringing effects of the line. This new value will be used to determine the impedance of the line.

1.2.1 Transmission Line Impedance

Transmission line impedance, Z_o, is greatly determined by the physical geometry of the line, and therefore, each configuration has its own equation for this calculation. The impedance of a microstrip transmission line described above is governed by the following equation:

$$Z_O = \frac{60}{\sqrt{\varepsilon_{eff}}} \ln\left(\frac{8h}{W} + \frac{W}{4h}\right)$$

Equation 2: Impedance equation for microstrip

1.2.2 Coupling

When two unshielded transmission lines are close together, power can be coupled between the lines due to the interaction of the electromagnetic fields generated by each line. Such lines are commonly referred to as coupled transmission lines.

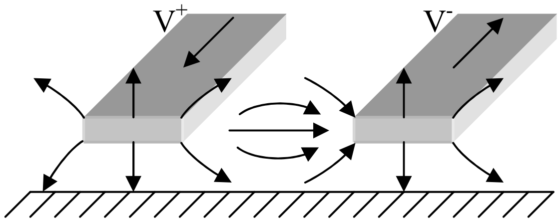


Figure 4: Electric fields for odd mode coupled lines

The coupling that takes place during this type of geometry is called distributive-type coupling. One requirement is that the length of coupling must be five to six wavelengths long so the transmission lines are supporting non-TEM modes. The further the lines are from each other, the more shallow the coupling fields propagate into the substrate.

1.2.3 Time Domain Reflectometry (TDR)

Generally speaking, there are two types of propagating waves on transmission lines: forward and backward. Forward waves travel from the source (generator) to the load and vice versa for the backward waves. If a transmission line is terminated with a load having the same exact impedance as the line, then there will be no backward waves. This line is deemed to be "matched". When the load and line do not match by any degree, then backward waves are generated, and these waves are called reflections, and the line is deemed to be "mismatched". As there is no such thing as a perfect match, there are always reflections inherent in cable transitions, and we can measure these using the method of Time Domain Reflectometry.

1.2.4 Velocity of Propagation (VOP)

The VOP is a specification of the cable indicating the speed at which a signal travels down the cable. Different cables have different VOPs. In order to assure the most accurate distance measurements, the cable VOP must be determined.

Signals transmitted into transmission lines would travel at the speed of light if there were no losses in the material. This velocity of propagation is governed by the effective relative dielectric constant, ε_{eff}, and the relative permeability, μ_r. The relative permeability is a measure of the materials effect on magnetic fields. By definition, it's the specific capacity of a body for magnetic induction or conducting power for lines of magnetic force. When this is less than 1.0, the substance is diamagnetic; above 1.0, paramagnetic; and when high, ferromagnetic. In most cases for transmission lines, this value is approximately equal to 1. The velocity of propagation in a transmission line is:

$$v_p = \frac{c}{\sqrt{\varepsilon_r * \mu_r}}$$

where c is the speed of light (3×10^8 m/s). A signal traveling in air will travel at the speed of light, and its VOP is represented by the number 1 (100%). All other signals are slower as they have an associated dielectric constant which slows the signal. A cable with a VOP of .85 would transmit a signal at 85% of the speed of light. A twisted pair cable, which typically has a lower VOP (such as .65), can transmit a signal at 65% of the speed of light.

1.2.5 Time Delay

In most applications, a signal consisting of 1 or more frequencies is sent down a transmission line. In others, such as computer networks, a pulsed signal may be sent down the line. A single pulse consists of multiple frequencies, so it can be regarded as sending multiple signals at once.

When any signal is transmitted into the end of a transmission line, it takes a certain amount of time to get to the other end. This time is called the transient and is represented by the letter tau, τ. Naturally, the longer the length of transmission line and the slower the VOP, the longer the signal will take to reach the end. This time is calculated by dividing the length of cable by the velocity of propagation.

$$\tau = \frac{l_{cable}}{v_p}$$

1.2.6 Rise Time Degradation

Rise time degradation is a measure of the increase in rise time as the input pulse travels through a connector or probe head before entering the cable or substrate used for transmission. It is given as a percentage and is given by:

$$\%\Lambda t_r = \frac{t_{r_2} - t_{r_1}}{t_{r_1}} x100\%$$

where $\%\Lambda t_r$ is the percent of degradation, t_{r_1} is the input rise time, and t_{r_2} is the output rise time from the connector or probe head.

Rise time degradation can be viewed as an impedance mismatch, attenuation of the signal, additional inductance or capacitance, or both. Care must be taken to properly match the launch to the line or substrate to minimize any reflections that could prevent precise measurements from being taken.

1.2.7 Loss Tangent

The loss tangent is a value that determines how lossy a material is.

A typical application requires a constant loss tangent from near DC to 20 GHz. The dielectric constant numbers, as an example, for FR4 are Er = 4.3 and tan delta = .035. The requirement is for the loss tangent to be constant across the band.

1.2.8 Reflection & Return Loss

When a signal, V_o^+, reaches a discontinuity which can be another transmission line, via, device, resistor, capacitor, etc, a certain percentage of the signal will be absorbed and a certain percentage will be reflected. The amplitude of the reflected voltage normalized to the amplitude of the incident voltage is known as the reflection coefficient (Γ).

$$\Gamma = \frac{V_o^-}{V_o^+} = \frac{Z_L - Z_O}{Z_L + Z_O}$$

The reflection coefficient is a value ranging from -1 to 1. From this equation, you can see that if the load matches the transmission line then $\Gamma=0$, and we have a perfect match which is indicative of no reflected signal and maximum power delivered to the load. If the load is open or a very large impedance ($Z_L >> Z_O$), then $\Gamma=1$, and there is no power delivered to the load, and the signal is completely reflected having the same polarity as the incident signal. For a short or $Z_L = 0$, we have $\Gamma=-1$, and again, there is no power delivered to the load and the signal is completely reflected having the opposite polarity as the incident signal. When the load is mismatched, $\Gamma \neq 0$, then all of the power is not delivered to the load. This amount not delivered is the amount of "loss" and is called the Return Loss (RL) and has units of decibels (dB).

$$RL = -20\log|\Gamma|\ (dB)$$

For example, say a 50Ω transmission line is terminated with a 75Ω load and driven with a 3 volt signal then $\Gamma = .3333$. This means that only 2 volts are delivered to the load, and 1 volt is reflected. These signals are noise on the line, and if they are large enough, can cause unwanted triggering of digital signal or unwanted noise in amplifiers.

1.2.9 Cross Talk

Cross talk is the undesirable capacitive and inductive coupling between two or more conducting paths. Figure 5 will be used for discussion. Both lines are separated by a very small distance, and therefore are closely coupled. They are related by inductive (L_m) and capacitive (C_m) coupling.

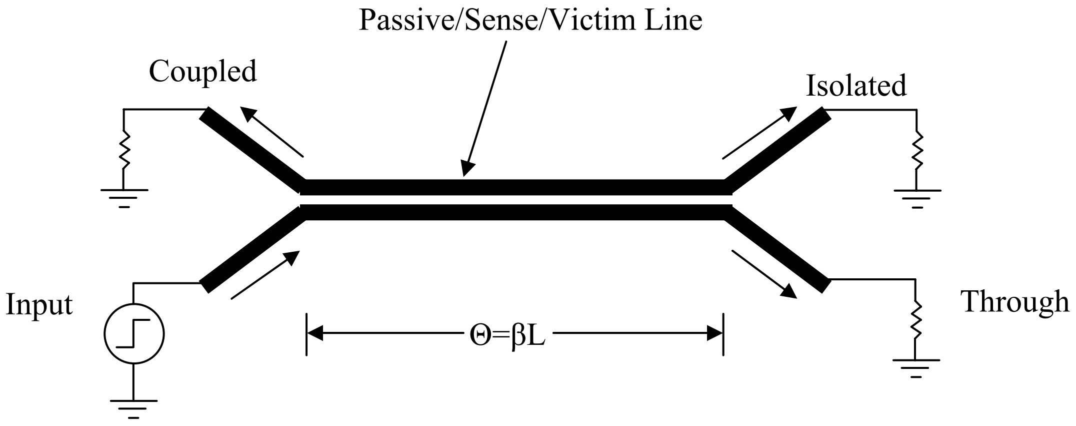


Figure 5: Coupled line couple

Each transmission line load is matched to the characteristic impedance of the line. At t=0, a fast rise time pulse is driven into the input of the coupler, and there is no signal present in the sense line. The signal propagates down the line at it's velocity of propagation while being slowly attenuated. While there is some coupling along the diagonal lines, we will ignore these for clarity purposes. When the launched pulse reaches the parallel section of the lines, the electric and magnetic fields change creating a similar signal, but of much lower magnitude, on the sense line.

