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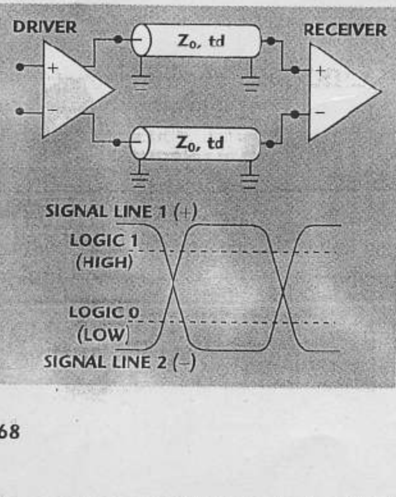
CHARACTERIZATION OF DIFFERENTIAL INTERCONNECTS FROM TIME DOMAIN REFLECTOMETRY MEASUREMENTS

Differential signaling schemes are a common approach to achieving higher noise immunity for critical signals in a high speed digital design. However, measurement and modeling of the transmission lines carrying differential signals pose several different challenges, which need to be addressed in order to achieve an accurate picture of differential signal transmission in digital system design and simulations.

A differential pair constitutes a set of coupled transmission lines and, therefore, can be modeled and simulated as such. Short differential lines can be modeled using coupled LC matrices, but a distributed model is required for longer lines. In this article, a technique for extracting such a distributed coupled line model from time domain reflectometry (TDR) measurements is presented. This model can be easily utilized in a SPICE or IBIS simulator, making it extremely usable for high speed differential interconnect modeling and simulation. The resulting accurate models help the designer to achieve a better understanding of the differential interconnects, resulting in higher performance system design.

Fig. 1 Differential signaling. ▼



DIFFERENTIAL LINE SIGNALING AND ANALYSIS

As shown in ***Figure 1***, differential signaling means that two transmission lines and two signals are used to transmit a single data bit from a driver to a receiver. The lines are not completely independent: When the signal on one line is logical low, the signal on the other line is logical high and vice versa. In addition, the two lines are typically laid out quite close to each other and exhibit signal coupling from one line to the other to a varying degree.

The reason for sacrificing precious space on a circuit board is to allow signal transmission between a driver and a receiver where a clean, reliable common ground between the driver and the receiver cannot be achieved. For example, this may be the case when the spacing between the driver and receiver is large. In such an event, the ground voltage potential of the driver circuit actually may be different from the ground voltage potential of the receiver circuit.

As an additional benefit, differential signaling schemes provide increased immunity to the common-mode noise in the system be-

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cause the receiver only sees relative (differential) voltage between the two transmission lines in the differential pair. In addition, because the fields radiated by each signal are of opposite polarity, they cancel out to significantly reduce the radiated energy, which is the main cause for electromagnetic interference between devices. Differential signals are also more immune to signal attenuation in the transmission medium because the receiver design typically allows sufficient gain to reproduce the original signal.[1] Typical applications of differential signals are low voltage differential signaling, fiber channels, disk drive flexible interconnects and Rambus™ clock signals.

A DIFFERENTIAL LINE CIRCUIT DESCRIPTION

The characteristic impedance of a transmission line can be described using its series resistance and inductance and shunt capacitance and conductance per unit length:

$$Z=\sqrt{\frac{R+j\omega L}{G+j\omega C}} \qquad (1)$$

This equation reduces for a lossless transmission line to

$$Z=\sqrt{\frac{L}{C}} \qquad (2)$$

The electrical length of such a line can be determined using

$$t=l\sqrt{LC} \qquad (3)$$

where

l = physical length of the line

Typically, differential lines are routed fairly close together. Because of the interaction (coupling) between the lines, propagation of the signal through the differential pair cannot be described by a single capacitance and inductance value per unit length, but instead is described as a set of L and C matrices per unit length:

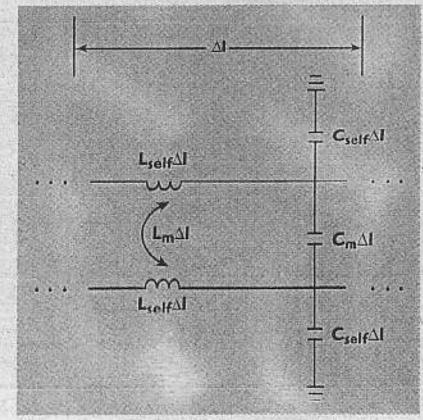
$$L=\begin{bmatrix} L_{self} & L_m \\ L_m & L_{self} \end{bmatrix}$$

$$C=\begin{bmatrix} C_{tot} & -C_m \\ -C_m & C_{tot} \end{bmatrix} \qquad (4)$$

where

$C_{tot} = C_{self} + C_m$

As shown in ***Figure 2***, these quantities are related to a physical circuit by an electrically short section (Δl) of a transmission line in which C_{self} is the



