IEEE Standard for Backplane Electrical Performance

Sponsor

Microprocessor and Microcomputer Standards Subcommittee of the IEEE Computer Society

Approved April 5, 1991

IEEE Standards Board

Abstract: The proper treatment of the electrical elements of the backplane that provides a physical and electrical connection between different modules in a computer system is covered. Techniques for measuring impedance, capacitance, and crosstalk are provided; and crosstalk, ground bounce, and decoupling are discussed. Relationships among the key elements controlling backplane performance, such as driver and receiver characteristics, the distance of the driver and receiver from the backplane, the impedance of the bus traces treated as transmission lines, lumped elements such as connectors and vias, and the termination are examined.
Keywords: backplane, backplane bus, bus, high-performance backplanes, low-performance backplanes, measurement techniques, medium-performance backplanes, transmission lines

The Institute of Electrical and Electronics Engineers, Inc.
345 East 47th Street, New York, NY 10017-2394, USA

Library of Congress Cataloging in Publication Data will be found on the following page.

IEEE Standards documents are developed within the Technical Committees of the IEEE Societies and the Standards Coordinating Committees of the IEEE Standards Board. Members of the committees serve voluntarily and without compensation. They are not necessarily members of the Institute. The standards developed within IEEE represent a consensus of the broad expertise on the subject within the Institute as well as those activities outside of IEEE which have expressed an interest in participating in the development of the standard.

Use of an IEEE Standard is wholly voluntary. The existence of an IEEE Standard does not imply that there are no other ways to produce, test, measure, purchase, market, or provide other goods and services related to the scope of the IEEE Standard. Furthermore, the viewpoint expressed at the time a standard is approved and issued is subject to change brought about through developments in the state of the art and comments received from users of the standard. Every IEEE Standard is subjected to review at least every five years for revision or reaffirmation. When a document is more than five years old, and has not been reaffirmed, it is reasonable to conclude that its contents, although still of some value, do not wholly reflect the present state of the art. Users are cautioned to check to determine that they have the latest edition of any IEEE Standard.

Comments for revision of IEEE Standards are welcome from any interested party, regardless of membership affiliation with IEEE. Suggestions for changes in documents should be in the form of a proposed change of text, together with appropriate supporting comments.

Interpretations: Occasionally questions may arise regarding the meaning of portions of standards as they relate to specific applications. When the need for interpretations is brought to the attention of IEEE, the Institute will initiate action to prepare appropriate responses. Since IEEE Standards represent a consensus of all concerned interests, it is important to ensure that any interpretation has also received the concurrence of a balance of interests. For this reason IEEE and the members of its technical committees are not able to provide an instant response to interpretation requests except in those cases where the matter has previously received formal consideration.

Comments on standards and requests for interpretations should be addressed to:

Secretary, IEEE Standards Board
445 Hoes Lane
P.O. Box 1331
Piscataway, NJ 08555-1331
USA

Library of Congress Cataloging in Publication Data

Institute of Electrical and Electronics Engineers.
IEEE standard for backplane electrical performance / sponsor, Microprocessor and Microcomputer Standards Subcommittee of the IEEE Computer Society ; approved April 5, 1991, IEEE Standards Board.
p. cm.
"IEEE Std. 1194-1991."
Includes bibliographical references and index.
ISBN 1-55937-127-7
1. Backplanes (Microcomputers)—Standards—United States. I. IEEE Computer Society. Microprocessor and Microcomputer Standards Subcommittee. II. IEEE Standards Board. III. Title.
TK7895.B33I58 1991 91-17328
621.39' 16—dc20 CIP

Foreword

(This Foreword is not a part of IEEE Std 1194-1991, IEEE Standard for Backplane Electrical Performance.)

This standard was prompted by the conflicting performance claims of various backplane implementations. The claims were made with differing conditions, making true comparisons of performance difficult. The definitions and techniques presented in this standard provide a guideline upon which these comparisons can be made. This standard is not a backplane benchmark, nor is it meant as a tutorial on transmission line theory. Rather, it is a practical guide to the problems associated with measuring backplane electrical performance.

Special thanks go to those who contributed the most toward making this standard possible. R. V. Balakrishnan served as chairman in the first few months and got the standard effort started. Jeff West served as the draft editor, writing most of it and helping review the ballot comments. Joel Martinez handled most of the administrative duties as the Working Group secretary, contributed to several sections of the standard, and helped review the ballot comments. Dan Yaklin helped review the ballot comments. Paul Borrill made technical contributions in several areas. Clyde Camp, Chair of the Microprocessor Standards Committee, helped tremendously, answering my "What needs to be done next?" questions.

At the time this document was submitted for sponsor ballot, the members of the Working Group were as follows:

Rick Henderson, *Chair*

R. V. Balakrishnan
Harrison Beasley
Paul Borrill
Rick Curtis
Emil Hahn
Anatol Kaganovich
Suren Kodical
Joel Martinez
Thanos Mentzelopoulos
Jeff West
Dan Yaklin

The following persons were on the balloting committee that approved this standard for submission to the IEEE Standards Board:

Mohammad Al-Malki
Duane L. Anderson
Jack Arabian
R. V. Balakrishnan
Harrison A. Beasley
John Black
Paul L. Borrill
Lyle Burnett
Todd Busniewski
Clyde Camp
Jay Cantrell
Donald Chi
Paul D. Cook
James Davis
Wayne Fischer
Gordon Force
Willard Graves
David Hawley
David V. James
Anatol Kaganovich
Richard Karcich
Ezzat Korany
Takahiko Kuki
Tuvia Lamdan
Glen Langdon
Gerry Laws
Rollins Linser
Andy J. Luque
Tulin Mangir
William McDonald
John Montague
Cuong Nguyen
Chandresh J. Patel
James M. Pexa
Shlomo Pri-Tal
William T. Rhoades
Hans Roosli
Fred U. Rosenberger
Philip Shutt
Michael Smolin
Paul Sweazey
Serge Szukalow
Darius Tanksalvala
Michael Teener
Michael Thompson
Carl Warren
Hans A. Wiggers
Mark Williams
Oren Yuen

The final conditions for approval of this standard were met on April 5, 1991. This standard was conditionally approved by the IEEE Standards Board on December 6, 1990, with the following membership:

Marco W. Migliaro, *Chairman* **James M. Daly,** *Vice Chairman*
Andrew G. Salem, *Secretary*

Dennis Bodson
Paul L. Borrill
Fletcher J. Buckley
Allen L. Clapp
Stephen R. Dillon
Donald C. Fleckenstein
Jay Forster*
Thomas L. Hannan
Kenneth D. Hendrix
John W. Horch
Joseph L. Koepfinger*
Irving Kolodny
Michael A. Lawler
Donald J. Loughry
John E. May, Jr.
Lawrence V. McCall
L. Bruce McClung
Donald T. Michael*
Stig Nilsson
Roy T. Oishi
Gary S. Robinson
Terrance R. Whittemore
Donald W. Zipse

*Member Emeritus

Contents

SECTION PAGE

FIGURES

FIGURES PAGE

TABLE

IEEE Standard for Backplane Electrical Performance

1. Scope

The backplane provides a physical and electrical connection between different modules in a computer system. Each module communicates with other modules by way of the backplane bus. The wires or traces on the bus and the traces on the modules are more than just a common connection point for electrical signals; they are electrical elements. Proper treatment of the electrical elements of the backplane (such as capacitance, inductance, impedance, delay constants, termination, connectors, driver, and receiver characteristics) is essential if the design of the backplane is to be successful.

This standard covers the proper treatment of the electrical elements of the backplane. It contains measurement techniques for impedance, capacitance, and crosstalk; a discussion of crosstalk, ground bounce, and decoupling; and a study of the relationships between the key elements that control backplane performance. Some of these key elements are driver and receiver characteristics, the distance of the driver and the receiver from the backplane, the impedance of the bus traces treated as transmission lines, lumped elements such as connectors and vias, and the termination. Each element contributes to the performance of the backplane.

The performance of a backplane can be categorized into three groups, as identified in Chapter 6 of [B3]:[1]

(1) Lumped model for low-performance backplane designs
(2) Transmission line reflected wave switching for medium-performance backplanes
(3) Transmission line incident wave switching for the highest-performance backplane designs

The lumped model treats the backplane as a capacitor equal to the total capacitance along the signal path, backplane, and plug-in modules. The lumped model can be used where the rise time of the signal is slow compared to the signal propagation time down the backplane, or where the bus performance is not critical and only a final state at some later time is needed. The lumped capacitance is charged or discharged by the driving device or the termination and is controlled by an RC time constant. The signal transition from a low to a high is of the form $1-e^{-t/R_1C}$, and a high-to-low signal transition is of the form e^{-t/R_2C}, where R_1 and R_2 are the equivalent resistance of the charging and discharging device.

As an example, Fig 1 shows the equivalent bus schematic of the open collector bus analyzed as a lumped model, where the open collector driver releases the bus and the voltage, $v(t)$, makes a transition from low to high.

[1]The numbers in brackets correspond to those of the bibliography in Section 7.

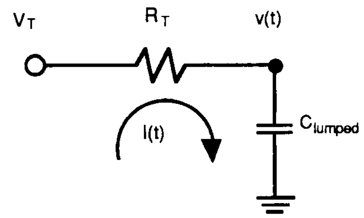

Fig 1
Low-to-High Lumped Model for Analysis

V_T is the termination voltage, C is the total lumped capacitance, and R_T is the termination resistance. Analysis of the circuit gives the following:

$$I(t) = C\frac{dv(t)}{dt} = \frac{V_T - v(t)}{R_T} ,$$

or

$$\frac{dv(t)}{dt} + \frac{v(t)}{R_T C} = \frac{V_T}{R_T C} .$$

This is a first-order differential equation and has a solution of the form

$$v(t) = V_T\,(1 - e^{-t/R_T C}) .$$

When the driver asserts the line low, the equivalent diagram for analysis of the behavior of the backplane is shown in Fig 2. R_d is the series resistance of the driver when it is in the ON state.

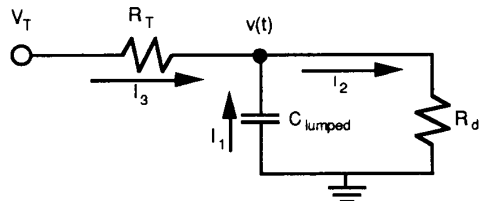

Fig 2
High-to-Low Lumped Model for Analysis

Analysis of this circuit gives the following:

$$I_2 = I_3 + I_1,$$

$$\frac{1}{R_d}v(t) = \frac{V_T - v(t)}{R_T} - C\frac{dv(t)}{dt},$$

$$C\frac{dv(t)}{dt} + \frac{1}{R_d}v(t) = \frac{V_T - v(t)}{R_T} ,$$

or

$$\frac{dv(t)}{dt} + \left(\frac{1}{CR_d} + \frac{1}{CR_T}\right)v(t) = \frac{V_T}{CR_T} ,$$

and has a solution of the form

$$v(t) = V_T\left(\frac{R_d}{R_d + R_T} + \frac{R_T}{R_d + R_T}\, e^{-t/RC}\right)$$

where

$$R = \frac{R_T R_d}{R_d + R_T}\ .$$

By knowing the lumped capacitance of the backplane and the value for the resistances charging or discharging the backplane, the performance of the backplane can be analyzed for time and voltage.

If the signal transition times are short compared to the time a signal propagates down the bus, then it is better to analyze the backplane as a transmission line. That is,

$$\text{transmission line length} > \frac{T_R\,(\text{ns})}{2T_{PD}\,(\text{ns/ft})}\ ,$$

where T_R is the signal rise time at the source, and T_{PD} is the one-way propagation delay of the line per unit length.

General transmission line properties are found in texts on the subject (see, for example, [B12], p. 490), but for transmission line analysis of backplanes only a few key concepts need be well understood. A few of these include the characteristic impedance, Z_0, in ohms; the characteristic delay, τ_0, usually in nanoseconds per meter; and the reflection coefficient, ρ, which is defined as the amplitude of the reflected wave divided by the amplitude of the incident wave.

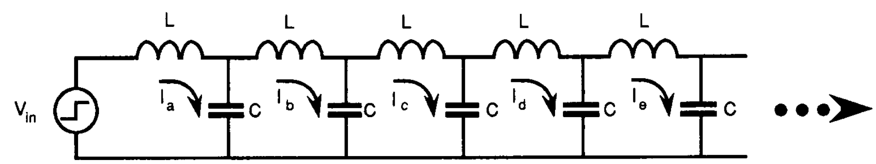

Fig 3
Simplified Discrete View of a Transmission Line

Figure 3 shows a simplified view of a transmission line as distributed inductance and capacitance. A voltage wave that is launched onto the transmission line at the left causes current to flow into the line. This current charges the distributed capacitance, but is delayed down the line by the distributed inductance. As long as the signal sees the same distributed L's and C's, the impedance of the line can be written as

$$Z_0 = \sqrt{\frac{L_0}{C_0}}\ , \qquad \text{(Eq 1)}$$

where L_0 and C_0 are the distributed inductance *(L)* and capacitance *(C)* per unit length. The current flowing into the transmission line is

$$I = \frac{V_{in}}{Z_0}\ .$$

The time it takes for the signal to travel down the transmission line is calculated using the relationship, delay = distance * τ_0, where

$$\tau_0 = \sqrt{L_0 C_0}\ . \qquad \text{(Eq 2)}$$

On a backplane there are always connectors, which have capacitance and inductance, that connect the backplane to branch transmission lines called stubs. These branch transmission lines are on the plug-in modules and are the communication paths between the plug-in modules and the backplane.