Mutual Capacitive Coupling - a signal, SC, caused by capacitive coupling between the two traces which travels along the victim trace in both the forward and backward direction with the same polarity.

Mutual Inductive Coupling - a signal, SL, caused by inductive coupling between the two traces which travels along the victim trace in both the forward and backward direction with opposite polarity.

Directionality - Crosstalk goes in both the forward and backward direction. Mutual capacitive and inductive forward crosstalk are approximately equal and opposite and tend to cancel. They are approximately equal and reinforcing in the reverse direction, and therefore tend to be additive.

1.2.10 Near-End Crosstalk (NEXT)

NEXT is simply defined as the unwanted signal coupling from a near-end transmitter into a pair measured at the same end.

1.2.11 Far-End Crosstalk (FEXT)

FEXT is defined as a measure of the unwanted signal coupling from a transmitter at the near-end into a neighboring pair measured at the far-end.

Time domain reflectometry is a test whereby a signal analyzer sends a fast rising step of energy into the DUT and then picks up any reflected signals from any discontinuities in the path being measured. Recall that with ANY change in impedance, the reflection coefficient will be nonzero, and reflected signals will be generated. These signals echo back to the source and are picked up by the signal analyzer. The TDR test can then determine the impedance of the line and its nature (inductive, capacitive, or resistive).

Now, if the transmission line is matched with the load, there will be no reflection. If a mismatch occurs, part of the signal will be absorbed and part will be reflected. The reflected part will reach the source in the same amount of time it took to reach the load. So from a source perspective, the time it takes to receive a reflected signal will therefore take 2τ. This will tell us how long the line is and how well the line is matched by the magnitude of the reflected signal.

1.2.12 Single-Ended Test

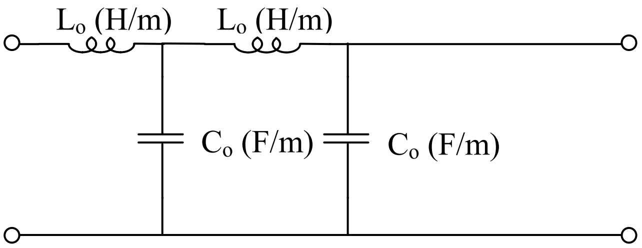
A single-ended test requires two conductors:one at ground potential, and the other the excited line, whereby the signal is injected.

1.2.13 Differential Test

Two conductors are injected with the same signal, but with opposite polarity. Many communication protocols use a differential setup and are done so to increase signal integrity especially in high noise environments. The benefit is that the field generated by the one conductor is nearly cancelled out by the field of the other, thereby resulting in very little crosstalk and radiated emissions.

1.3 Lossy versus Lossless Transmission Line Models

A lossless line has no real resistance or conductance in either the conductor or the dielectric. It can be represented as



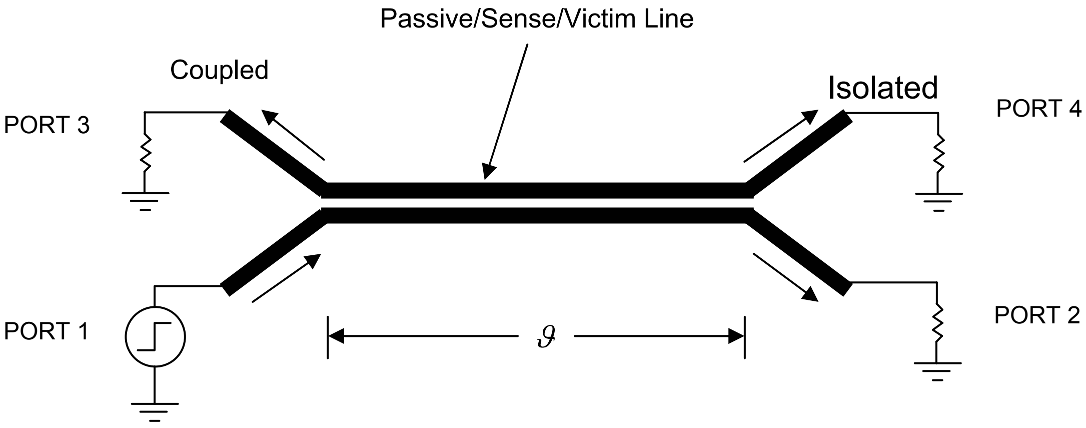
With characteristic impedance of

$$Z_o = \sqrt{\frac{L_o}{C_o}}$$

A lossy transmission line has a resistance and conductance and is dramatically effected by the materials from which it's made.

1.4 Theory

The following theoretical discussion is accompanied by validation through simulation using a Method of Moments field solver. Much of the background to the following section is supported through field theory and beyond the scope of this paper. The analytical solutions achieved are in terms of S-parameters and are calculated from the even and odd mode propagation characteristics (phase delay and impedance levels). The initial solver curve-fit a capacitance function, based upon the geometric and electrical properties of the structure, to generate the analytical solution in terms of the capacitance function. This solution was then normalized to the capacitance of the geometric components, and calculations were performed to get the propagation characteristics. Those results were then compared to that of a separate numerical field solver and found to correlate very well. From that point, the curve fitting technique was not used since the accuracy of the numerical solver was far greater; therefore, that solver was used for these solutions.



Some commonly used terms are:

- S_{11} (RL or Return Loss at coupler input)
- S_{21} (IL or Insertion Loss)
- S_{31} (NEXT or Near-End Cross Talk or TEM directional coupler output)
- S_{41} (FEXT or Far-End Cross Talk or distributed directional coupler output)

The ***electrical length*** of a transmission line coupler for the primary mode is:

$$\vartheta = \frac{2 \cdot \pi \cdot f_c \cdot \ell}{\upsilon_p}$$

where ℓ is the physical coupling length of the lines, f is the operating frequency, and υ_p is the mean phase velocity of the signal.

Velocity of propagation or phase velocity is a widely used term in electromagnetics, but the propagation delay is a more applicable quantity. This value is the reciprocal of the velocity of propagation. The signals in a transmission line are slowed by the square of the dielectric constant. For example, the most common printed circuit board dielectrics have a dielectric constant of approximately 4. Therefore, common printed circuit board delays are about 67 nanoseconds per meter.

Signals propagating through microstrip lines have two mode velocities: even and odd. The even mode phase velocity is the speed at which the wave travels in the board, and the odd mode velocity is in the air. Knowing the dielectric constant of air is much lower than that of the board, we are assured the odd mode velocity being much faster than the even.

The odd and even mode propagation delays are different due to the fact that microstrip utilizes two dielectric materials: air and the substrate. For example, the odd mode propagation delay between microstrip lines is less than that of the even mode propagation delay. This is because in the former case, more of the signal propagates in air.

The wavelength of the difference wave between the even and odd modes is the point at which maximum coupling occurs from one line to the other between two parallel, transmission lines is

$$\lambda_d = \frac{\pi}{\beta_e - \beta_o}$$

where β_o and β_e are the odd and even mode propagation constants of the wave, respectively. Knowing that $\beta = \omega/c = \omega/\upsilon_p$, then

$$\lambda_d = \frac{\pi}{\frac{\omega}{\upsilon_e} - \frac{\omega}{\upsilon_o}}$$

Where υ_e and υ_o are the even and odd mode phase velocities or velocities of propagation. We can replace velocity of propagation with delay and write the equation as

$$\lambda_d = \frac{\pi}{\omega(\tau_e - \tau_o)} \qquad [0]$$

For the six inch Golden Standard board used in this paper, the length used is only the section where the lines are parallel to each other. This length is 5.25 inches or 0.13335 meters. We have assumed the launch length, and diagonal feed lines are negligible to the coupling.

Maximum coupling for S_{31}, measurement at Port 3 relative to Port 1 is referred to as the TEM mode and is mostly due to inductive coupling. This occurs at $\vartheta = \frac{\pi}{2}$ and is normally where a coupler is operated. This is considered to be the

midband of the coupler.

1.4.1 Matching Impedance

The matching impedance for a lossy or lossless directional coupler is the square root of the even and odd mode impedances.

$$Z_k = \sqrt{Z_e \cdot Z_o}$$

For our Golden Standard board after simulation, the field solver calculated even and odd mode impedances to be **Z_e = 69.7Ω and Z_o = 45Ω**.

The resulting simulated matching impedance for our Golden Standard board is then 56Ω.

In a lossy medium, you will have the same matching impedance as the lossless providing the losses are relatively small relative to the characteristic impedance of the line.

1.4.2 S_{31} Coupling Level

For this particular coupler, the maximum coupling occurs at a quarter wave

$$f = \frac{1}{4 * \tau * \ell}$$

The coupling equation and value for this board using the previously simulated even and odd mode impedances is

$$k = \frac{Z_e - Z_o}{Z_e + Z_o} = 0.215$$

Converting to decibels using

$$C = -20\log|k|$$

we have a coupling constant of

$$C = -13.3\ dB$$

This simulated value is shown in Figure 6.