▲ *Fig. 2 An electrically short section Δl of a transmission line.*

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capacitance per unit length of one line to ground and C_m is the mutual capacitance per unit length between lines. The quantities L_{self} and L_m are the self-inductance per unit length of one line and the mutual inductance per unit length between lines, respectively.

But what about the differential and common-mode impedances of the lines, which are often used to describe the differential transmission line behavior? Differential impedance is typically defined as the impedance measured between two conductors driven differentially, that is, with identical, but opposite, polarity signals. Odd-mode impedance is the impedance of a single conductor (transmission line) when the two conductors are driven differentially.[3] Even-mode impedance is the impedance of either conductor when the differential pair is driven with identical, same-polarity (even- or common-mode) signals. Various common-mode impedance definitions are used in the industry; the definition that considers the common-mode and even-mode impedances to be identical will be used in this article.

In some cases, the differential impedance alone is the parameter of interest to a board designer. Based on the differential impedance value, a designer can make a first cut at predicting the propagation of the signal through the differential pair. In addition, the common-mode impedance can help analyze the common-mode noise rejection; if the common-mode impedance is much higher than the differential impedance, the common-mode rejection will be high. If the ratio of the common-mode signal to the differential signal present on the differential transmission line pair and the values for the differential and common-mode impedances are known, the amount of common-mode noise that will propagate through the differential interconnect can be estimated.

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Odd- and even-mode impedances for a differential pair can be computed using

$$Z_{odd} = \sqrt{\frac{L_{self} - L_m}{C_{tot} + C_m}}$$
$$Z_{even} = \sqrt{\frac{L_{self} + L_m}{C_{tot} - C_m}} \qquad (5)$$

and

$$t_{odd} = l\sqrt{(L_{self} - L_m)(C_{tot} + C_m)}$$
$$t_{even} = l\sqrt{(L_{self} + L_m)(C_{tot} - C_m)} \qquad (6)$$

Using these definitions, it is easy to conclude that

$$Z_{differential} = 2Z_{odd}$$
$$Z_{common} = \frac{Z_{even}}{2} \qquad (7)$$

with the delays for differential and odd mode and common and even mode being equal. Note that in the case where no coupling between the lines in the differential pair is present, both even- and odd-mode impedance values simply collapse to the characteristic impedance of each line. Normally, coupling between the lines would be considered a negative char-