These connectors and stubs change the impedance of the transmission line. Lumped capacitance along a transmission line, as in the case of regularly placed connectors on a backplane, changes the impedance and delay constant by the relationships

$$Z_L = \frac{Z_0}{\sqrt{1+\frac{C_d}{C_0}}} ,$$

and

$$\tau_L = \tau_0\sqrt{1+\frac{C_d}{C_0}} ,$$

where C_d is the added capacitance per unit length, and C_0 is the intrinsic capacitance per unit length of the transmission line, Z_0. These relationships can be derived by replacing C_0 by $C_0 + C_d$ in Eqs 1 and 2. (These equations do not take into account the effects at very high frequencies. In those cases, skin effects and other effects start to become significant. The assumption of frequency-independent values of Z_0, L_0, and C_0 are then no longer valid.)

A point along the transmission line where the impedance changes is called a discontinuity. An example of this occurs when the backplane is loaded with only one-half of the plug-in modules, and they are grouped to one end of the backplane. The impedance of the backplane has a discontinuity at the interface between the unloaded and loaded sections. At each point where a voltage wave traveling down a transmission line meets a discontinuity in the impedance, some of the signal is reflected and some is transmitted. The ratio of the reflected wave to the incident wave is called the reflection coefficient, ρ. The reflection coefficient has a major influence on the design and performance of a backplane. If the drivers or transceivers are placed too far from the backplane, or there is too great a mismatch of impedance on the backplane or to the termination, or there are improper driver and receiver characteristics, then reflections can cause serious noise problems. These problems include multiple switching and reduced noise margins. In this case, a bus-settling time would be required before valid data could be presumed.

In high-performance backplane designs the bus impedance, termination resistance, termination voltage, stub length, and the driver and receiver characteristics are all carefully controlled so that valid data can be presumed on the incident wave of the signal.

The 1194 standard is intended to be used to design and evaluate high-performance backplanes. The standard has four sections. The first section covers measurement techniques for the impedance of a backplane and lumped capacitance placed along the backplane. These techniques are based on time-domain reflectometry. The next section deals with two types of noise that can be generated on a bus: ground bounce and crosstalk. The third section evaluates decoupling techniques, looking at both the current needed during switching and the capacitor's equivalent RLC network. The last section applies transmission line theory to worst-case bus loading configurations to derive backplane design constraints. These constraints relate the unloaded and loaded bus impedances, the termination resistance, and stub length. The purpose of this last section is to show that by knowing the desired noise margins and the characteristics of the drivers and receivers on the bus, derivations can be made that relate the key bus parameters, Z_U, Z_L, R_T, stub length, and lumped capacitance so that the desired bus performance can be achieved.

Although these derivations were based on the open collector type of bus and incident wave switching, the techniques can be applied to other bus structures. This is done by creating the Thevenin-equivalent termination and calculating the equivalent currents that charge and discharge the line. The process of creating a Thevenin-equivalent circuit for the termination, where an equivalent termination resistance and voltage are used instead of the actual circuit, is shown as an example in Section 6, Fig 20. This shows how to convert a split-resistor termination from V_{CC} to ground, to a equivalent termination resistance and voltage.

The equivalent currents for charging and discharging the line are calculated from the termination, dc-quiescent currents, and the output driver. This usually requires solving simultaneous equations. With CMOS buses, full-logic swings are not usually accomplished on the incident wave because of low output drive and low termination currents. In both the low-to-high step and the high-to-low step, the driver current must be added to the bus current to determine the total current available to charge or discharge the line. The voltage step is calculated based upon the dc current flowing through the bus, plus the current the CMOS driver is supplying at that particular V_{OH} level, and the bus impedance. With TTL buses, the high-to-low transition is very similar to the open collector case where the current to discharge the line is the change in voltage divided by the transmission line impedance, assuming there is enough available current to make the full transition. If there is not enough current, then it must be analyzed like the CMOS case. The current to charge the line from low to high is current flowing through the line while in the low state, plus any current that the Darlington structure supplies at that voltage. Standard ECL buses operate almost exactly like open collector (or IEEE Std 1194.1-1991[2] based) buses except that current is sourced onto the bus from the driver instead of sinking current from the bus when the driver is ON.

[2]As this standard goes to press, IEEE Std 1194.1-1991 is not yet published. It is, however, available in manuscript form from the IEEE Service Center, 445 Hoes Lane, P.O. Box 1331, Piscataway, NJ 08855-1331, 1-800-678-4333.

2. Conventions Used in This Standard

C_0 = The capacitance per unit length of Z_0.
C_d = The lumped capacitance per unit length added to Z_0, caused by the vias and connectors that are added to the backplane.
C_{stub} = The capacitance per unit length added to the backplane impedance by the effects of the stub or plug-in board. This includes lumped capacitance such as drivers and receivers.
L_0 = The inductance per unit length of Z_0.
ρ = The reflection coefficient. It is the ratio of the reflected wave amplitude divided by the incident wave amplitude.
R_T = The value of the termination resistor.
$V+$ = The incident voltage wave traveling on the transmission line, also labeled V_I.
$V-$ = The reflected voltage wave traveling on the transmission line, also labeled V_R.
V_{IH} = The high-level threshold voltage for the receiver.
V_{IL} = The low-level threshold voltage for the receiver.
V_{OH} = The actual high-level voltage produced by the driver on the bus.
V_{OL} = The actual low-level voltage produced by the driver on the bus.
V_T = The termination voltage.
Z_0 = The intrinsic characteristic impedance of the backplane trace. No capacitive effects of vias or connectors are included.
Z_L = The fully loaded backplane impedance. This includes the backplane trace impedance, and the capacitive effects of the vias and connectors, as well as all of the plug-in modules inserted with their vias, connectors, traces, transceiver, etc.
Z_S = The characteristic impedance of the stub between the driver and the backplane.
Z_U = The unloaded impedance of the backplane. This includes the backplane trace impedance and the capacitive effects of vias, connectors, and everything on the backplane without the plug-in modules.

incident wave switching. Describes a voltage transition that is large enough in magnitude to change the state of a input to a receiver just once on the first edge of the wave. This assumes that subsequent reflections will not cause the receiver to switch back to a previous state.

noise margin. The difference between the actual voltage on the bus and the receiver threshold voltage, V_{IH} or V_{IL}.

stub. The signal path on the plug-in board between the transceiver and the backplane. This includes the printed circuit board trace, connector, and all lumped capacitance such as drivers and receivers.

3. Measurement Techniques

3.1 Impedance Measurement Techniques. There are several methods of determining the impedance of a transmission line (backplane). The most direct method is to use time-domain reflectometry (TDR). TDR systems can measure exact impedance values and are easy to use. The user of TDR systems is referred to the TDR manufacturer's documentation and application notes (see [B11]). However, if a TDR system is not available, the backplane impedance can be derived using TDR techniques with common laboratory instruments.

The instruments needed are a high-bandwidth oscilloscope and pulse generator. The bandwidth needed should be greater than or equal to the reciprocal of 3.5 times the rise time of interest. The rise time of interest should be at least as fast as the transition times of devices used on the backplane. The procedure is to measure the incident and reflected waves caused by impedance mismatch between the unknown backplane impedance and a known impedance, and then calculate the impedance using the equations given. The known impedance is the 50 Ω test environment (50 Ω input cabling and 50 Ω input to scope).

This method should be easy to use and is very accurate. This is because the measurement system is maintained in a 50 Ω environment and thus any loading caused by the use of probes is eliminated. Also, this method most closely replicates a TDR system.

Figure 4 shows the setup for the impedance measurement. The 50 Ω input to the scope is connected to the input cable by a T-connector.

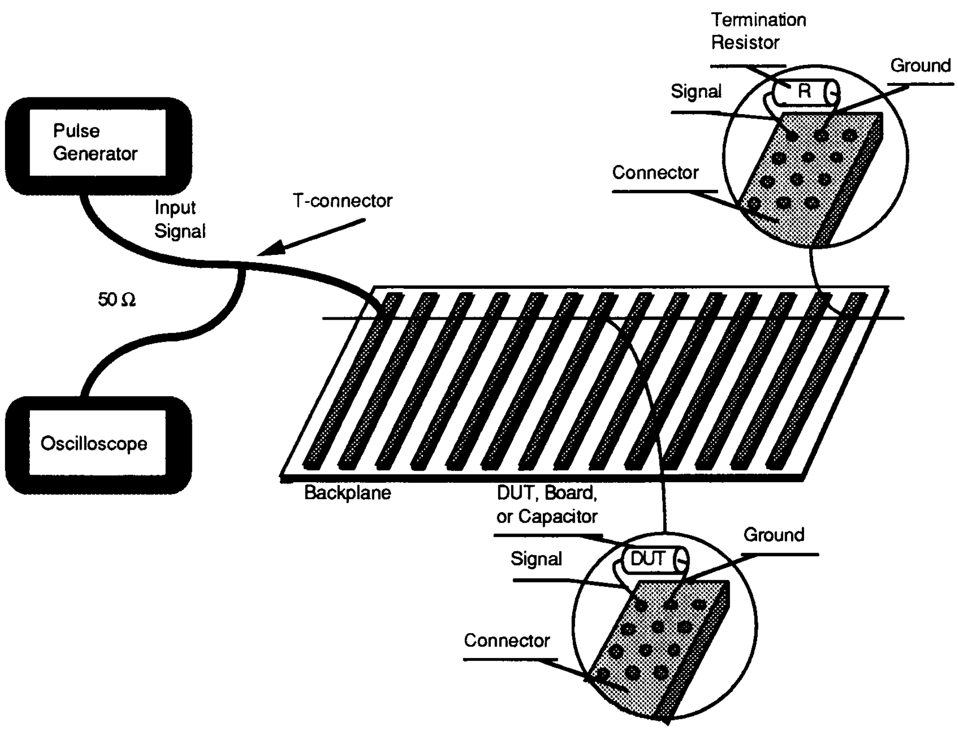

Fig 4
Test Setup for Backplane Impedance Measurement

The derivation for the impedance of the backplane being measured (Z_U) can be aided by the use of a lattice diagram (see Chapter 6 of [B3], pp. 231–235). Figure 5 shows the lattice diagram for measuring the backplane as set up in Fig 4.

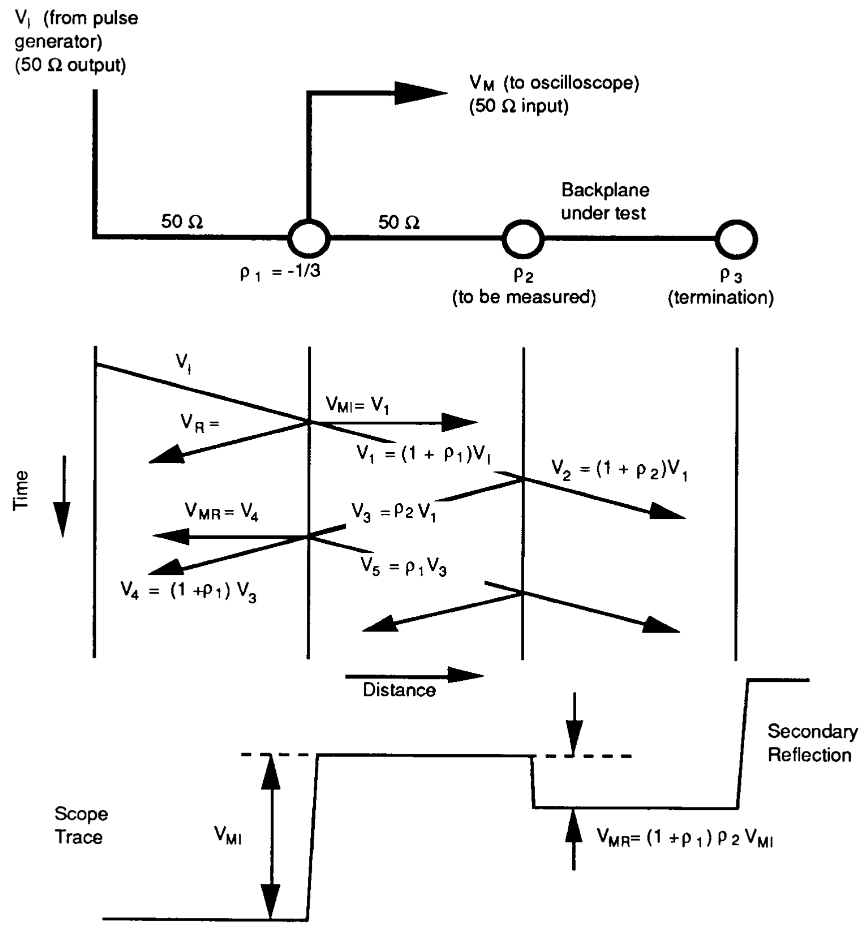

Fig 5
Lattice Diagram for Impedance Measurement

From the lattice diagram in Fig 5, the reflection coefficient, ρ_2, at the cable to backplane interface can now be derived, as follows:

$$\rho_2 = \frac{V_{MR}}{V_{MI}} * \frac{1}{1 + \rho_1}$$

where V_{MR} and V_{MI} are measured reflected and incident voltages at the oscilloscope, respectively. Since

$$\rho_1 = \frac{25 - 50}{25 + 50} = -1/3 ,$$

and the reflection coefficient to be calculated, ρ_2, equals

$$\rho_2 = \frac{Z_U - 50}{Z_U + 50},$$

the measured backplane impedance, Z_U, can now be written in terms of the measured voltages at the scope only, V_{M1} and V_{M2}, as

$$Z_U = 50\,\frac{2V_{MI} + 3V_{MR}}{2V_{MI} - 3V_{MR}}\ .$$

If the measurement for the reflected wave is the incident plus the reflected wave, as shown in Fig 6, then the impedance of the backplane can be calculated as follows:

$$Z_U = 50\frac{2V_{MI} + 3\,(V_{MR} - V_{MI})}{2V_{MI} - 3\,(V_{MR} - V_{MI})}\ .$$

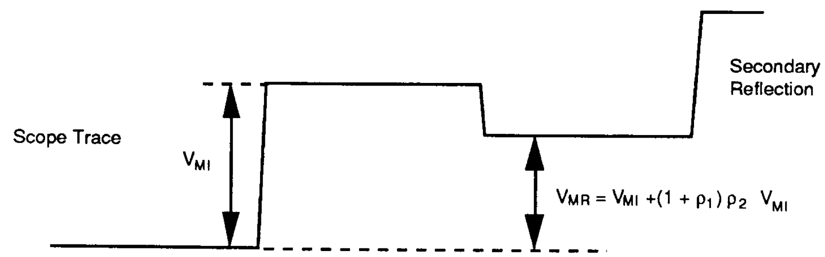

Fig 6
Measuring the Reflected as the Incident Plus the Reflected Wave

In TDR-type measurements, the input signal rise time has to be fast enough to separate the input wave (V_I) from the reflected wave (V_R) and any incidental reflections. The secondary reflection from the far end of the backplane to the termination resistor boundary would be an incidental reflection.

So that the desired reflections can be seen clearly, a rule of thumb is that the rise time must be less than 1/5 the time between reflections. The reflected wave of interest, that of the 50 Ω cable to backplane, will be seen in the round trip from the T-connector to the backplane and back. This distance is set by the cable length. The secondary reflection at the backplane to termination boundary will be seen in the round-trip delay over the entire backplane length. Typically the propagation delays are about 6.5 ns/ft on the backplane. A round-trip delay on a one-foot backplane is 2 times 6.5 ns or 13 ns. The rise time must then be 13 ns/5 or 2.6 ns or faster.

The minimum transition time on the input pulse should also be considered. In order to smooth out the reflections on the backplane created by the connectors for the plug-in cards and the distance between cards, a minimum pulse transition time of more than the two-way delay between connectors should be used. As an example, a typical τ_0 of the intrinsic backplane material might be 1.80 ns/ft. At a card spacing of 1 in, the round-trip delay between card connectors is about 305 ps. The rise time of the input pulse should then be slower than this.

3.2 Capacitance Measurement Techniques. The capacitive effect of a device or board that is to be plugged into a backplane can be determined by the method of reflection matching. This method compares the deflection on a TDR trace caused by the reflection in a wave traveling down the backplane from the device under test (DUT) with a known capacitive load. As in the case for measuring the impedance of the backplane, a TDR would be best suited for the job. However, a similar configuration for measuring lumped capacitance on a backplane can be made like that of backplane impedance measurement, again using common laboratory instruments; see Fig 7.

A voltage step, V_I, applied to the test transmission line, or backplane, shown in Fig 7, should produce a reflection of V_{RDUT} from the device under test (see Fig 8). This reflected voltage bump is equal to a voltage V_{RT} obtained from a test capacitor of equal capacitance to the device under test. The correct capacitive value is obtained by trying capacitors of different values until the same deflection is seen as with the device under test. The test capacitors should be a low-inductance type, e.g., an SMD type. This test should be carried out with incident voltage rise times of 1 ns and 10 ns (see [B1], p. 133).

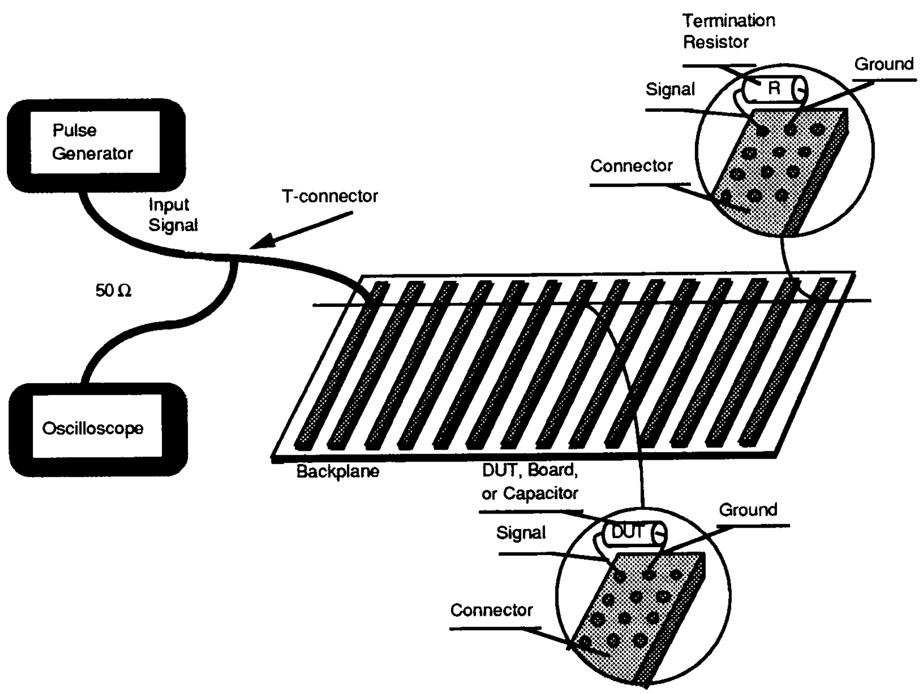

Fig 7
Test Setup for Capacitance Measurements

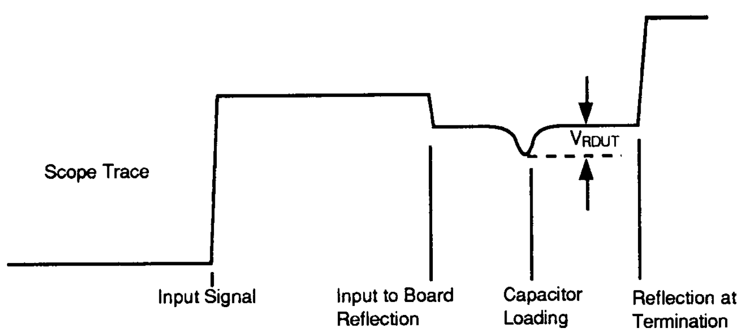

Fig 8
Voltage Deflection Caused by Lumped Capacitance

4. Noise Evaluation and Solutions

A single-ended bus is an unbalanced transmission medium that is time shared by several system elements. Like any unbalanced system, it is highly susceptible to common-mode noise. Noise is any unwanted signal or signal perturbation. Two known sources of noise are ground bounce and crosstalk.

4.1 Ground Bounce. Ground bounce is usually generated by high-speed drivers switching a lot of current in a short period of time. Ground leads and traces have inductance. The voltage generated across an inductor (ground bounce) is as follows:

$$V_L = L * \frac{dI}{dt} . \tag{Eq 3}$$

This equation shows that the faster the current is forced through an inductor, the greater the voltage across the inductor. In backplane applications it is common to use quad or octal drivers or transceivers. Since all the drivers in a package typically use the same ground lead, any inductive ground bounce created by the switching of one or more of the drivers is seen by all the outputs in the package. This ground bounce is transmitted on the data lines and reduces noise margins in the unbalanced system.

Ground bounce is often mistaken for crosstalk or noise coupling because it occurs simultaneously with the switching outputs. A simplified schematic of an output driving a transmission line is shown in Fig 9.

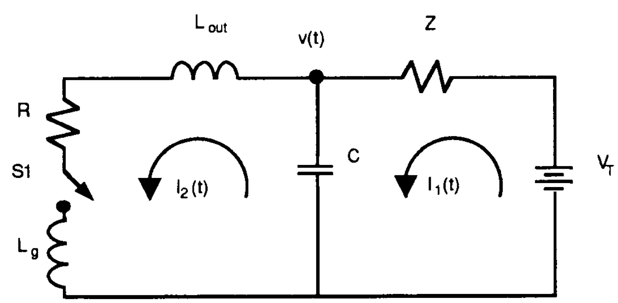

Fig 9
Simplified Output for Ground-Bounce Analysis

The output transistor is represented by the switch, S1, and its series resistance, R. The package lead inductors are L_g for the ground lead and L_{out} for the output lead. The stray capacitance of the package and connection of the package to the trace is C. Z is the impedance of the transmission line, terminated in a voltage V_T.

The differential equations governing circuit behavior when switch S1 is closed can be derived from the following basic relationships:

$$L_g \frac{dI_2(t)}{dt} + R * I_2(t) + L_{out} \frac{dI_2(t)}{dt} = v(t) , \tag{Eq 4}$$

$$I_1(t) - I_2(t) = C \frac{dv(t)}{dt} , \tag{Eq 5}$$

$$I_1(t) = \frac{V_T - v(t)}{Z} . \tag{Eq 6}$$

Substituting Eq 6 into Eq 5,

$$\frac{V_T - v(t)}{Z} - I_2(t) = C\,\frac{dv(t)}{dt}\,. \qquad \text{(Eq 7)}$$

Differentiating Eq 4,

$$\frac{dv(t)}{dt} = (L_g + L_{out})\,\frac{d^2I_2(t)}{dt^2} + R\frac{dI_2(t)}{dt}\,, \qquad \text{(Eq 8)}$$

then substituting Eqs 4 and 8 into Eq 7 and rearranging it gives an equation in $I_2(t)$:

$$\frac{d^2I_2(t)}{dt^2} + \left(\frac{RCZ + L}{LCZ}\right)\frac{dI_2(t)}{dt} + \left(\frac{R + Z}{LCZ}\right)I_2(t) = \frac{V_T}{LCZ}\,,$$

where $L = L_g + L_{out}$.

This is a second-order linear differential equation of the form

$$\frac{d^2y}{dt^2} + 2\zeta\omega_n\frac{dy}{dt} + \omega_n^2 y = \omega_n^2 x \quad \text{(See [B15], section 3.17, pp. 39–40.)} \qquad \text{(Eq 9)}$$

The natural frequency or ringing frequency is as follows:

$$\omega_n = \sqrt{\frac{R + Z}{LCZ}}\,.$$

The damping factor,

$$\zeta = \frac{1}{2\omega_n}\left(\frac{RCZ + L}{LCZ}\right),$$

and for the input, x,

$$\omega_n^2 x = \frac{V_T}{LCZ},$$

or

$$x = \frac{V_T}{LCZ} * \frac{LCZ}{R + Z} = \frac{V_T}{R + Z}\,.$$

Solving the characteristic equation,

$$D_2 + 2\zeta\omega_n D + \omega_n^2 = 0\,,$$

$$D_1 = -\zeta\omega_n + j\omega_n\sqrt{1 - \zeta^2}\,, \text{ and}$$

$$D_2 = -\zeta\omega_n - j\omega_n\sqrt{1 - \zeta^2}\,,$$

where $\zeta\omega_n$ is the damping coefficient, α, and $\omega_n\sqrt{1 - \zeta^2}$ is the damped natural frequency, ω_d. The homogeneous solution to Eq 9 is

$$y_h(t) = e^{-\alpha t}[K_1\cos(\omega_d t) + K_2\sin(\omega_d t)].$$

Knowing the initial conditions $I_2(0) = 0$ and $v(0) = V_T$, the constants K_1 and K_2 can be solved for.

$I_2(0) = 0$ means $K_1 = 0$.

$$\frac{dy(0)}{dt} = \frac{v(0) - RI_2(0)}{L} = \frac{V_T}{L} = \omega_d * K_2 \text{ means } K_2 = \frac{V_T}{L\omega_d}\,.$$

The solution, therefore, is

$$y_h(t) = \frac{V_T}{L\omega_d} e^{-\alpha t}\sin(\omega_d t) \ .$$

The weighted function is

$$\omega(t) = \frac{1}{\omega_d} e^{-\alpha t}\sin(\omega_d t) \ .$$

The weighted response to the unit step function at $t = 0$ is

$$y_u(t) = \int_0^t w(t-\tau)\omega_n^2 d\tau = 1 - \frac{\omega_n}{\omega_d} e^{-\alpha t} \sin(\omega_d t + \varphi) ,$$

where φ is defined as $\tan^{-1}\left(\frac{\omega_d}{\alpha}\right)$.