1.4.3 Maximum S_{31} Coupling Frequency

After simulation, we acquired values of even mode propagation delay = 5.98ns/m and odd mode propagation delay = 5.59 ns/m results in a mean delay of 5.78 ns/m. For the lossless line, this yields a frequency for maximum coupling in S_{31} of:

f_c= 324 MHz and (n +1) multiples [0]

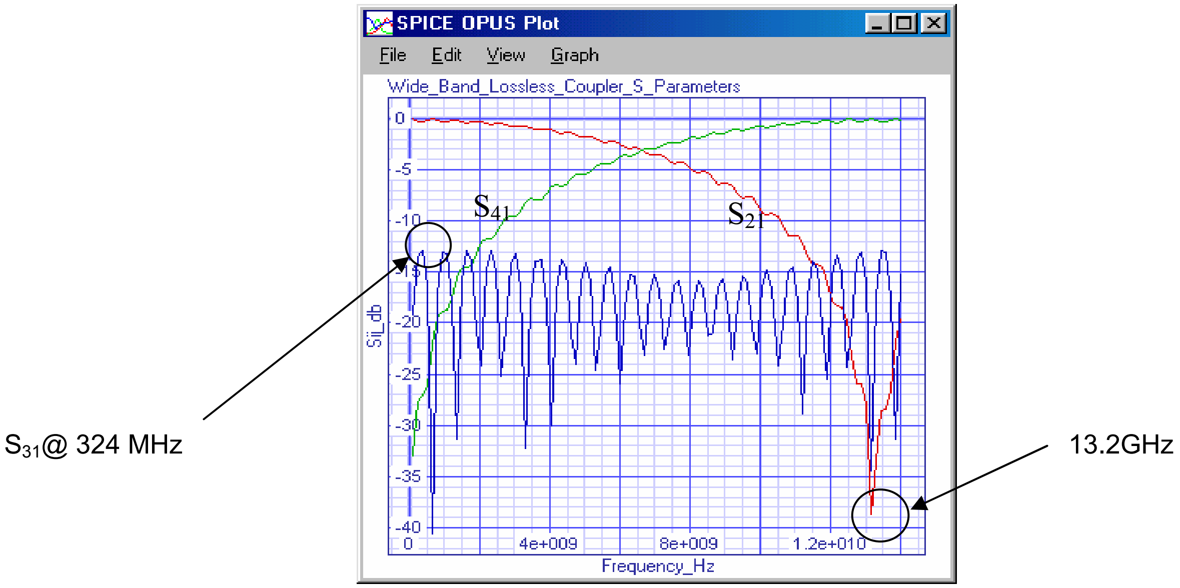


Figure 6: Simulated Wide Band Lossless Curves

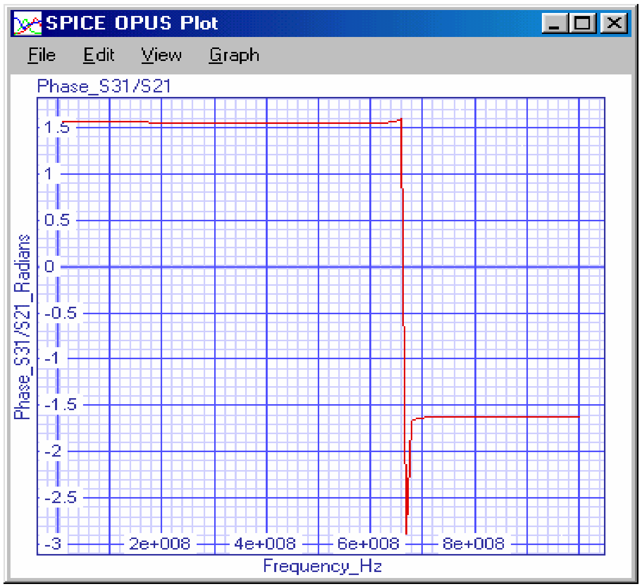


Figure 7: Simulated Phase change occurrence at maximum coupling frequency

The phase plot is shown to verify the frequency of least coupling for S_{31}. You will notice this is the point of deep absorption for S_{31}. This point does not tell us much other than just a known occurrence. The “glitch” is a simulation discontinuity and can be ignored.

1.4.4 S_{41} Coupling

For S_{41}, commonly known as the broadband or distributed coupler, the result for the electrical length is different than that of S_{21} or the TEM mode. The resultant phase velocity is not the mean, but rather the difference between the even and odd mode velocities.

$$\vartheta = \frac{2 \cdot \pi \cdot f_c \cdot \ell}{\upsilon_e - \upsilon_o}$$

As can be seen, the denominator is the difference between the modes and results in a much higher electrical length.

We first ask ourselves, at this particular length, what is the frequency by which maximum power is coupled between the lines?

First we must determine the propagation delay for the even and odd modes; we used a Method of Moments field solver for the results as the calculations are beyond the scope of this paper. In the case of the lossy model, an FR4 dielectric constant of 4.0 and loss tangent of 0.02 was used.

For the ***lossless*** coupled microstrip lines, the simulator determined the even and odd mode time delays to be:

For $\tau_e = 5.67558$ ns/m $\tau_o = 5.39457$ ns/m

Using these time delays, we calculate the frequency for maximum power transfer to be:

$f_c = 13.615 GHz$ (lossless line)

For the ***lossy*** coupled microstrip lines, the simulator determined the even and odd mode time delays to be:

For $\tau_e = 5.88178$ ns/m $\tau_o = 5.42947$ ns/m

Again, using these time delays, we calculated the frequency for maximum power transfer on S_{41} to be

$$f_c = 8.289 \text{ GHz} \quad \text{(lossy line)}$$

As you can recall, there are two modes of propagation: even and odd. The odd mode simulation, in air, will always be the same between the lossy calculation and the lossless calculation. It will be the even mode calculation that will generate the time delay change and therefore contribute to the shift downward in frequency as this is the mode that travels within the medium. See the Appendix for the lossy simulation report.

Knowing at what frequency maximum power is transferred, we can determine a few of the S-parameters using the following equations from [0].

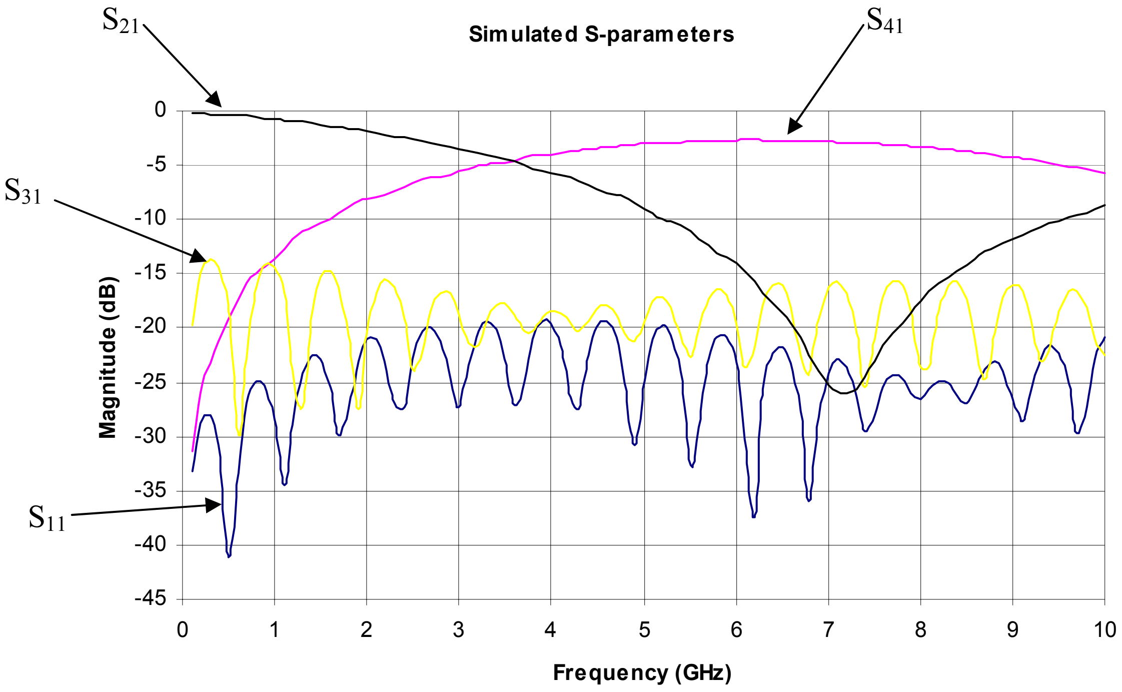
$$|S_{41}| = \left|\sin\left(\frac{\pi \cdot \ell}{2 \cdot \lambda_d}\right)\right| \qquad |S_{21}| = \left|\cos\left(\frac{\pi \cdot \ell}{2 \cdot \lambda_d}\right)\right|$$

S_{41} becomes a maximum and S_{21} becomes a minimum when the resultant wavelength of the even and odd modes of the coupler is equivalent to the physical length. In a lossless medium, S_{41} would continue to reach a maximum for every 2n+1 multiples of $\pi/2$. As S_{41} is at a maximum, S_{21} is at a minimum. For the lossy model, S_{41} will continue to oscillate, but it will be attenuated as the frequency increases due to ohmic and primarily dielectric losses. The two equations denote the energy transfer between the active line and the far-end cross talk on the quiet line.