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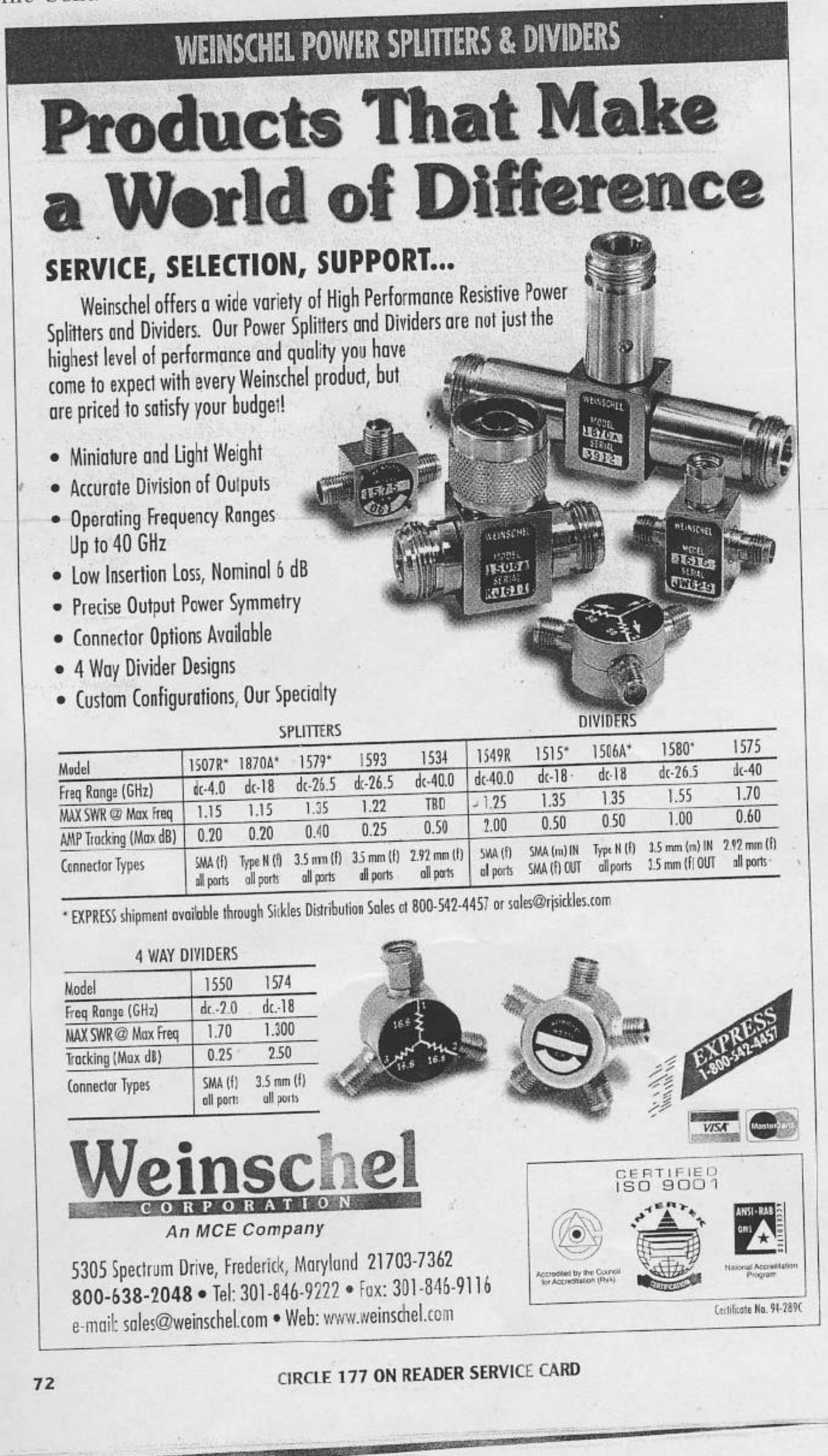
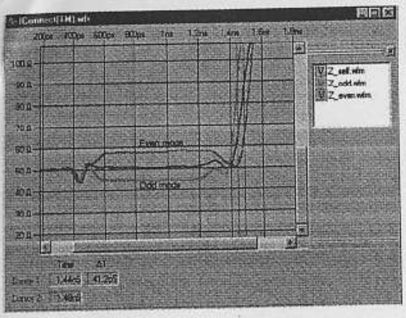




▲ *Fig. 3 Even- and odd-mode impedance profiles obtained from the differential TDR measurements.*

Fig. 4 Differential pair model based on the even and odd impedance and delay values. ▼

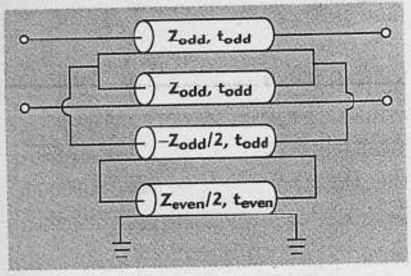
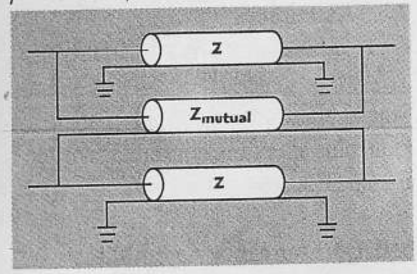


Fig. 5 A simplified differential pair model. ▼



acteristic. However, in the differential signaling case, a higher common-mode rejection actually is obtained due to coupling between the lines. Note that unless there is no coupling between the lines, the even-mode impedance will always be higher than the odd-mode impedance.