The total solution is $I_2(t) = y_h(t) + \frac{V_T}{R+Z} y_u(t)$,

$$I_2(t) = \frac{V_T}{L\omega_d} e^{-\alpha t}\sin(\omega_d t) + \frac{V_T}{R+Z}\left(1 - \frac{w_n}{w_d} e^{-\alpha t} \sin(\omega_n \sqrt{1-\zeta^2}\, t + \varphi)\right) ,$$

$$I_2(t) = V_T\left[\frac{1}{R+Z} + e^{-\alpha t}\left(\frac{1}{L\omega_d}\sin(\omega_d t) - \frac{\omega_n}{\omega_d(R+Z)}\sin(\omega_d t + \varphi)\right)\right].$$

Figures 10 and 11 illustrate the current, $I_2(t)$, the voltage, $v(t)$, and the voltage across L_g, the ground lead, using typical values for V_T, R, Z, C, L_{out}, and L_g.

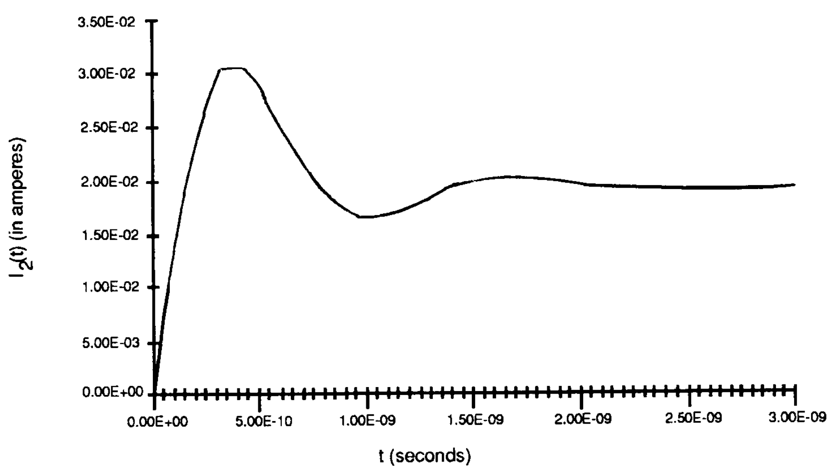

Fig 10
Current Through Output Leads

The voltage across the ground lead is

$$V_g = L_g \frac{dI_2(t)}{dt} ,$$

and the voltage $v(t)$ is

$$v(t) = L\,\frac{dI_2(t)}{dt} + R\,I_2(t)\ .$$

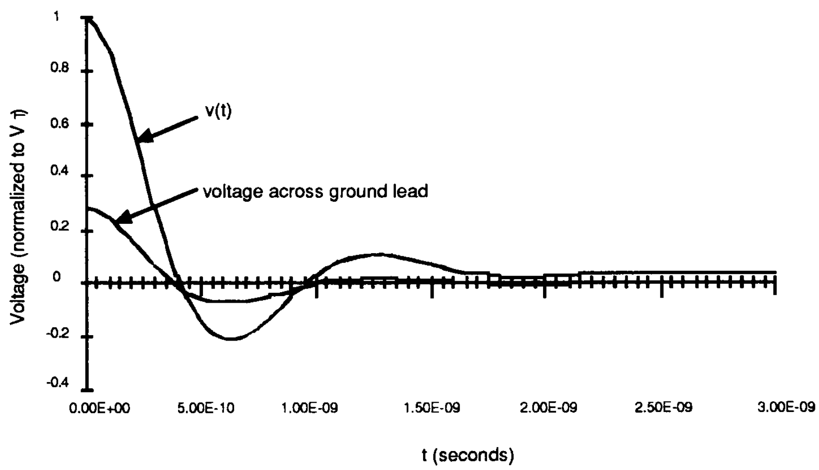

Fig 11
Voltage Across Ground Lead and $v(t)$

The magnitude of the initial ground bounce also can be estimated by knowing $I_2(0) = 0$, so all the voltage, $v(0) = V_T$, is across the inductors, $L_g + L_{out}$. The portion that will be across the ground lead, L_g, is

$$V_g = V_T * \frac{L_g}{L_g + L_{out}}\ .$$

In actual devices, the ground voltage, V_g, does not step instantaneously to $V_T * \frac{L_g}{L_g + L_{out}}$ at $t = 0$, as the step function response shows. This is because the output transistor has capacitance associated with it and takes time to turn on.

If the output driver is slower in turning on than the ringing frequency of the package (the damped natural frequency, ω_d, in the above discussion), then the ground bounce can be written with respect to the edge rate on the output.

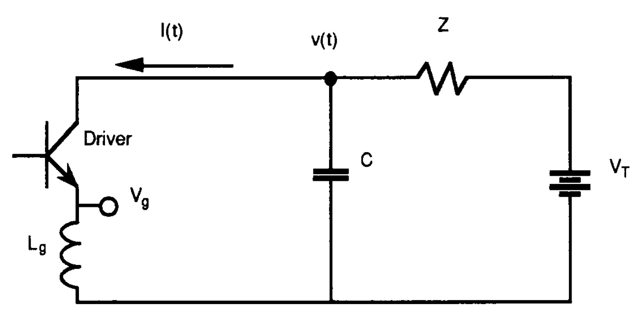

Fig 12
Simplified Output for Ground Bounce With Controlled Output

Figure 12 shows a simplified output driver connected to the bus, Z, and stray capacitance, C. The current the driver sinks, $I(t)$, is

$$I(t) = -C\frac{dv(t)}{dt} + \frac{V_T - v(t)}{Z} . \qquad \text{(Eq 10)}$$

The current $I(t)$ must flow through the inductor L_g. The voltage across L_g is

$$V_g = L_g\frac{dI(t)}{dt} . \qquad \text{(Eq 11)}$$

Differentiating Eq 10 and inserting it into Eq 11 gives

$$V_g = -\left(L_g\, C\frac{d^2v(t)}{dt^2} + \frac{L_g}{Z}\frac{dv(t)}{dt} \right) .$$

This gives a more accurate view of the output ground bounce, showing that the ground bounce, V_g, is not only related to the inductance, L_g, and rate of change of the output voltage but also to the second derivative of the output voltage. This means that both the output edge rate and the speed at which the edge rate changes while the output turns on and off are important.

Ground bounce can be reduced by using drivers with controlled rise and fall times, e.g., trapezoidal drivers, and reducing the inductance of the ground leads. The inductance of the package ground leads can be minimized by selecting the shorter side leads on DIP packages, by selecting smaller packages, or by using multiple (parallel) ground pins. The inductance of the printed circuit board can be reduced by providing large ground planes at the driving part.

Some related discussions on ground bounce are found in [B7].

4.2 Crosstalk. Crosstalk determines the maximum physical length of a bus that can be incorporated with acceptable reliability. Crosstalk is due to the distributed capacitive coupling, C_C, and distributed inductive coupling, L_C, between the lines. When crosstalk is measured on an undriven sense line next to a driven line (both terminated at their characteristic impedance), the near-end crosstalk and the far-end crosstalk have quite distinct features, as shown in Fig 13. Near-end crosstalk is the noise on the undriven sense line measured near the driving end where the signal originates. Far-end crosstalk is the noise on the undriven sense line measured at the far end to where the signal is propagating.

The peak amplitudes of near-end crosstalk, V_{NE}, and far-end crosstalk, V_{FE}, are given as follows:

$$V_{NE} = K_{NE}(2\tau_L)(V_I/T_R) \quad \text{for } T_R > 2\tau_L ,$$
$$V_{NE} = K_{NE}(V_I) \quad \text{for } T_R < 2\tau_L ,$$
$$V_{FE} = K_{FE}(L)(V_I/T_R) .$$

(See [B10], pp. 511–525.)

The pulse width of the near-end crosstalk, V_{NE}, is

$$T_{pulse} = T_R + 2\tau_L .$$

The pulse width of the far-end crosstalk, V_{FE}, is

$$T_{pulse} = T_R .$$

V_I is the signal swing on the driven line, T_R is the transition time of the input signal, and τ_L is the bus delay constant given by

$$\tau_L = L\sqrt{L_0 * C_0} ,$$

where

L = Length of the bus
L_0 = Distributed inductance per unit length
C_0 = Distributed capacitance per unit length

The coupling constants, K_{NE} and K_{FE}, are given by the expressions

$$K_{NE} = \frac{L(C_c * Z_0 + L_c/Z_0)}{4\tau_L}, \quad \text{(Eq 12)}$$

$$K_{FE} = \frac{C_c * Z_0 - L_c/Z_0}{2} \quad \text{ns/ft}, \quad \text{(Eq 13)}$$

where Z_0 is the characteristic impedance of the lines and is given by

$$Z_0 = \sqrt{\frac{L_0}{C_0}},$$

and

L_c = Distributed inductance coupling per unit length
C_c = Distributed capacitance coupling per unit length

The near-end component of crosstalk reduces to zero at the far end and the far-end component of crosstalk reduces to zero at the near end (see Fig 13). At any point in between, the crosstalk is a fractional sum of the near-end and far-end waveforms as shown.

Equations 12 and 13 show that the far-end crosstalk can have either polarity, while the near-end crosstalk always has the same polarity as the signal causing it. From the idealized depiction of crosstalk shown in Fig 13, several useful observations that apply to the general case can be made as follows:

(1) Crosstalk always scales with the signal amplitude.
(2) Absolute crosstalk amplitude is proportional to slew rate V_I/T_R, not just $1/T_R$.
(3) Far-end crosstalk width is always T_R.
(4) For $T_R < 2\tau_L$, the near-end crosstalk amplitude V_{NE} expressed as a fraction of the signal amplitude V_I is always a function of physical layout only.
(5) The higher the value of T_R, the lower the percentage of crosstalk (relative to signal amplitude).

The corresponding design implications are as follows:

(1) The noise margin expressed as a percentage of the signal swing is what is important, not the absolute noise margin. Therefore, to improve noise immunity, the percentage noise margin has to be maximized. This is achieved by reducing the receiver threshold uncertainty region and by centering the threshold between the high and low levels.
(2) Smaller signal amplitude with the same transition time reduces bus drive requirements without reducing noise immunity.
(3) Far-end crosstalk is eliminated if the receiver is designed to reject pulses having pulse widths less than or equal to T_R.
(4) When $T_R < 2\tau_L$, the near-end crosstalk immunity for a given percentage noise margin has to be built into the backplane printed circuit board layout. Since $(V_{NE}/V_I) = K_{NE}$ for this case, K_{NE} should be kept lower than the available worst-case noise margin. K_{NE} may be reduced by either increasing the spacing between lines

or by introducing a ground line between the lines. The ground line, in addition to increasing the spacing between the signal lines, forces the electric field lines to converge on it, significantly reducing crosstalk.

(5) For minimum crosstalk, the rise and fall times of the signal waveform should be as large as possible consistent with the minimum pulse width requirements of the bus. A driver that automatically limits the slew rate of the transition can go a long way towards reducing crosstalk.

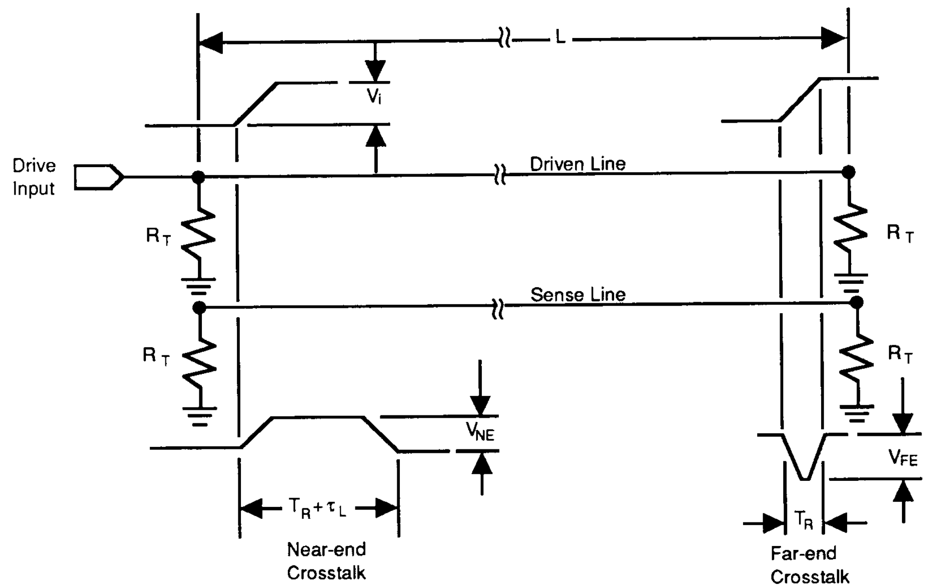

Fig 13
Crosstalk Under Ideal Conditions

4.3 Crosstalk Measurement. When multiple lines on either side of the sense lines switch simultaneously, the magnitude of the crosstalk is several times larger than single line switching, typically 3.5 times for microstrip backplanes. Also, the location of the drivers on the driven lines and the receiver on the sense line for worst-case crosstalk differs for the near-end and far-end cases as shown in Figs 14 and 15 for a uniformly loaded bus. But if the far-end crosstalk is not of the opposite polarity, then the combined effect of far-end and near-end crosstalk could have a larger amplitude and pulse width at a point near the middle of the sense line in Fig 14. So in the general case, or in the case of a nonuniformly loaded bus, it is advisable to check the sense line at several locations along the length of the bus to determine the worst-case crosstalk. The measurement should be made for both the positive and the negative transitions of the drive signal because of differences in transition times between the rising and falling edges.