1.5 Lossy Test Data

1.5.1 Lossy Test Board Frequency Simulations

All of the following frequency domain simulations were performed using a SPICE simulator that was integral to our simulation software.



The above graph shows all four, simulated scattering parameters for our reference board. You will notice the very similar comparison between the lossless simulation and the lossy simulation. We used a loss tangent (dissipation factor) of .02 and dielectric constant of 4.0 for our model.

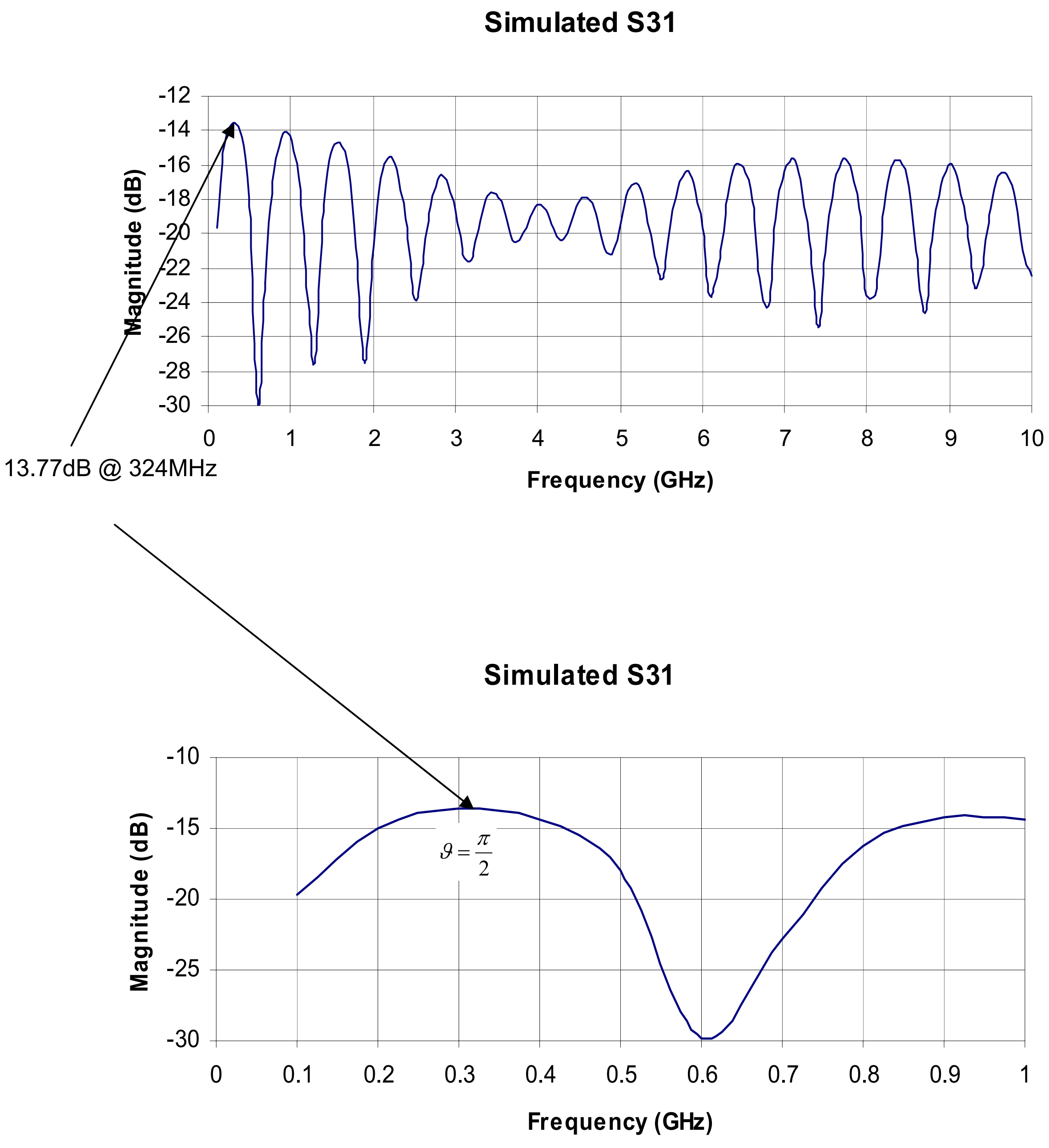
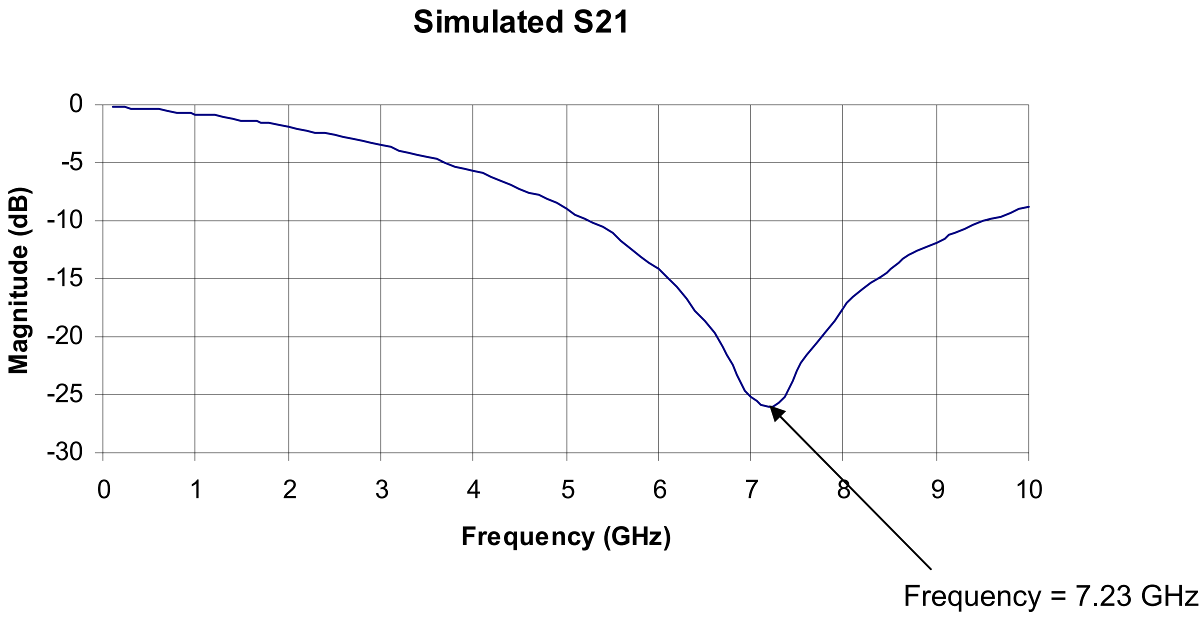


Figure 8: Simulated TEM (S_{31}) curve at 324MHz

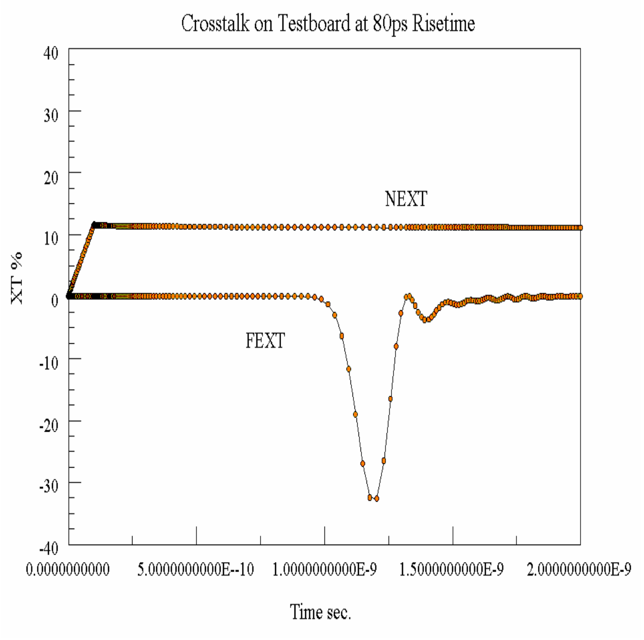
Recall the simulated S_{31} is the Near-End Cross Talk (NEXT) and is governed by the mean of the propagation delays and hence the much lower frequency.