It is also interesting to observe that the time delays for the even and odd modes will be different unless the ratio of inductive and capacitive crosstalk in the two lines is the same. The difference in delays and impedances for the even and odd modes for a symmetric differential transmission line pair is shown in ***Figure 3*** using the Z-line algorithm in TDA Systems' IConnect™ software. (Note that in the presence of coupling between the two transmission lines, the even-mode impedance will always be higher.) In most practical cases, the even- and odd-mode delays will be different. Therefore, if a significant amount of common-mode energy is present in the differential signal, the designer will observe signal splitting,[4] resulting in bit errors and intersignal interference.

DIFFERENTIAL LINE SIMULATION IN SPICE

How is the propagation of the differential signal through the interconnect modeled? Since a differential pair is typically just a pair of closely routed symmetric and coupled transmission lines, it would be logical to model them as a symmetric coupled pair. A coupled LC matrix is generally used to describe such a coupled transmission line structure. However, a high number of lumped LC components will be required to simulate transmission lines that are electrically long, which is often the case when a differential transmission scheme must be used.

The electrical length of each LC network must be significantly shorter than the rise time of the signal propagating through the interconnect. A practical rule of a short or lumped interconnect can be described as

$$t_{rise} > \frac{t_{delay}}{6}$$

$$t_{rise} > \sqrt{\frac{L \bullet C}{6}} \qquad (8)$$

where L and C are the total capacitance and inductance values, respectively, for the given interconnect segment. For a rise time of 600 ps, this rule means that the lumped model can be applied to an interconnect segment no longer than 100 ps, or approximately 2/3" in FR4 board material. For a high number of long traces and for complex simulations, using the lumped model approach becomes impractical and cumbersome, slowing down the simulation and making a comprehensive signal integrity analysis more difficult to achieve.

Of course, an alternative is to use a distributed approach. Single-line impedance clearly is not enough to characterize the differential transmission line pair since line coupling must be taken into account. Differential impedance alone is not enough either since common-mode rejection and propagation must be accounted for. The solution comes from a relatively simple mathematical analysis of the differential pair, which also can be viewed as a symmetric coupled transmission line pair. The model shown in ***Figure 4***[5] is an accurate representation of a coupled transmission line pair based on its even- and odd-mode impedances.

It is a simple exercise in circuit analysis to demonstrate that this transmission line configuration will present twice the odd-mode impedance to a differential signal and two separate even-mode impedances to a common-mode signal. Since any signal can be decomposed in its differential- and common-mode components, this model will predict propagation of any combination of differential- and common-mode signals, accurately representing the behavior of the differential transmission line pair. Note that, for practical purposes, it may be preferable to avoid using a transmission line with a negative impedance; however, the line with negative impedance can be easily substituted with a positive impedance line and a dependent voltage source.

One disadvantage of this four-line model is its relative complexity. In addition, this model is difficult to extend to a case of more than two coupled lines. However, if the common-mode impedance is significantly larger than the differential-mode impedance, the common-mode signal will be mostly rejected, and mainly the differential signal will be observed at the receiver end. The high common-mode rejection will make the difference in differential- and common-mode signal delay irrelevant due to the small amplitude of the common-mode signal, and will allow the use of a simplified model, shown in ***Figure 5***. Here,

$$Z_{mutual} = \frac{2Z_{odd}Z_{even}}{Z_{even} - Z_{off}} \qquad (9)$$

odd ?

Again, under assumption of $t_{odd} = t_{even}$, it can be easily shown that this model will accurately represent both differential- and common-mode signal propagation through the differential pair. This article focuses on the measurement-based approach to extracting the even and odd imped-

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ces for the distributed differential ne models presented previously.

OBTAINING A DIFFERENTIAL LINE MODEL FROM MEASUREMENTS

The modeling of a differential line from measurements is shown in ***Figure 6***. The choice of instrumentation for performing this measurement-based modeling work is limited to frequency domain instruments such as vector network analyzers (VNA) or impedance analyzers and time domain instruments such as the TDR.

For designers with significant microwave background, the frequency domain is quite often a more understood and more intuitive domain to work with. However, a problem arises from the fact that differential network analyzer measurements, or four-port measurements, require a multiport network analyzer system. It is only recently that such a system has become available from major measurement equipment manufacturers. A two-port VNA measurement system can be used to obtain four-port network parameters, but this approach requires a large number of measurements[6] and is not easy to complete or simple to analyze.