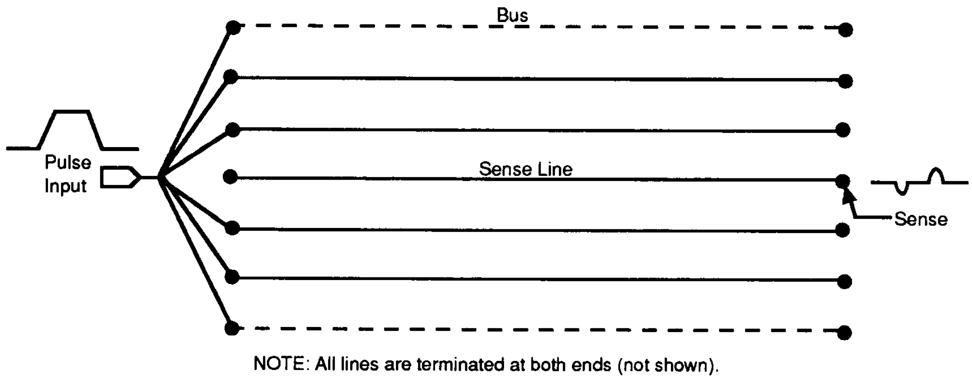

Fig 14
Worst-Case Far-End Crosstalk Measurement

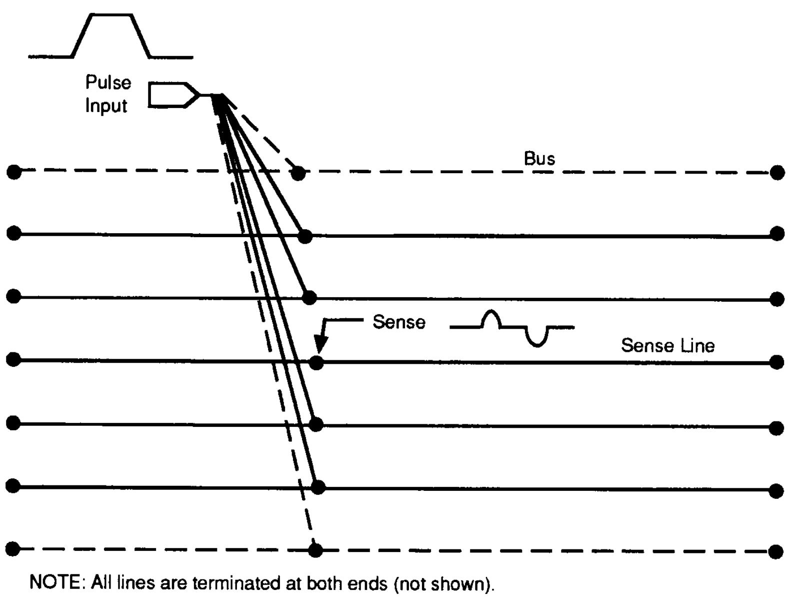

Fig 15
Worst-Case Near-End Crosstalk Measurement

5. Decoupling Techniques

Decoupling is a term that describes the lumped capacitances placed near devices, between the power supply potentials and ground. The capacitor supplies current for rapidly changing signals and helps keep the supply lines from dipping because of the changing demand for current through the inductance in the supply current paths. The capacitor also acts to isolate two circuits on a common supply line. The decoupling network is a low-pass filter (RC or RLC) (see [B8]). Since the capacitors help keep noise in one part from affecting the supply lines and thus the other parts on the board as well, the term *decoupling* is used. Another term used for the capacitor is *bypass*. A bypass capacitor, like the decoupling capacitor, is placed across the integrated circuit from V_{CC} to ground. It has the effect of shunting high-frequency current around the circuit.

The amount of capacitance needed can be calculated from the capacitor equation

$$I = C * \frac{dV}{dt},$$

or

$$C = \frac{I * \Delta t}{\Delta V}.$$

In a typical application, the current needed, *I*, might be 100 mA, the duration of the pulse might be 10 ns, and the tolerable change in voltage, 10 mV.

$$C = \frac{10^{-1} * 10^{-8}}{10^{-2}},$$

$$C = 10^{-7} = 0.1\ \mu\text{F}.$$

There are two other points to consider—the frequency response of the capacitors to be used and the distance from the device. Both of these concerns arise because of parasitic resistance and inductance in the capacitor. The equivalent model for a capacitor with all its parasitics is shown in Fig 16.

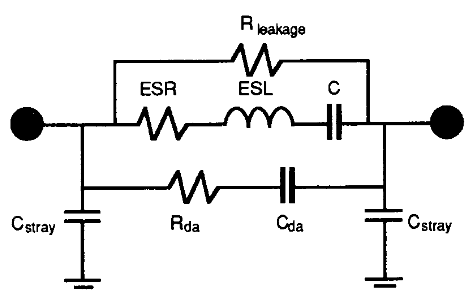

Fig 16
Equivalent Capacitor Diagram

The "da" in the R_{da} and C_{da} stands for dielectric absorption and relates to stray charge that can be absorbed into the dielectric material. A simplified model to consider just the critical element needed for an evaluation of decoupling is shown in Fig 17.

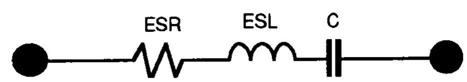

Fig 17
Simple Equivalent Capacitor Diagram

ESR is the effective series resistance (R_S) and ESL is the effective series inductance (L). The impedance of the simplified equivalent circuit is as follows:

$$Z_C = R_S + j\omega L + \frac{1}{j\omega C} ,$$

$$Z_C = R_S + j\left(\omega L - \frac{1}{\omega C}\right) \qquad \text{(see [B9]).} \qquad \text{(Eq 14)}$$

The magnitude of the impedance is $|Z_C| = \sqrt{R^2 + X^2}$, so

$$|Z_C| = \sqrt{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2} .$$

The resonant frequency is where $\omega L = \frac{1}{\omega C}$, or

$$\omega = \frac{1}{\sqrt{LC}} .$$

The minimum impedance is $Z_C = R$, and occurs at its resonant frequency.

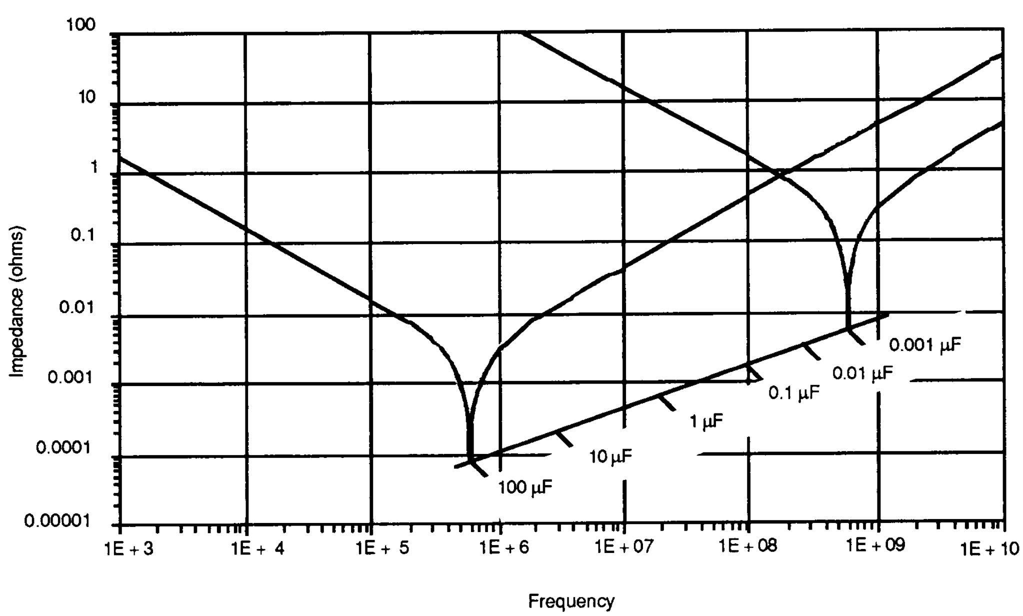

Fig 18
Example of Capacitor Impedance versus Frequency (Hz)

Figure 18 shows how the impedance varies with frequency. Typically as the capacitance goes down, so does the inductance, which means the resonant frequency goes up. The series resistance also usually goes up as the capacitance goes down.

Equation 14 can also be written as an equivalent capacitance, as follows:

$$Z_C = R_S + \frac{1}{j\omega C_{eq}},$$

where

$$C_{eq} = \frac{C}{1 - \omega^2 LC}. \qquad \text{(See [B4].)}$$

Because capacitors behave like RLC networks, they do not always look capacitive. Because of the series inductance, the capacitor has a resonant frequency, above which the capacitor looks more like an inductor. Some basic guidelines for choosing capacitors are as follows:

(1) Calculate the amount of capacitance needed by the capacitor equation,

$$I = C * \frac{dV}{dt},$$

based on the amount of current needed, how long the current will be needed, and what change in voltage is allowable.

(2) Choose a capacitor whose resonant frequency (Hz) is at least as high as the corresponding edge rates of the switching signals. The key equation that links the time domain to the frequency domain is that the rise time (10% to 90%) of a system with a single high-frequency roll-off (6 dB per octave) and a small-signal input depends upon the closed-loop small-signal bandwidth as

$$T_R = \frac{0.35}{f_{(-3\ \mathrm{dB})}},$$

where f is the frequency in Hertz.

(3) If possible, connect two (or more) capacitors in parallel to achieve the total capacitance required. The values of capacitors connected in parallel add algebraically. However, the frequency response (Fig 18) of each capacitor is not changed.

(4) In general, the smaller the value of the capacitor, the higher the series resonant frequency. Capacitors of smaller values may be connected in parallel to implement a band-pass filter, whose pass band is several decades wide.

(5) Place the capacitors as close to the switching device as possible. In the case of a transceiver, place the capacitance between the V_{CC} and ground planes at the part. In the case of termination, place the capacitance from V_T to ground at the termination resistor. See Fig 20 in 6.2.

6. Noise Margin Calculations and Backplane System Design Constraints

The design of a backplane requires care if it is going to be successful. The traces on the board and backplane act like transmission lines with a characteristic impedance, Z_0, and a characteristic delay, τ_0. Design constraints can be calculated for the unloaded and loaded characteristic impedance of the bus, the termination and the stub length using transmission line theory, noise margins, and the characteristics of the components driving and receiving the signals on the bus.

The basic theory behind transmission lines will be presented. The theory will be applied to backplane design. Relationships will be derived that can be used to set the values for the termination resistance, R_T, and the unloaded and loaded bus impedance, Z_U and Z_L. These same relationships can also be used to calculate the maximum stub length between the driver and the backplane.