This is the frequency by which maximum coupling occurs and is governed by the delta between the propagation delays.

With the SPICE simulated S_{21}, you will notice a discrepancy between the calculated frequency of 8.2GHz versus the SPICE frequency of 7.23 GHz. There are many areas where differences can occur. The numbers that are used for this calculation are extremely small, on the order of 10^{-9}, so any round off error of any type will create a shift in the solution. There are innate numerical errors that occur within models, and when using many of these models together for simulation can attribute to these differences. This is commonly seen with different software programs having different solutions to the same problem.

1.5.2 Lossy Test Board, Single-Ended, Time Domain Simulations



The NEXT is approximately 12% while the far-end FEXT is 32%-the result of modal dispersion on the test board. The output step exhibits a rise time of ~250ps for an input rise time of 80ps; a rise time degradation of 312% which is a result of the combination of the attenuation and the modal dispersion. 3D simulations in the time domain (or IFT) should produce similar results since time domain simulations are less sensitive to the effects of 3D mode conversion to waveguide modes etc.

1.6 Measured Test Data

1.6.1 Test Setup

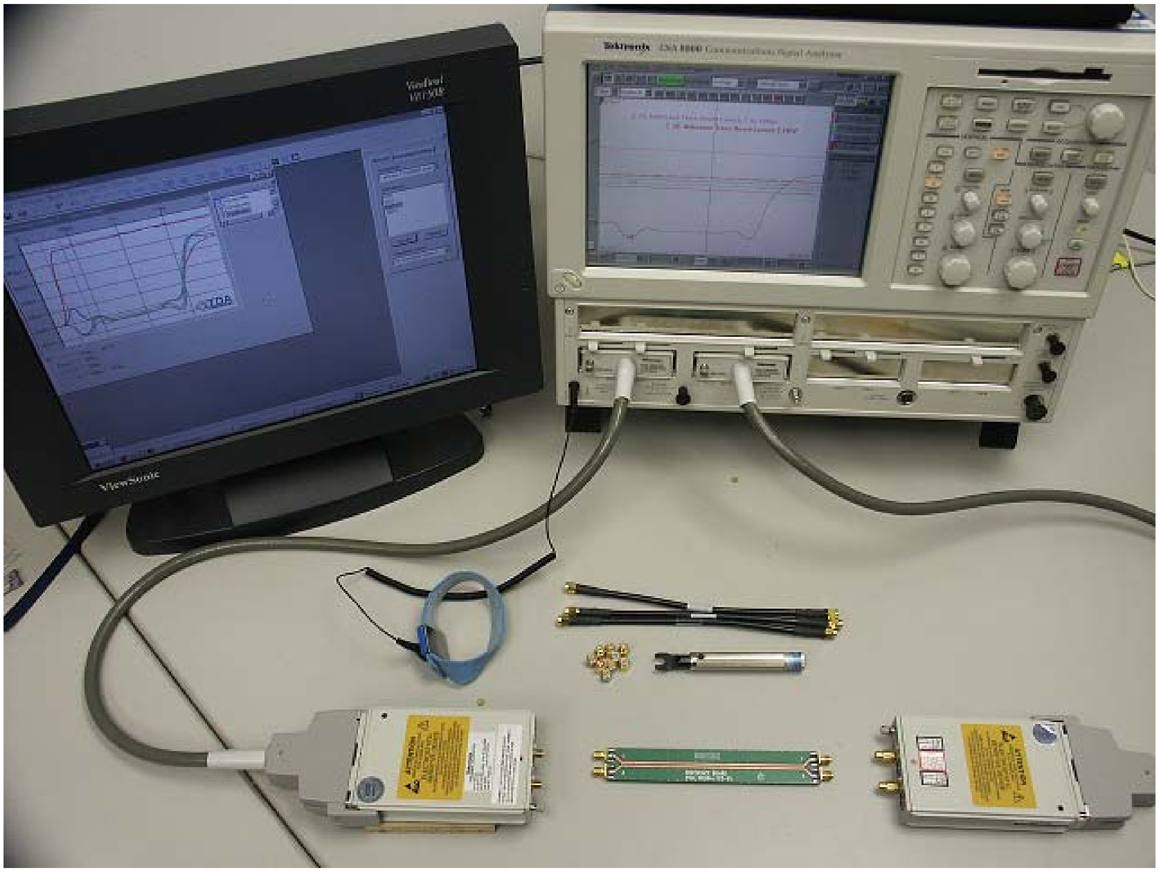


Figure 9: Test Setup

- Tektronix CSA8000 communication signal analyzer
- Tektronix 1 meter module extender model # 012-1568-00
- Tektronix sampling module model # 80E04
- Iconnect software
- 50Ω coaxial cables of equal length with male SMA connectors
- Samtec reference board
- Loads and shorts for board
- ESD wrist strap
- Calibrated SMA torque wrench

Figure 10: Samtec Golden Standard Board

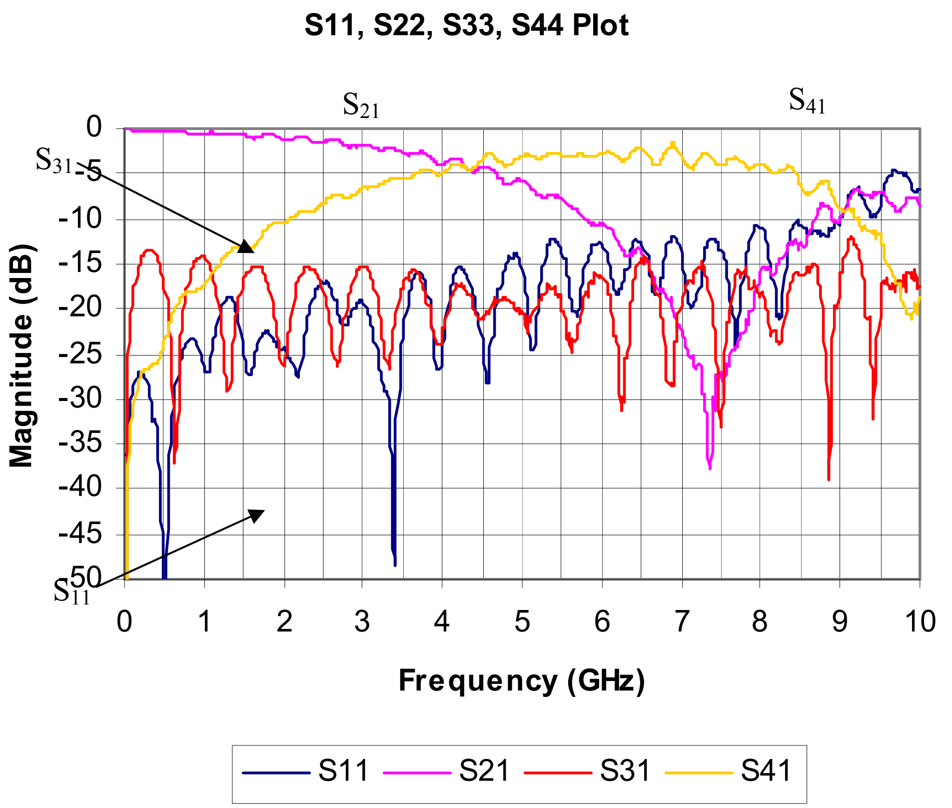
The Samtec Golden Standard board is made out of FR4 substrate material with an overall thickness of .062”. The copper is one ounce copper. The traces can be modeled as an array of conductors with the outer two conductors being ground and the internal two as signals. The bottom ground plane is tied to the upper ground plane with the use of many vias. A via fence or stitching is performed along the length of the traces and around the edge of the board. This is to ensure a solid ground plane with negligible difference in ground potential. The end launch SMA connectors are Johnson Components-part number 142-0701-851. The initial reference board had only a few vias near the SMA launch points to tie the grounds together. The difference between the grounding schemes can have an unknown effect.