On the other hand, four and more ports in the TDR instruments have been readily available for quite some time. A TDR may or may not support a comprehensive frequency domain calibration procedure available in most VNA systems, but for purposes of extracting the even and odd impedances and delays the accuracy of the measurement is more than sufficient. Moreover, the propagation delay is more easily obtained in the time domain due to the visual nature of this domain.

TDR MEASUREMENT BASICS

In the simple TDR setup shown in ***Figure 7***, the impedance of the board trace can be determined from the waveform measured by a TDR oscilloscope. The measured waveform is the superposition of the incident waveform at the device under test (DUT) and the reflected waveform, with the reflected waveform offset by two electrical lengths of the cable interconnecting the oscilloscope TDR sampling head to the DUT.[7] The multiple reflection effects must be deconvolved to achieve better accuracy in impedance measurements. A TDR measurement setup for differential line characterization is shown in ***Figure 8***.

Differential TDR measurements can come in handy when it is difficult to achieve a good ground plane reference, or when a differential line analysis must be performed. A virtual ground plane, created by two TDR sources of the same shape and different polarity that arrive simultaneously at a DUT interface, helps achieve the desired measurement results.

It was mentioned previously that TDR measurement accuracy suffers from multiple reflection effects when multiple discontinuities are involved in the measurement. However, a true impedance profile of the DUT can be obtained through an inverse scattering algorithm reported previously.[8,9] Based on the incident step and TDR response of the system, the multiple reflections can be dynamically deconvolved from the TDR response; because of that process, another name used for this algorithm is dynamic deconvolution.

Even- and odd-mode analysis, based on the even and odd impedance profiles computed from the differential TDR measurements, is an extremely useful tool for characterizing symmetric transmission line systems, such as cables and connectors. The odd impedance profile is obtained from a differential TDR measurement with two TDR sources of opposite polarity; the even impedance profile is obtained through a measurement with two TDR sources of the same polarity. In each case, only a single TDR channel needs to be acquired. A differential reference short is used for computing the impedance profile. The reference short waveform is most easily obtained by disconnecting the DUT and connecting the two signals to ground in close proximity to each other, or connecting them directly to each other and then connecting them to ground. Once the even and odd impedance profiles of the differential pair are obtained, the model can be easily computed. ?

In addition, the LC matrices can be easily extracted from the even and odd TDR impedance profiles using ①

$$L_{self} = \frac{1}{2\Delta l}\left(Z_{even}t_{even} + Z_{odd}t_{odd}\right)$$

$$L_{mutual} = \frac{1}{2\Delta l}\left(Z_{even}t_{even} - Z_{odd}t_{odd}\right)$$

$$C_{mutual} = \frac{1}{2\Delta l}\left(\frac{t_{odd}}{Z_{odd}} - \frac{t_{even}}{Z_{even}}\right)$$

$$C_{tot} = \frac{1}{2\Delta l}\left(\frac{t_{odd}}{Z_{odd}} + \frac{t_{even}}{Z_{even}}\right) \qquad (10)$$

From a practical modeling perspective, the capacitor and inductor values from Equation 10, for example,

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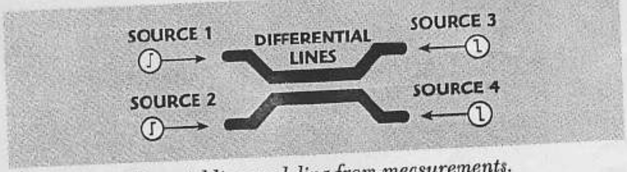
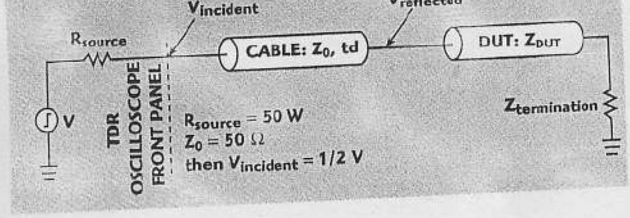
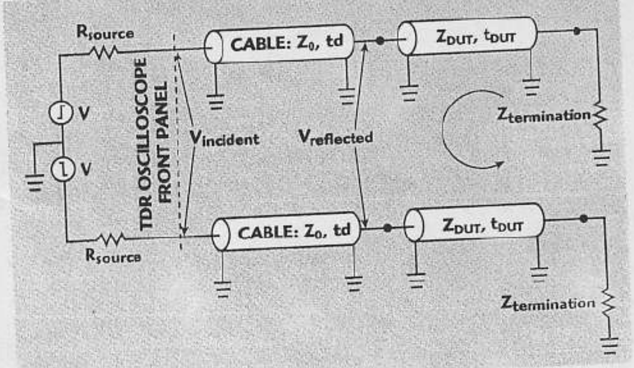