Table 1
Summary Table of Critical Backplane Impedance Relationships

(Eq 15)	$Z_U < Z_L \dfrac{V_T + V_{IL} - 2V_{OL}}{V_T - V_{IL}}$
(Eq 16)	$Z_U < R_T \dfrac{V_T + V_{IL} - 2V_{OL}}{V_T - V_{IL}}$
(Eq 17)	$R_T < Z_L \dfrac{V_T - V_{OL}}{V_{IH} - V_{OL}}$
(Eq 21)	$Z_U < Z_L \dfrac{2V_{OH} - V_{OL} - V_{IH}}{V_{IH} - V_{OL}}$
(Eq 22)	$Z_U < R_T \dfrac{2V_{OH} - V_{OL} - V_{IH}}{V_{IH} - V_{OL}}$
(Eq 27)	$C_{stub} = \left[\left(\dfrac{Z_U}{Z_L}\right)^2 - 1\right](C_d + C_0)$

6.1 Summary of Transmission Line Theory as It Applies to Backplane Design. The derivation of the general wave equations for transmission lines can be found in texts on the subject (see, for example [B12], pp. 482–502). The equations for voltage and current on a transmission line are as follows:

$$V = V_1 e^{\alpha x} e^{j(\omega t + \beta x)} + V_2 e^{-\alpha x} e^{j(\omega t - \beta x)},$$

$$I = \frac{V_1}{\sqrt{Z/Y}} e^{\alpha x} e^{j(\omega t + \beta x)} - \frac{V_2}{\sqrt{Z/Y}} e^{-\alpha x} e^{j(\omega t - \beta x)}.$$

V_1 and V_2 are constants and represent the magnitude of the wave. x is the distance along the transmission line and t is time. $e^{\alpha x}$ and $e^{-\alpha x}$ are attenuation factors, and α is the attenuation constant. $e^{j(\omega t - \beta x)}$ and $e^{j(\omega t + \beta x)}$ are phase factors, with β being the phase constant and ω the frequency in radians per second. Z and Y are the impedance per unit length and admittance per unit length. These equations show that the voltages add for waves that are traveling in both directions and the currents subtract. To solve for the characteristic impedance of the line for a single wave traveling down the transmission line, just the first term of each equation need be used:

$$V = V_1 e^{\alpha x} e^{j(\omega t + \beta x)} ,$$

$$I = \frac{V_1}{\sqrt{Z/Y}} e^{\alpha x} e^{j(\omega t + \beta x)} .$$

The ratio V/I gives

$$V/I = Z_0 = \sqrt{\frac{Z}{Y}} = \sqrt{\frac{R + j\omega L}{G + j\omega C}} .$$

In backplane analysis, the transmission line can be considered lossless; that is, R and G are small or $\omega L >> R$ and $\omega C >> G$, so

$$Z_0 \sqrt{\frac{L_0}{C_0}},$$

and, v_0, the phase velocity,

$$v_0 = \frac{1}{\sqrt{L_0 C_0}} ,$$

or τ_0, the characteristic delay, is

$$\tau_0 = \sqrt{L_0 C_0} ,$$

where L_0 and C_0 are the inductance and capacitance per unit length of the transmission line.

Adding lumped capacitance along the backplane, as in the case of connectors and plug-in boards, modifies the characteristic impedance and characteristic delay as

$$Z_L = \frac{Z_0}{\sqrt{1+\frac{C_d}{C_0}}} , \quad \tau_L = \tau_0 * \sqrt{1+\frac{C_d}{C_0}} ,$$

where C_d is the added capacitance per unit length and C_0 is the capacitance per unit length of the unloaded Z_0.

At a boundary between two lines of different characteristic impedance, V_L, the voltage, and current are the sum of the incident and reflected waves,

$$V_L = V^+ + V^- ,$$
$$I_L = I^+ - I^- .$$

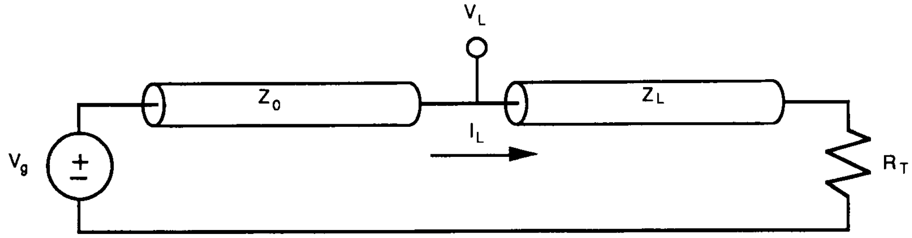

Fig 19
Transmission Line Boundary Between Z_0 and Z_L

The ratio, $V_L/I_L = Z_L$, so

$$Z_L = \frac{V+ + V-}{I+ - I-}\ .$$

Since $I+$ and $I-$ equal $V+/Z_0 + V-/Z_0$, respectively,

$$Z_L = \frac{V+ + V-}{V+/Z_0 - V-/Z_0}\ ,$$

or

$$Z_L = Z_0 \frac{V+ + V-}{V+ - V-} = Z_0 \frac{1 + V-/V+}{1 - V-/V+}\ .$$

The ratio, $V-/V+$, is the reflection coefficient, ρ.

$$Z_L = Z_0 \frac{1+\rho}{1-\rho}\ ,$$

or

$$\rho = \frac{Z_L - Z_0}{Z_L + Z_0}\ .$$

This relationship can be used to calculate the voltage due to impedance mismatch on a transmission line because the reflected voltage wave equals the incident voltage wave times ρ, or

$$V_{\text{reflected}} = V_{\text{incident}} * \rho.$$

The voltage is, then,

$$V_{\text{boundary}} = V_{\text{quiescent}} + V_{\text{incident}} + V_{\text{reflected}}.$$

6.2 Derivation of Critical Relationships. Using these transmission line equations, critical relationships can be written. These relate the loaded to unloaded impedance and the termination resistance to the loaded and unloaded impedance. Two methods for calculating maximum stub length are also derived. The first relates the maximum allowable lumped capacitance to the length of the stub. The second relates the edge rate to the length of the stub and uses the reflection at the backplane to limit overshoots and undershoots from the threshold region.

These relationships and constraints are derived by writing the equations of the voltages and currents at the impedance mismatch points, under worst-case allowable conditions, for both the mismatch on the bus, Z_U to Z_L, and for the mismatch between the backplane (Z_U or Z_L) and the termination resistance. Using the specifications for the drivers, receivers, and noise margins wanted, the critical relationships on the bus can be derived.

These relationships can be used in two ways. First, given the loaded and unloaded bus impedance, the termination resistance and the voltages, a value of noise margin on a given backplane can be calculated. Or, second, knowing the noise margin wanted and the driver and receiver specifications, the relationships between Z_U, Z_L, R_T, V_T, and stub length can be calculated to design a backplane.

In this discussion, the termination uses Thevenin-equivalent resistance and voltage. Some buses use a single resistor connected to a terminating voltage. In this case, these are the same as the Thevenin equivalents. In other cases, the Thevenin equivalents must be calculated. In the case of split-resistor termination,

$$\frac{1}{R_T} = \frac{1}{R1} + \frac{1}{R2},$$

$$V_T = V_{CC} \frac{R2}{R1 + R2}.$$

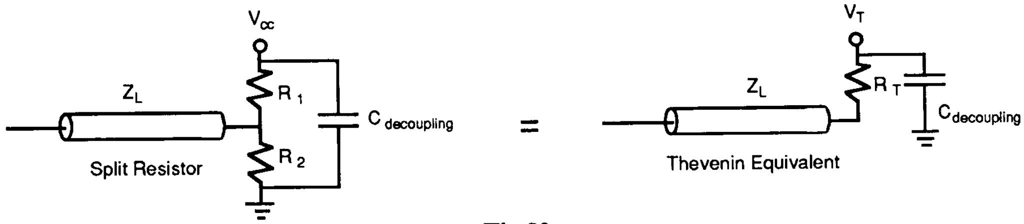

Fig 20
Thevenin-Equivalent Termination

In the following cases, configurations of backplane loading are evaluated that set up worst-case mismatches in Z_U and Z_L, and R_T and Z_U or Z_L. (It is these mismatches that cause reflections and degrade noise margins. For incident wave switching, these reflections cannot cause the reflected voltage wave to impinge on the threshold region.) Allowing the boards to be placed on the backplane in any position permits a configuration called "split backplane," where most of the boards are placed at one end with a large space of unloaded backplane between them and some others isolated at the other end. Boards grouped together create other problems that are also evaluated.

6.2.1 Case 1—High-to Low-Transition Driving From the End. Figure 21 shows the split-backplane configuration, driven from one end.

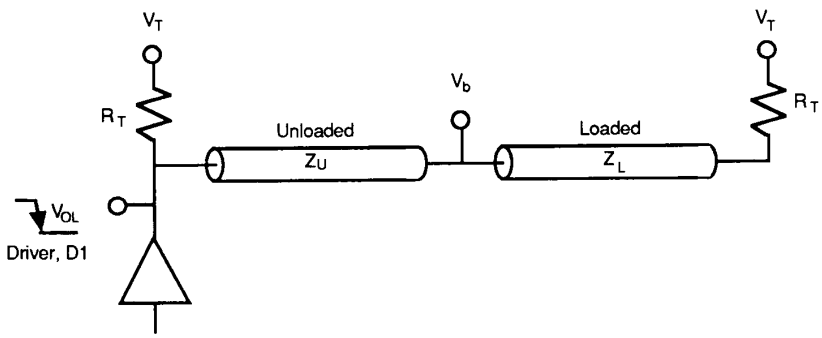

Fig 21
Split Backplane With Driver D1 Driving High to Low From the End

Since Z_L is lower than Z_U, the discontinuity at V_b (the boundary between the loaded and unloaded sections of the backplane) creates a negative reflection coefficient; the value of the voltage at V_b will be higher than the V_{OL} the driver generates. V_b incident wave voltage of a high-to-low transition caused by driver D1 can be solved for by the reflection coefficient,

$$V_{reflected} = V_{incident} * \rho .$$

$$V_b = V_{quiescent} + V_{incident} + V_{reflected},$$

$$= V_T + (V_{OL} - V_T) + (V_{OL} - V_T)\rho ,$$

$$V_b = V_{OL} + (V_{OL} - V_T)\rho .$$

The noise margin is defined as the difference between bus voltage, V_b, and the threshold voltage, V_{IL},

$$\text{Noise margin} = V_{IL} - V_b = V_{IL} - V_{OL} - (V_{OL} - V_b)\rho$$

Setting the noise margin at $V_b = V_{IL}$,

$$\rho = \frac{V_{IL} - V_{OL}}{V_{OL} - V_T} .$$

Knowing $\rho = \frac{(Z_L - Z_U)}{(Z_L + Z_U)} = \frac{\left(1 - \frac{Z_U}{Z_L}\right)}{\left(1 + \frac{Z_U}{Z_L}\right)}$, the ratio $\frac{Z_U}{Z_L}$ can be solved for:

$$\frac{Z_U}{Z_L} = \frac{V_T + V_{IL} - 2V_{OL}}{V_T - V_{IL}} ,$$

or

$$\boxed{Z_U < Z_L \frac{V_T + V_{IL} - V_{OL}}{V_T - V_{IL}} . \qquad \text{(Eq 15)}}$$

This case creates a relationship for the split-loaded backplane between Z_L *and* Z_U *that is independent of* R_T.

6.2.2 Case 2—High-to-Low Transition Driving From the Middle. Figure 22 shows a split-loaded backplane configuration driven from the middle at the loaded-to-unloaded boundary.

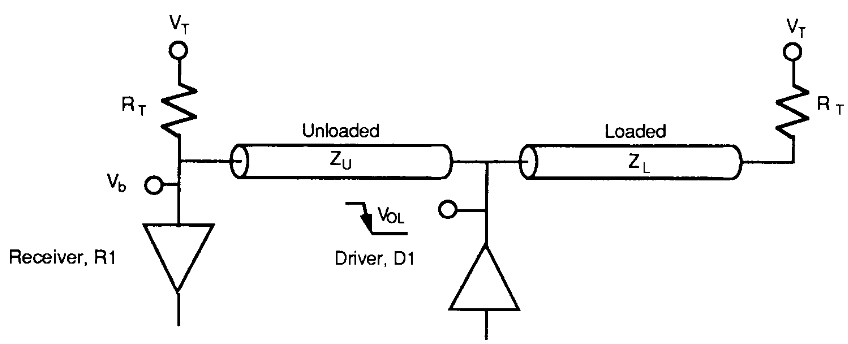

Fig 22
Split Backplane With Driver D1 Driving High to Low From the Middle

A receiver placed near the termination resistance, V_b, sees a reflected voltage caused by the mismatch between the termination resistance, R_T, and the unloaded backplane impedance, Z_U. Assuming that R_T is less than Z_U, the same negative reflection coefficient exists at V_b, as in case one, causing the voltage at V_b (incident wave) to be higher than the voltage, V_{OL}, generated at the driver. The incident wave voltage at V_b, high-to-low transition, caused by driver D1 can again be calculated.

$V_b = V_{\text{quiescent}} + V_{\text{incident}} + V_{\text{reflected}}$,

$= V_T + (V_{OL} - V_T) + (V_{OL} - V_T)\rho$,

$V_b = V_{OL} + (V_{OL} - V_T)\rho$.

The noise margin is again defined as the difference between the bus voltage, V_b, and the threshold voltage, V_{IL}:

Noise margin = $V_{IL} - V_b = V_{IL} - V_{OL} - (V_{OL} - V_T)\rho$

Setting the noise margin at $V_b = V_{IL}$,

$$\rho = \frac{V_{IL} - V_{OL}}{V_{OL} - V_T} .$$

Knowing $\rho = \dfrac{(R_T - Z_U)}{(R_T + Z_U)} = \dfrac{\left(1 - \dfrac{Z_U}{R_T}\right)}{\left(1 + \dfrac{Z_U}{R_T}\right)}$, the ratio $\dfrac{Z_U}{R_T}$ can be solved for:

$$\frac{Z_U}{R_T} = \frac{V_T + V_{IL} - 2V_{OL}}{V_T - V_{IL}} ,$$

or

$$\boxed{Z_U < R_T \frac{V_T + V_{IL} - 2V_{OL}}{V_T - V_{IL}} . \qquad \text{(Eq 16)}}$$

This case creates a relationship for the split-loaded backplane between R_T *and* Z_U *that is independent of* Z_L.

6.2.3 Case 3—Low-to-High Transition Driving From the Middle of a Fully Loaded Section. Figure 23 shows a general backplane loading configuration.