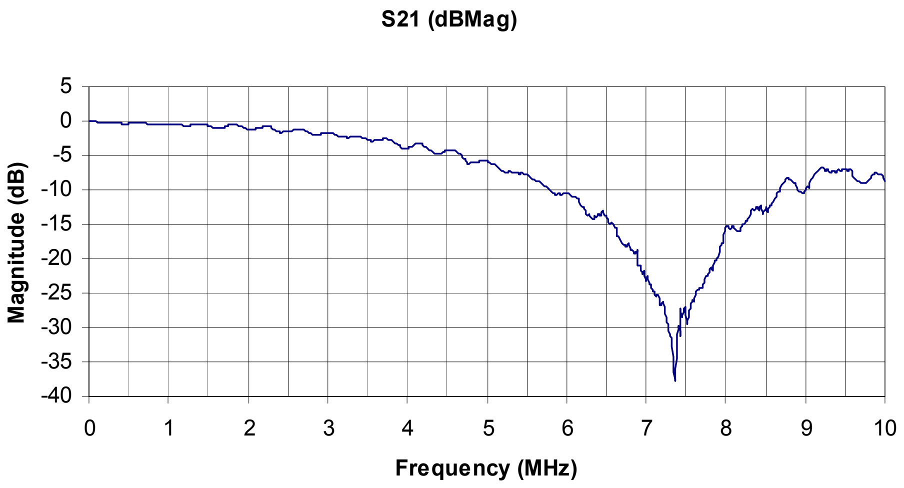
The initial Golden Standard board was built using a milling machine. This was done so the traces would be as close to a true rectangle as possible to match the theoretical calculations. This procedure is not realistic for sample boards using mass production, so the Golden Standard boards were created through a commercial board house. The standard board house fabrication process creates traces that are more similar to trapezoids than true rectangles, and this complicates theoretical calculations. This must be considered when comparing the theoretical results with the real test data.

A soldermask was placed on the produced boards, but only on the ground on the top layer and over most of the bottom layer. The soldermask was refrained from being placed over the traces as the loading effect of the mask would have dramatically changed the response of the coupler. An anti-oxidant, such as gold, was not deposited over the traces, but rather they were kept completely uncovered. The traces will oxidize, and some consider the oxidation to be a ceramic of sort, but for our purposes, it is best kept uncovered.

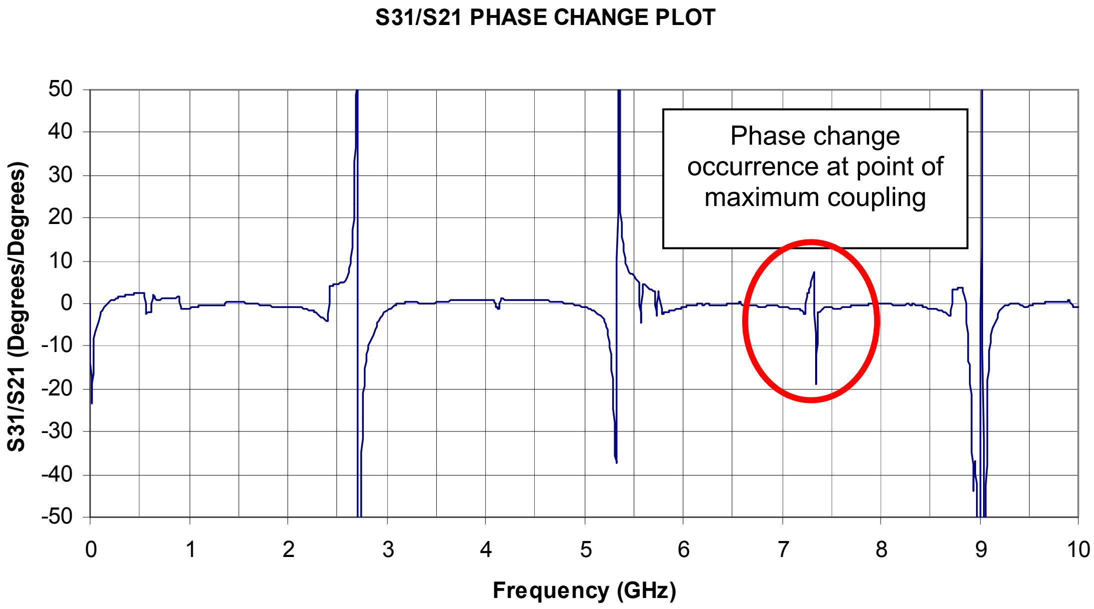
1.6.2 Measured Frequency Domain Measurements

We first took our S-parameter measurements in our setup, and all are shown below. Initially, you can see the graph is very similar to the simulated graph earlier in this paper. Notice the large coupling for S_{41} during the very low coupling for S_{21}. Also notice the point of maximum coupling for S_{31}.

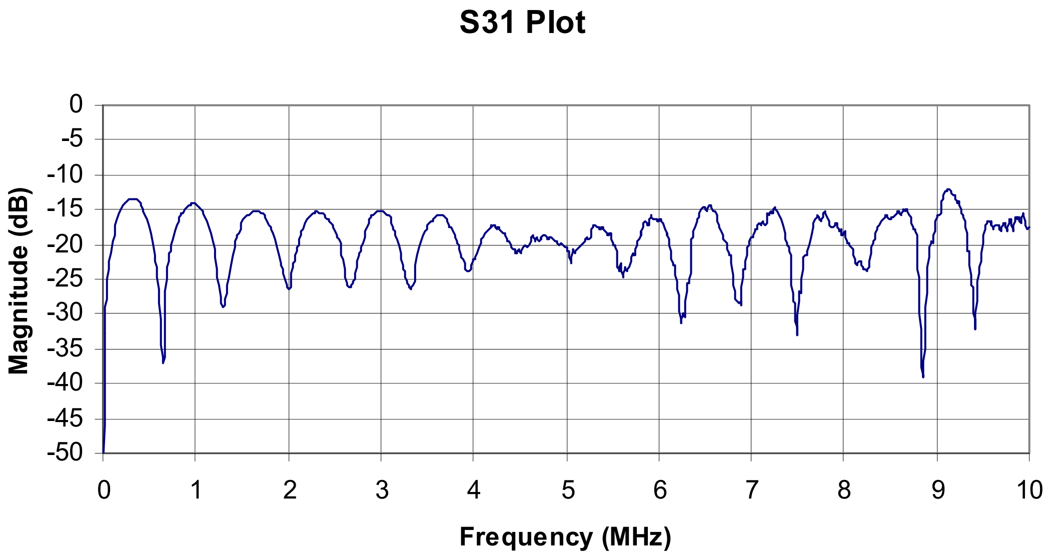


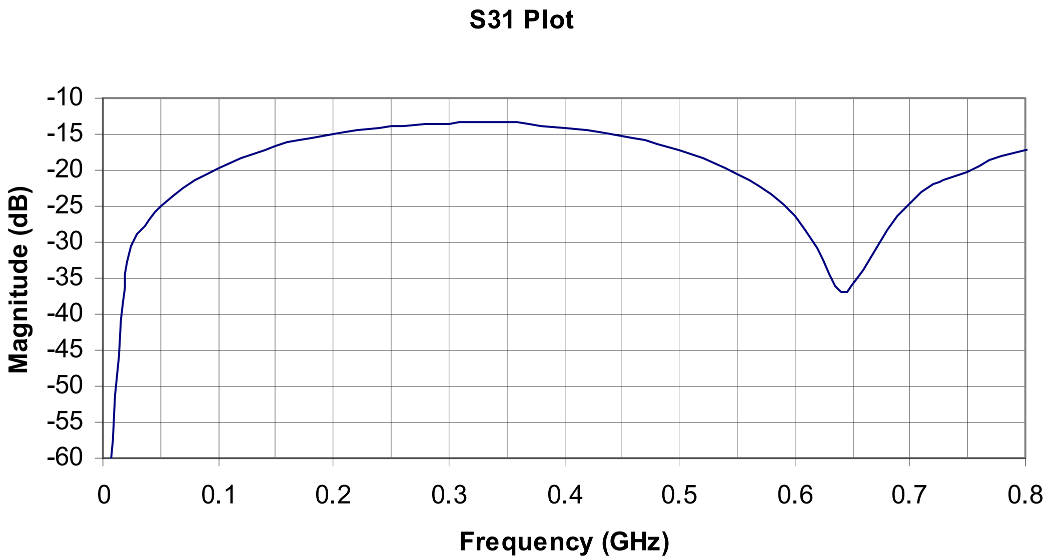


As you can see in, the frequency where minimum coupling occurs for S_{21} at 7.3GHz versus the 7.3GHz that was found theoretically. Even though the data matched the simulated value perfectly, this is not always the case. Many errors can be introduced that create discrepancies between the simulated and measured results. The reader must be aware of how the software performs its calculations in order to make the best decision for the design.