▲ *Fig. 6 Differential line modeling from measurements.*

▼ *Fig. 7 A TDR block diagram.*





▲ *Fig. 8 TDR measurement setup for differential line characterization.*

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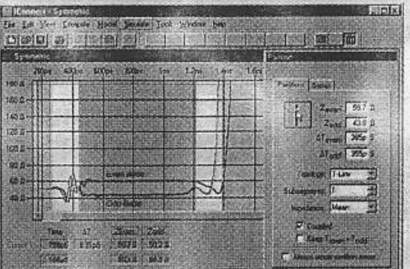
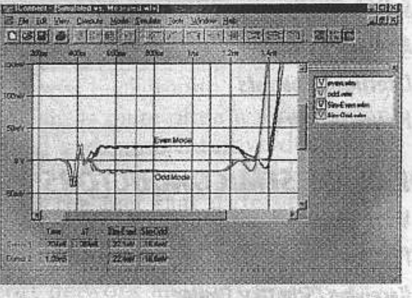


▲ *Fig. 9 Partitioning the impedance profile waveforms in the symmetric coupled line modeling window.*

Fig. 10 Verifying the differential line model. ▼



$L_{self}\Delta l$, may be computed if a lumped model is desired. The above equations also demonstrate that if the mutual capacitance $C_m\Delta l$ is to remain positive, it is necessary that

$$\frac{t_{odd}}{Z_{odd}} - \frac{t_{even}}{Z_{even}} > 0 \qquad (11)$$

A model that does not satisfy this constraint should not be used since it does not represent a physically realizable structure and could produce inaccurate simulations.

A DIFFERENTIAL LINE MODELING EXAMPLE

As an example, a differential pair on an FR-4 board was measured and modeled. The lines in this DUT have SMA connectors as an interface to a TDR oscilloscope. The lines are closely coupled. (The spacing to the ground plane on the board was half the gap between the lines.) The impedance profiles for the same differential pair were shown previously.

After the data were acquired from a TDR oscilloscope, they were processed using the Z-line impedance deconvolution algorithm in the IConnect software. As mentioned previously, two waveforms are necessary to compute the impedance profile: the DUT waveform and the reference step waveform. The reference step waveform for a differential measurement can be obtained by connecting the signals on both of the differential TDR channels to each other and then to ground, if possible. When using cables with SMA connectors, the easiest way to achieve this measurement is to connect the cables using an SMA barrel interconnect. Only the waveform on one channel of a TDR instrument needs to be acquired; no additional adding or subtracting of the waveforms in the scope is necessary. The resulting even and odd impedance profiles were shown previously and, based on these impedance profiles, the model for the DUT is easily extracted. Note that without the impedance deconvolution algorithm, the impedance profiles are subject to the multiple reflection effects in TDR oscilloscopes and the impedance readout values may not be correct at each point on the TDR trace.

After the impedance profiles have been computed, the impedance profile waveforms are partitioned in the symmetric coupled line modeling window, as shown in ***Figure 9***. The distributed model is the most appropriate in this case since the electrical length of the lines approaches or exceeds 1 ns. The assumption that the even-mode delay is equal to the odd-mode delay is somewhat difficult to maintain, and the more accurate four-line model is more appropriate. When the model is saved, the equivalent SPICE circuit that describes the model is obtained. A sample listing of such a circuit is given in ***Appendix A***.