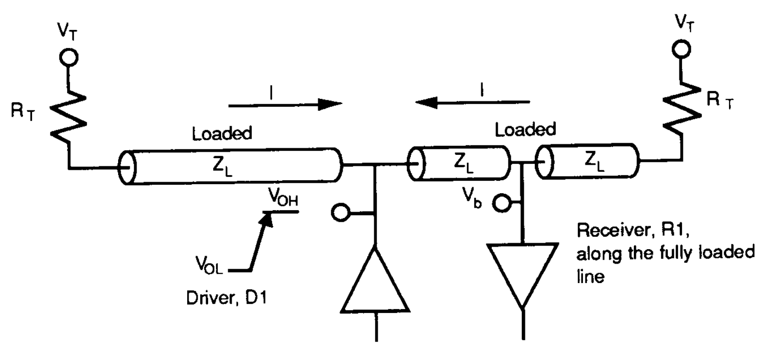

Fig 23
Fully Loaded Section of Backplane With Driver D1 Driving Low to High

This condition is valid for any fully loaded section or a fully loaded backplane, including the split-loaded backplane. The current to drive the line high is supplied by the termination resistor, R_T.

$$\Delta V = I * Z_L = (V_T - V_{OL})\frac{Z_L}{R_T},$$
$$V_{OH} = V_{OL} + (V_T - V_{OL})\frac{Z_L}{R_T},$$
$$\frac{Z_L}{R_T} = \frac{V_{OH} - V_{OL}}{V_T - V_{OL}}.$$

The noise margin is the difference between the bus voltage, V_{OH}, and the threshold voltage V_{IH}:

$$\text{Noise margin} = V_{OH} - V_{IH} = V_{OL} + (V_T - V_{OL})\frac{Z_L}{R_T} - V_{IH}.$$

Setting the noise margin, $V_{OH} = V_{IH}$,

$$R_T < Z_L\frac{V_T - V_{OL}}{V_{IH} - V_{OL}}. \qquad \text{(Eq 17)}$$

This case generates a relationship between R_T *and* Z_L *that is independent of* Z_U.

6.2.4 Case 4 —Low-to-High Transition Driving From the End. Figure 24 shows the split-loaded backplane configuration, driven low to high from the end.

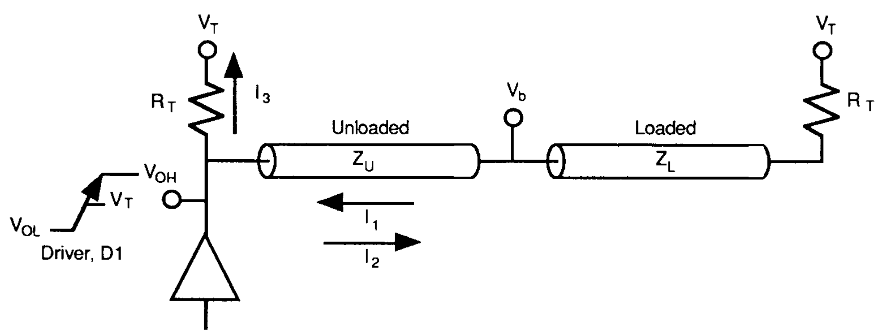

Fig 24
Split Backplane With Driver D1 Driving Low to High From the End

Since Z_U is always greater than Z_L, the negative reflection coefficient at V_b causes the voltage at V_b to be lower than that generated at the driver. V_b incident wave voltage of a low-to-high transition caused by driver D1 can be calculated by writing the equation of the currents.

$$I_1 = \frac{V_T - V_{OL}}{R_T}, \quad I_3 = \frac{V_{OH} - V_T}{R_T},$$
$$I_2 = I_1 - I_3,$$
$$I_2 = \frac{1}{R_T}(2V_T - V_{OL} - V_{OH}),$$
$$\Delta V = I_2 * Z_U = \frac{Z_U}{R_T}(2V_T - V_{OL} - V_{OH}),$$
$$V_{OH} = V_{OL} + \Delta V,$$
$$V_{OH} = V_{OL} + \frac{Z_U}{R_T}(2V_T - V_{OL} - V_{OH}),$$

solving for V_{OH},

$$V_{OH} + \frac{Z_U}{R_T} V_{OH} = V_{OL} + \frac{Z_U}{R_T}(2V_T - V_{OL}),$$

$$V_{OH}\left(1 + \frac{Z_U}{R_T}\right) = V_{OL}\left(1 - \frac{Z_U}{R_T}\right) + 2V_T\frac{Z_U}{R_T},$$

$$V_{OH} = V_{OL}\frac{1 - \frac{Z_U}{R_T}}{1 + \frac{Z_U}{R_T}} + 2V_T\frac{\frac{Z_U}{R_T}}{1 + \frac{Z_U}{R_T}}, \quad \text{or}$$

$$V_{OH} = V_{OL}\frac{R_T - Z_U}{R_T + Z_U} + 2V_T\frac{Z_U}{R_T + Z_U}. \qquad \text{(Eq 18)}$$

This generates the incident wave down the line which hits the discontinuity at V_b, causing a reflection. The voltage at V_b is calculated by

$$V_b = V_{OH} + V_{\text{reflected}},$$

$$V_b = V_{OH} + (V_{OH} - V_{OL}) * \rho, \qquad \text{(Eq 19)}$$

again where $\rho = \frac{(Z_L - Z_U)}{(Z_L + Z_U)}$. Combining Eqs 18 and 19,

$$V_b = V_{OL}\frac{R_T - Z_U}{R_T + Z_U} + 2V_T\frac{Z_U}{R_T + Z_U} + \left(V_{OL}\frac{R_T - Z_U}{R_T + Z_U} + 2V_T\frac{Z_U}{R_T + Z_U} - V_{OL}\right) * \rho,$$

$$V_b = V_{OL}\frac{1 - R}{1 + R} + 2V_T\frac{R}{1 + R} + \left(V_{OL}\frac{1 - R}{1 + R} + 2V_T\frac{R}{1 + R} - V_{OL}\right) * \rho,$$

where $R = \frac{Z_U}{R_T}$. Solving for R,

$$R = \frac{Z_U}{R_T} = \frac{V_b - V_{OL}}{2V_T - V_b - V_{OL} - 2V_{OL} * \rho + 2V_T * \rho},$$

Setting $V_b = V_{IH}$,

$$\boxed{R_T < Z_U \frac{2V_T - V_{IH} - V_{OL} - 2V_{OL} * \rho + 2V_T * \rho}{V_{IH} - V_{OL}}.} \qquad \text{(Eq 20)}$$

In this configuration, the ratio Z_U to R_T is dependent on the reflection coefficient at the loaded-to-unloaded boundary, $\rho = \frac{(Z_L - Z_U)}{(Z_L + Z_U)}$. Knowing the reflection coefficient at the termination, ρ_T, $\frac{Z_U}{R_T} = \frac{1 + \rho_T}{1 - \rho_T}$, the ratio, Z_L to Z_U, can be calculated.

If the V_{OH} level is clamped at a value lower than V_T, a simpler relationship can be derived.

$$\Delta V = V_{OH} - V_{OL},$$

$$V_b = V_{OH} + (V_{OH} - V_{OL})\rho,$$

$$\rho = \frac{\left(1 - \frac{Z_U}{Z_L}\right)}{\left(1 + \frac{Z_U}{Z_L}\right)} = \frac{V_b - V_{OH}}{V_{OH} - V_{OL}},$$

$$\frac{Z_U}{Z_L} = \frac{2V_{OH} - V_{OL} - V_b}{V_b - V_{OL}},$$

setting $V_b = V_{IH}$,

$$\boxed{Z_U < Z_L \frac{2V_{OH} - V_{OL} - V_{IH}}{V_{IH} - V_{OL}}\ .} \qquad \text{(Eq 21)}$$

6.2.5 Case 5—Low-to-High Transition Driving From the Middle. Figure 25 is the split-loaded backplane driven at the boundary of the loaded-to-unloaded interface.

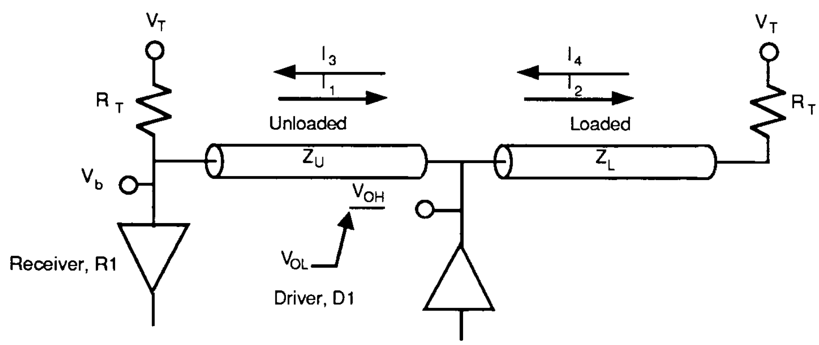

Fig 25
Split Backplane With Driver D1 Driving Low to High From the Middle

The negative reflection coefficient at the termination, V_b, causes the voltage at V_b to be lower than that generated by the driver. The incident wave voltage at V_b, low-to-high transition, caused by driver D1 can be solved for by writing the equations for the currents.

$$I_1 = I_2 = \frac{V_T - V_{OL}}{R_T}, \quad I_3 = \frac{V_{OH} - V_{OL}}{Z_U}, \quad I_4 = \frac{V_{OL} - V_{OL}}{Z_L},$$

$$I_2 + I_2 = I_3 + I_4,$$

$$2 * \frac{V_T - V_{OL}}{R_T} = \frac{V_{OH} - V_{OL}}{Z_U} + \frac{V_{OH} - V_{OL}}{Z_L}\ .$$

Solving for V_{OH},

$$V_{OH} = 2\left(\frac{V_T - V_{OL}}{R_T}\right)\frac{Z_U Z_L}{Z_U + Z_L} + V_{OL}\ .$$

$V_b = V_{quiescent} + V_{incident} + V_{reflected}$,
$= V_{OL} + (V_{OH} - V_{OL}) + (V_{OH} - V_{OL})\rho_T$,
$= V_{OH} + (V_{OH} - V_{OL})\rho_T$,
$V_b = (1 + \rho_T)V_{OH} - V_{OL} * \rho_T$,

$$V_b = (1 + \rho_T)\left[2\left(\frac{V_T - V_{OL}}{R_T}\right)\frac{Z_U Z_L}{Z_U + Z_L} + V_{OL}\right] - V_{OL} * \rho_T ,$$

where $\rho_T = \dfrac{(R_T - Z_U)}{(R_T + Z_U)} = \dfrac{1 - \dfrac{Z_U}{R_T}}{1 + \dfrac{Z_U}{R_T}}$. $Z_L = Z_U * \dfrac{1 + \rho_1}{1 - \rho_1}$.

ρ_T is the reflection coefficient at the termination, and ρ_1 is the reflection coefficient at the loaded-to-unloaded boundary. This sets up a relationship between Z_L, Z_U, and R_T. Knowing ρ_1, and setting $V_b = V_{IH}$, the ratio $\frac{Z_U}{R_T}$ can be solved for. Or knowing ρ_T and setting $V_b = V_{IH}$, the ratio $\frac{Z_L}{R_T}$ can be solved for.

If the V_{OH} level is clamped at a value lower than V_T, a simpler relationship can be derived.

$$\Delta V_{OH} = V_{OH} - V_{OL} ,$$

$$V_b = V_{OH} + (V_{OH} - V_{OL})\rho ,$$

$$\rho = \frac{(R_T - Z_U)}{(R_T + Z_U)} = \frac{1 - \dfrac{Z_U}{R_T}}{1 + \dfrac{Z_U}{R_T}} = \frac{V_b - V_{OH}}{V_{OH} - V_{OL}} ,$$

$$\frac{Z_U}{R_T} = \frac{2V_{OH} - V_{OL} - V_b}{V_b - V_{OL}} ,$$

setting $V_b = V_{IH}$,

$$\boxed{Z_U < R_T \frac{2V_{OH} - V_{OL} - V_{IH}}{V_{IH} - V_{OL}} . \qquad \text{(Eq 22)}}$$

6.2.6 Case 6—High-to-Low Transition Driving From the Middle, Receiving Anywhere Along the Line. In Fig 26 the incident wave, high to low, generated by the driver, D1, must pass through the threshold region, V_{IL}.