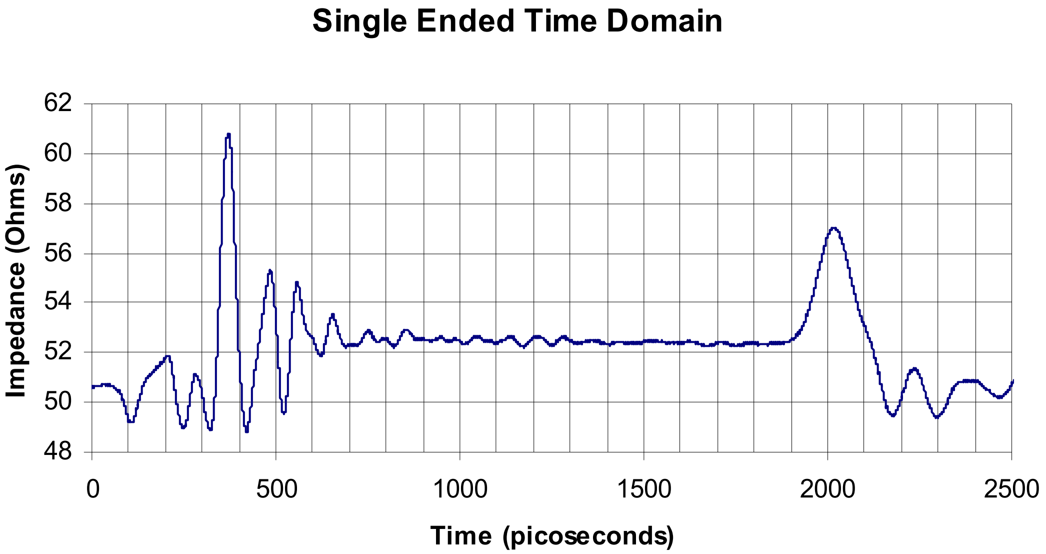
The phase change above at 7.3 GHz is shown simply to indicate a phase change occurring at the point where S_{21} becomes a minimum and S_{41} becomes a maximum. This data is included to simply show the phase change and nothing more.





The S$_{31}$ plot above shows an enlarged section for the TEM coupling. It compares extremely well to the 324MHz TEM coupling frequency that was theoretically determined. The measured value was 320MHz, and the measured power level of -13.41dB compares extremely well to the -13.3dB that was calculated.

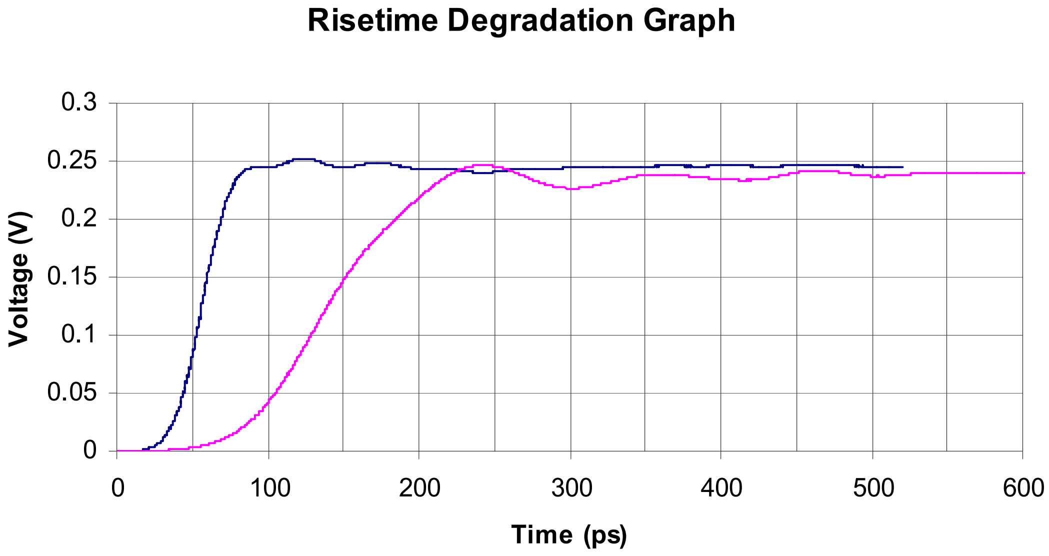
1.6.3 Measured Single-Ended Time Domain Responses of the Lossy Test Board to a 37ps Rise Time Input Step



The graph above shows the single-ended time domain measurement for impedance looking into Port 1. All ports were terminated with 50Ω, and the signal has been average 128 times. The impedance of the line is about 52.5Ω which compares well to our calculated 56Ω.

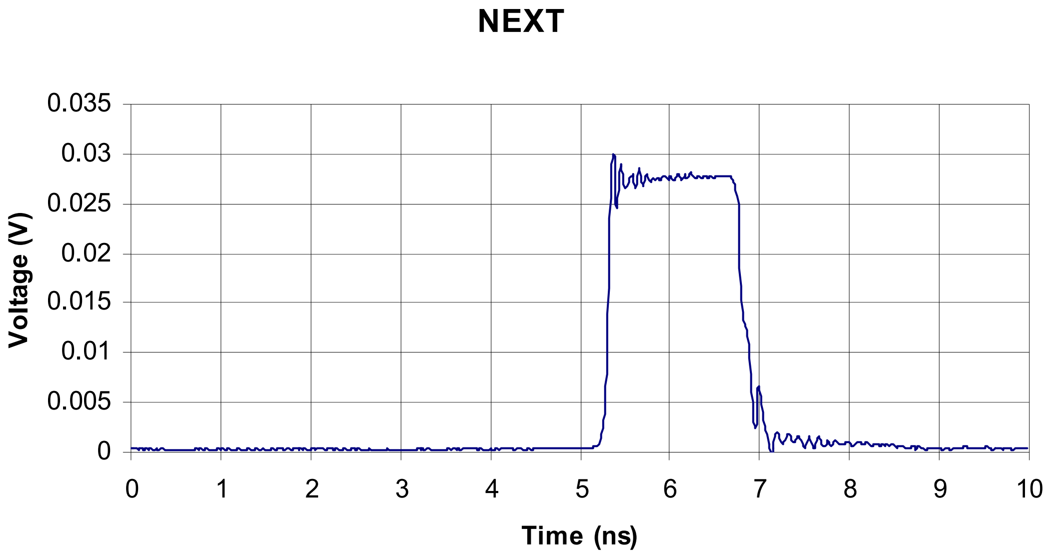
When mounting the SMA connector, there is a small gap between the board edge and where the inner conductor leaves the Teflon dielectric. This small length is where the inner conductor is not soldered to the pad and is represented by the very large spike in the graph. The second, smaller spike is the diagonal trace leading to where the two lines become parallel. After which, there is some ringing before the measurement settles and is where we measure our impedance. The last large spike is the diagonal trace leading to port 2.

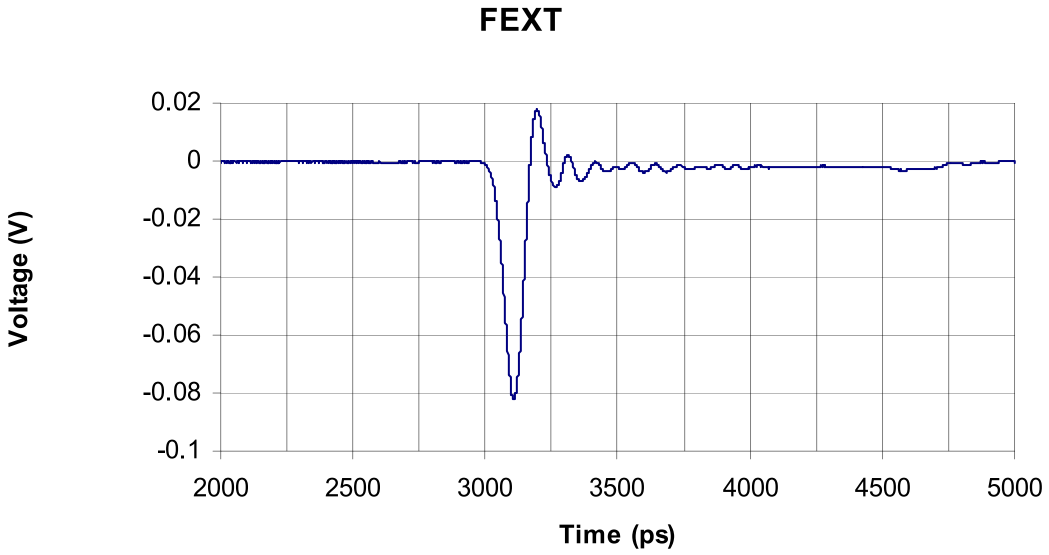
1.6.4 Measured Rise Time Degradation



From the graph, you can see the rise time degradation from Port 1 to Port 2 due to the lossiness of the board. The input rise time is around 37 ps, and the output rise time is around 116 ps which correlate to a 309% degradation. This also compares very well to the simulated values. This degradation is occurring over the course of around 6" of trace length. If the reader has concerns of the rise time of their digital signal and the bit error rate of their system, they must be very concerned with the degradation through their traces.