To verify the created model, a designer must create a composite model using the IConnect software. The composite model complements the extracted DUT model with the source and termination that emulate the TDR measurement source and termination. Using an integrated interface to a SPICE simulator, the designer can simulate this composite model and, based on the resulting simulation waveforms, verify the accuracy of the DUT model. Both even- and odd-mode stimuli must be used in simulations to ensure that the model accurately predicts both even and odd modes of signal propagation, as shown in ***Figure 10***. (Simulations

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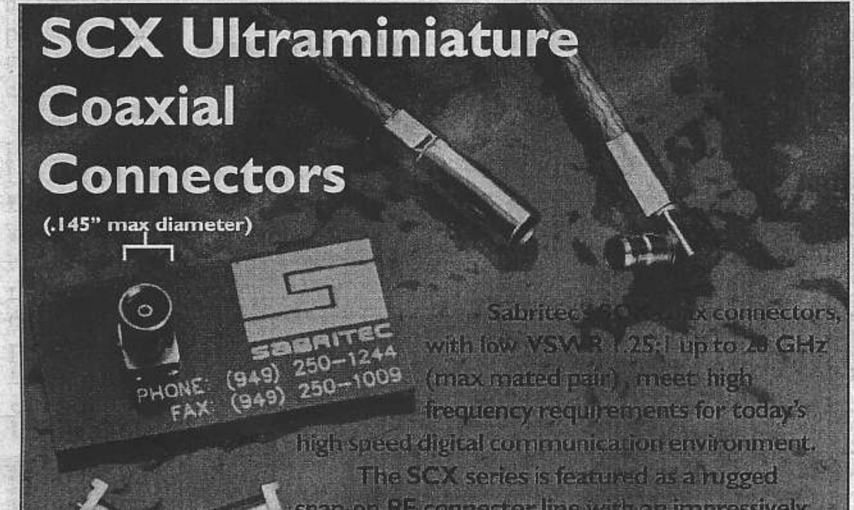

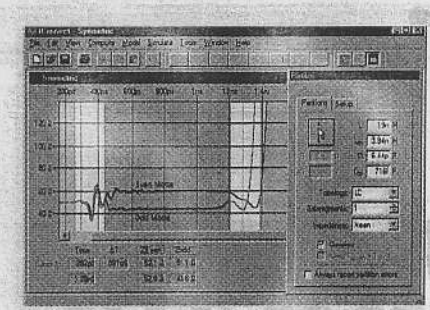

▲ *Fig. 11 The even and odd analysis used to compute LC matrices for the coupled lines.*

with both even- and odd-mode stimuli are performed using an integrated interface to a SPICE simulator and the simulation results are compared to the measured data.) It can be seen that with the exception of the SMA connectors, which are not part of the differential line, the model accurately predicts the signal propagation. In addition, the connectors can be modeled using the IConnect software's lumped circuit modeling capability.

A simplified three-line model also can be used to model the differential transmission lines if the difference between the even- and odd-mode delays can be ignored. In this model, the even-mode delay must be used for the main lines, and the delay for the line responsible for the coupling in the structures must be adjusted to achieve good match between the simulation and the measurement.

Even- and odd-mode analysis also can be utilized for characterization of lumped interconnect structures, even when single-ended signaling schemes are utilized. Structures such as high speed connectors, ball grid array packages and high performance automatic test equipment sockets can be easily modeled using even and odd TDR measurements and Equation 10. Based on the even and odd impedance profiles, the LC matrices for the coupled structure are easily computed, as shown in ***Figure 11***. (Note that for long lines a large number of subsegments must be used to accurately model the line.)

CONCLUSION

A technique for extracting a distributed coupled line model from TDR measurements has been demonstrated. This model can be easily used in any standard time domain simulator (SPICE or IBIS) and can accurately predict the propagation of a digital signal through a differential transmission line pair. As a result, the digital system simulation will predict the system behavior more accurately, resulting in higher performance system designs. ■

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APPENDIX A

SAMPLE SPICE CIRCUIT LISTING

```
* Syntax: PSpice
* Name: Automatically Generated
.subckt Symmetric 1 2 3 4 5
****** Partition #1
***** Subsegment #1 *****
   t1   1   5   6   5   Z0=49.7  TD=92.3p
   t2   3   5   7   5   Z0=49.7  TD=92.3p
****** Partition #2
***** Subsegment #1 *****
   t3   6   8   9   10  Z0=43.8  TD=345p
   t4   7   8   11  10  Z0=43.8  TD=345p
   e1   12  8   12  13  2
   e2   14  10  14  15  2
   t5   12  13  14  15  Z0=21.9  TD=345p
   t6   13  5   15  5   Z0=29.9  TD=385p
****** Partition #3
***** Subsegment #1 *****
   t7   9   5   2   5   Z0=44.4  TD=74.7p
   t8   11  5   4   5   Z0=44.4  TD=74.7p
.ends
```