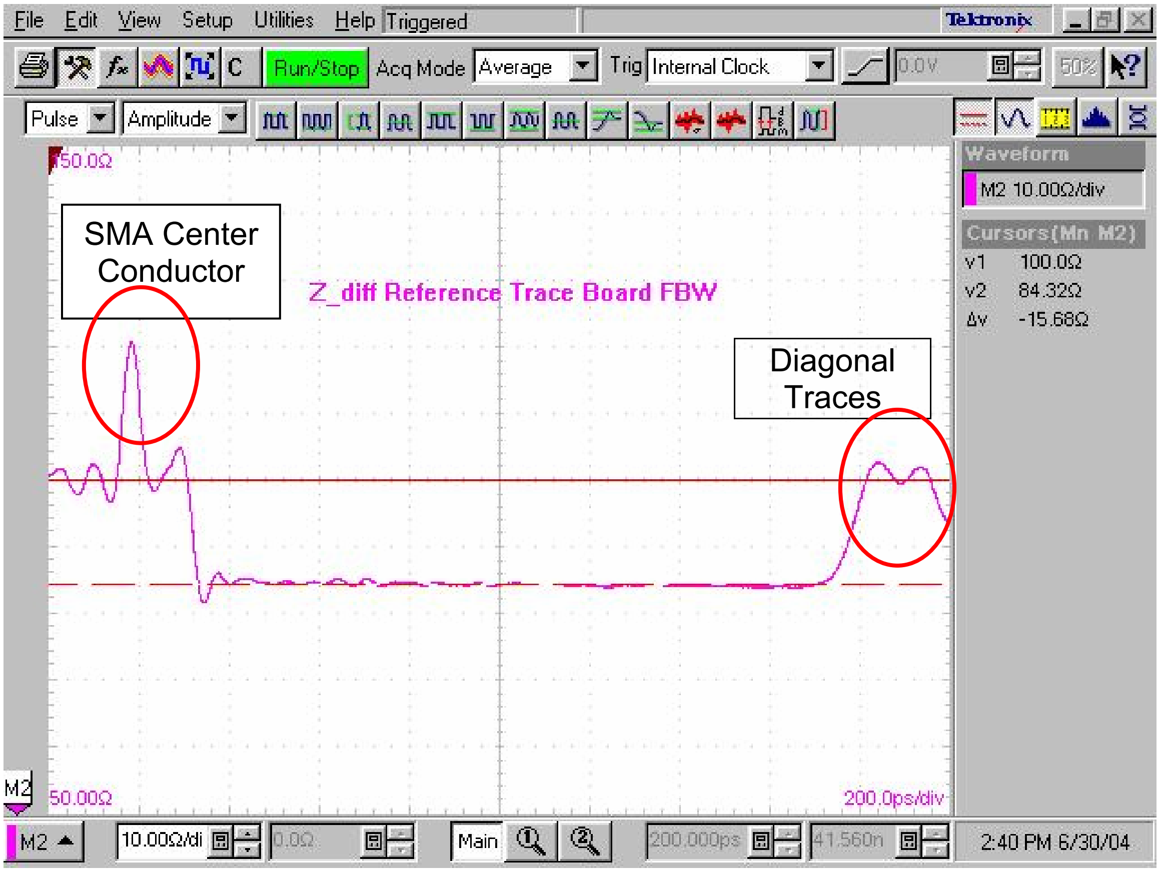
1.6.5 Measured NEXT and FEXT





The input signal had a voltage level of 258.5 which was used for the percentage calculations. The near-end cross talk for the board was approximately 27mv or 9%, and the far-end cross talk was around 82mV or 31.7% as is shown in the graphs.

1.6.6 Differential Time Domain Response with a Rise Time of 37ps.



Acq Mode Average
Trig Internal Clock
M2 10.00Ω/div
Cursors(Mn M2)
v1 100.0Ω
v2 84.32Ω
Δv -15.68Ω

The above graph shows the differential time domain response where the input signal was sent into Ports 1 and 3 and output at Ports 2 and 4. The large spike is where the center conductor of the SMA is not soldered to the board, and therefore is in air. We found the simulated differential impedance to be 86.6Ω which compares very nicely with the 84.3Ω for the measured. The effects of attenuation on the differential impedance are not as pronounced as it is for single-ended due to the coupling.

1.7 Conclusion

Through the course of this article, we have given a basic electromagnetic background, theoretical discussion, and real measured data of a four port microstrip coupler. The intent was to shed light and to clarify some confusing points about some very important test requirements using a very simple, predictable microstrip configuration.

Many facets of industry are effected by cross talk, rise time degradation, loss tangents, etc., and not understanding these phenomena can result in a design not meeting requirements. Sometimes simply being aware of these issues and making some simple preventative measures can be enough to maintain signal integrity and throughput.

Simulation software is a very useful tool for determining how a system will operate, how much noise it will generate, how a signal with be changed, where connector pins and traces should be positioned, and many other factors. The software saves valuable time and money if used by the right person. It is not well advised to run blindly into design with the data that is produced by a simulation program. Familiarity with the program and manufacturing processes are invaluable when finalizing a design, and a good designer will use all his/her tools to their advantage before releasing a design.

The Golden Standard board was made to allow customers to perform their own tests while having confidence in their results. It is our hope that, after running their own TDR and network analyzer tests using the Golden Standard board and data, our customers will gain a better understanding in many of these principals. This knowledge provides better insight into what requirements are needed for faster transition into production and a more reliable, robust design.

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APPENDIX

Cross-Section Report

Written: 07/02/04 08:08:42

Cross-section name: brian.xsc

Analysis frequency: 1e+010 Hz

Conductor Number	Conductor Name
C1	m2:1_1
C2	m2:1_2

Results included in this cross-section report are:

[L] [R] [C] [G] Td,v [Z] [Y] [Ev] [Ei] Zo

[L] unit inductance matrix in H/m

	C1	C2
C1	317.79e-9	82.564e-9
C2	82.564e-9	317.79e-9

[R] unit resistance matrix in Ohm/m

	C1	C2
C1	18.983e0	2.787e0
C2	2.787e0	18.983e0

[Rdc] unit resistance matrix in Ohm/m

	C1	C2
C1	287.2e-3	0.0e0
C2	0.0e0	287.2e-3

[C] unit capacitance matrix in F/m

	C1	C2
C1	105.87e-12	-19.456e-12
C2	-19.456e-12	105.87e-12

[G] unit conductance matrix in S/m

	C1	C2
C1	110.48e-3	-16.167e-3
C2	-16.167e-3	110.48e-3

Td mode time delay vector in s/m

v mode speed of propagation vector in m/s

Mode number	Time Delay	Speed of propagation
1	5.42947e-9	184.180e6
2	5.88178e-9	170.17e6

Used for maximum coupling calculations

[Ev] voltage eigenvectors in V

	mode 1	mode 2
C1	4.6542e0	5.8338e0
C2	-4.6542e0	5.8338e0

[Ei] current eigenvectors in A

	mode 1	mode 2
C1	107.43e-3	85.708e-3
C2	-107.43e-3	85.708e-3

[Y] admittance matrix in S

	C1	C2
C1	18.887e-3	-4.1954e-3
C2	-4.1954e-3	18.887e-3

Zo characteristic impedance vectors in Ohms

Zg(grounded) 0	load = Zu	Zu(user def.)	Zc(calculated) load = Ze	Ze(effective)	Zf(floated) load = load = open
C1	52.946e0	50.0e0	54.247e0	54.303e0	55.695e0
C2	52.946e0	50.0e0	54.247e0	54.303e0	55.695e0

Coefficient Report for Conductor 0 and Conductor 1

I	j	Lij	Cij	Ze	Zo	Se	So	Fwdx	Rvsx	Se	So
From	to	(nh/in)	(pf/in)	(ohms)	(ohms)	(ns/ft)	(ns/ft)	(s/s)	(v/v)	(ps/in)	(ps/in)
1	1	8.072	2.689	54.3	-	1.724	1.655	-	-	-	143.7
1	2	2.097	0.4942	68.07	43.32	1.793	1.655	0.07997	0.2221	149.4	137.9
2	2	8.072	2.689	54.3	-	1.724	-	-	-	143.7	-

I	j	Rsij	Gij	Rdcij	Gdcij
From	to	(ohm-nsec^.5)	(mS-ns)	(ohms)	(mS) PER INCH
1	1	0.1924	0.4466	0.00729	0
1	2	0.02106	0.06536	0	0
2	2	0.1924	0.4466	0.00729	0

[Z] impedance matrix in Ohms

	C1	C2
C1	55.695e0	12.372e0
C2	12.372e0	55.695e0

Coefficient Report

Even mode = 68.066

Odd mode = 43.323

Differential = 86.646

Coupling = 30.09