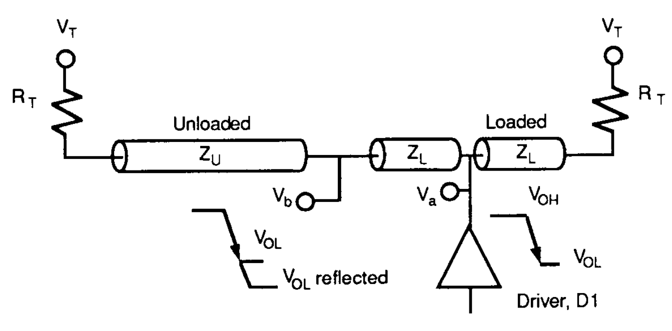

Fig 26
Partially Loaded Section of Backplane With Driver D1 Driving High to Low

The current (I_{OL}, output current low state) needed to do this is the voltage change divided by the impedance.

$$I_{OL} \text{ needed} = \frac{(V_T - V_{IL})}{\frac{Z_L}{2}} .$$

Generally backplane driver specifications guarantee an I_{OL} at a specified V_{OL} or a V_{OL} when driving the fully allowable Z_L. This I_{OL} is, then,

$$I_{OL} \text{ guaranteed} = \frac{(V_T - V_{OL})}{\frac{Z_L}{2}} ,$$

Since the discontinuity at the boundary, Z_L to Z_U, creates a positive reflection coefficient, $Z_U > Z_L$, the reflected wave voltage will reinforce the incident wave. The low reflected wave travels back down the line to the driver at V_a. If the driver, D1, tries to release the line during the reflected wave, it will be at a lower voltage, putting an additional constraint on the reflection coefficient, ρ.

$$V_b = V_{quiescent} + V_{incident} + V_{reflected} ,$$
$$V_b = V_{OL} + (V_{OL} - V_T)\rho .$$

$$\rho = \frac{(Z_U - Z_L)}{(Z_U + Z_L)} = \frac{\left(1 - \frac{Z_L}{Z_U}\right)}{\left(1 + \frac{Z_L}{Z_U}\right)} = \frac{(1 - R)}{(1 + R)} , \text{ where } R = \frac{Z_L}{Z_U} .$$

Since Z_U is larger than Z_L, the reflection coefficient is positive and the voltage at V_b will be lower than the V_{OL} of the driver. When the reflection returns to the driver, D1, it will have to release the line from that lower point. The current in the reflected wave is the reflected voltage divided by the impedance or

$$I_{reflected} = \frac{V_{reflected}}{Z_L} .$$

This current directly subtracts from the incident wave current, leaving the current to charge the line as

$$I_{available} = \frac{(V_{OL} - V_T)}{Z_L} - \frac{V_{reflected}}{Z_L}$$
$$I_{available} = \frac{(V_{OL} - V_T)}{Z_L} - \frac{(V_{OL} - V_T)\rho}{Z_L} .$$

The voltage change when the line is released is

$$\Delta V = I_{available} * Z_L .$$

$$(V_{OL} + (V_{OL} - V_T)\rho) - V_{IH} = \left[\frac{(V_{OL} - V_T)}{Z_L} - \frac{(V_{OL} - V_T)\rho}{Z_L}\right] * Z_L$$

Solving for the reflection coefficient,

$$\boxed{\rho < \frac{V_T - V_{IH}}{2V_T - 2V_{OL}} . \qquad \text{(Eq 23)}}$$

6.2.7 Case 7—Low-to-High Transition Driving From the Middle of a Partially Loaded Backplane, Receiving at the Loaded-to-Unloaded Interface. In Fig 27, the driver, D1, releases the line at point V_a, and the voltage makes a transition low to high.

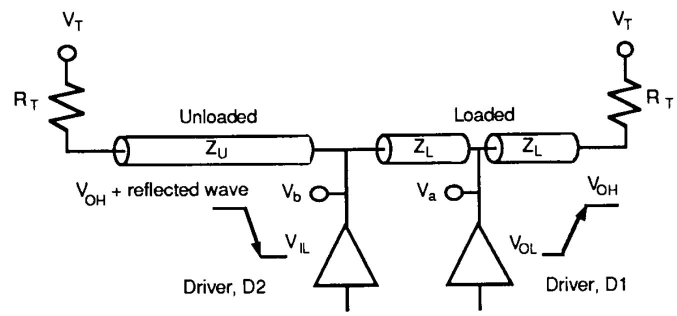

Fig 27
Partially Loaded Section of Backplane With Driver D1 Driving Low to High

The mismatch at V_b causes a positive reinforcement of the voltage. The reflected wave from the termination resistors cannot be relied upon to reduce the overshoot because the bus timing requires incident wave switching. The driver, D2 or D1, must be able to assert the line at least through the switching voltage, V_{IL}. The maximum current D1 or D2 can sink by specification is I_{OL}. (Typically an output driver can sink a lot more than this at V_{OUT} greater than V_{OL}, but that is neither specified nor guaranteed.) Therefore, a maximum voltage step is

$$\Delta V = I_{OL} * \frac{Z_L}{2} .$$

The maximum voltage D1 or D2 can see and still pull the line low is

$$V_{max} = V_{IL} + \Delta V,$$
$$V_{max} = V_{IL} + I_{OL} * \frac{Z_L}{2} . \qquad \text{(Eq 24)}$$

The low-to-high wave generated by D1 releasing the line is

$$V_{max} = V_{olmin} + \frac{(V_T - V_{olmin})}{\frac{R_T}{2}} * \frac{Z_L}{2} + \text{reflected wave},$$

$$V_{max} = V_{olmin} + (V_T - V_{olmin}) * \frac{Z_L}{R_T} + (V_T - V_{olmin}) * \frac{Z_L}{R_T} * \rho \qquad \text{(Eq 25)}$$

The reflection coefficient $\rho = \frac{(Z_U - Z_L)}{(Z_U + Z_L)}$. Putting Eqs 24 and 25 together,

$$V_{IL} + I_{OL}\frac{Z_L}{2} = V_{olmin} + (V_T - V_{olmin}) * \frac{Z_L}{R_T} + (V_T - V_{olmin}) * \frac{Z_L}{R_T} * \rho.$$

Solving for the reflection coefficient,

$$\boxed{\rho < \frac{V_{IL} + I_{OL} * \frac{Z_L}{2} - V_{olmin} - (V_T - V_{olmin}) * \frac{Z_L}{R_T}}{(V_T - V_{olmin}) * \frac{Z_L}{R_T}} . \qquad \text{(Eq 26)}}$$

Knowing $\rho = \frac{(Z_U - Z_L)}{(Z_U + Z_L)} = \frac{(1 - Z_L/Z_U)}{(1 + Z_L/Z_U)}$, the ratio Z_L/Z_U can be solved for. This is also the same calculation for a wave driving into a termination, but the reflection coefficient would be $\rho = \frac{(R_T - Z_L)}{(R_T + Z_L)}$ and the ratio would be Z_L/R_T.

6.2.8 Discussion of Switching During the Reflected Signals. Cases 6 and 7 deal with the maximum data transfer rate or bandwidth of the backplane. At high clock rates the output drivers are making transitions before reflections have a chance to die out. The two cases presented are necessary but not sufficient conditions for correct backplane operation. This means that the conditions of cases 6 and 7 must be met, but there might be more complex configurations of backplane loading or operation that would present more severe constraints. Setting the impedance mismatch determines a maximum transfer rate. This is so because the reflections caused by the impedance mismatch have an additive effect when switching faster than reflections on the bus can settle. A higher transfer rate can be achieved if the mismatch between Z_L and Z_U is reduced.

A measurement technique to determine the maximum frequency of the backplane is shown in Fig 28. The backplane is loaded with a worst-case mismatch allowable in Z_L and Z_U, and is driven by a pulse generator, as shown in Fig 28, or by an appropriate driver, if it is suitably fast enough. As the pulse frequency is increased, the reflected waves will not have a chance to settle before the next transition. When the reflected noise degrades into the threshold window it is considered to be operating beyond reliable limits. Probing the voltage along the bus at various points might be necessary to find the worst cumulative reflection.

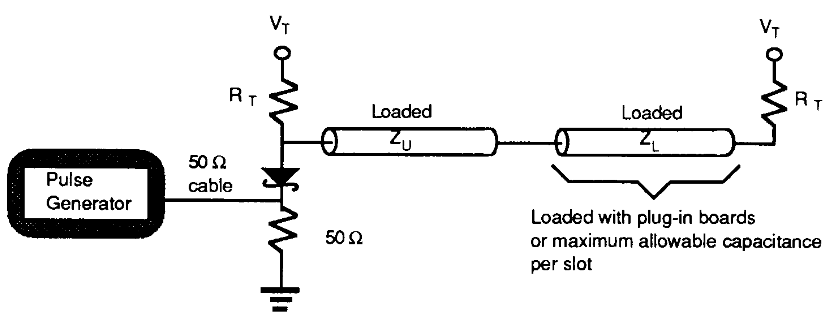

Fig 28
Measurement Scheme to Determine Maximum Transfer Rate

6.3 Stub Length Calculation (length of trace on plug-in board)

6.3.1 Calculating Stub Length by Maximum Capacitance. The allowable length between the bus transceiver and the backplane can be calculated by using the relationships established, the values for the capacitance per unit length, and the inductance per unit length for the intrinsic transmission line on the stub. From the equations

$$Z_0 = \sqrt{\frac{L_0}{C_0}} \text{ and } \tau_o = \sqrt{L_0 C_0},$$

L_0 and C_0 can be calculated by solving the simultaneous equations. Since the ratio between Z_L and Z_U is a limiting factor from case one, and knowing

$$Z_U = \frac{Z_0}{\sqrt{1 + \frac{C_d}{C_0}}} \text{ and } Z_L = \frac{Z_U}{\sqrt{1 + \frac{C_{stub}}{C_0 + C_d}}},$$

(where C_d is all the lumped capacitance of the connector and via, etc., except the plug in board, on backplane, and C_{stub} is capacitance of the stub),

$$R = \frac{Z_U}{Z_L} = \sqrt{1 + \frac{C_{stub}}{C_0 + C_d}},$$

or

$$\boxed{C_{stub} = \left[\left(\frac{Z_U}{Z_L}\right)^2 - 1\right](C_d + C_0) \qquad \text{(Eq 27)}}$$

If the ratio between Z_0 and Z_L is the limiting ratio, then

$$Z_L = \frac{Z_0}{\sqrt{1 + \frac{C_{stub}}{C_0 + C_d}}\sqrt{1 + \frac{C_d}{C_0}}},$$

from which C_{stub} may be calculated. Subtracting from C_{stub}, the lumped capacitive elements on the stub such as connectors, vias, sockets, and packaged parts, the remainder is used for allowable stub length. The length is calculated from the capacitance allowed for the trace, times the characteristic capacitance per unit length of the stub.

6.3.2 Calculating Stub Length by Edge Rate and Reflection. The criteria for using edge rate to determine stub length is that ringing caused by the impedance mismatch between the stub and the loaded line cannot impinge on the threshold window. The stub is the transmission line between the driver and the backplane and is usually of higher impedance than the parallel combination of the two backplane traces the stub sees. A signal initiated from the driver travels down the stub until it hits the backplane. The impedance mismatch at the stub backplane connection causes a reflection that travels back to the driver. This causes the ringing on the stub.

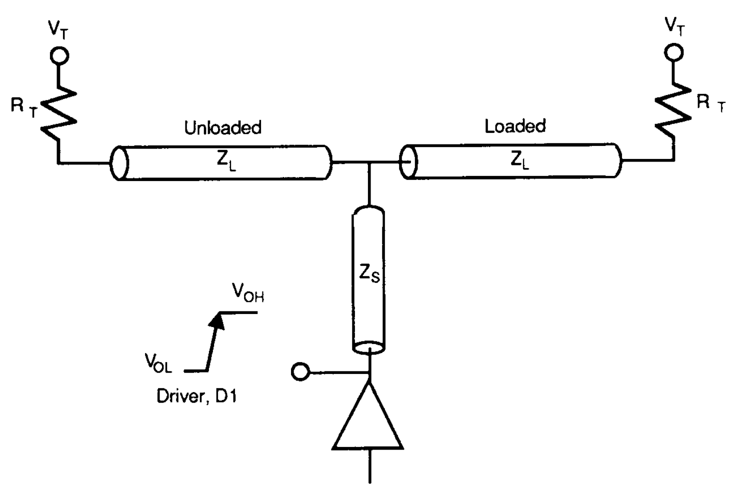

Fig 29
Stub Illustration Driving Low to High

The driver, D1, makes a transition from low to high. Assuming a step function, the V_{OH} level will be

$$V_{OH} = V_{OL} + 2(V_T - V_{OL}) \frac{Z_S}{R_T} .$$

The reflection coefficient at the boundary between the stub and the backplane is

$$\rho = \frac{\frac{Z_U}{2} - Z_S}{\frac{Z_U}{2} + Z_S} ,$$

$$\rho = \frac{1 - 2\frac{Z_S}{Z_U}}{1 - 2\frac{Z_S}{Z_U}} .$$

Usually this reflection coefficient $\rho \approx -1$. This means most of the incident wave is reflected back at the driver. If the rising ramp is slow compared to the round-trip delay of the stub, the reflection will limit the ringing. To prevent the ringing from dipping into the threshold region and causing multiple switching, a good rule of thumb is that the ramp should be five times slower than the round trip delay of the stub, and three for short stubs under 2 in. Propagation delay on the stub is determined by the propagation delay constant

$$\tau_0 = \sqrt{LC} .$$

With the distance, d, and the round trip, $2d$, the round-trip time is

$$\text{delay} = \tau_0 * 2d,$$

Using our rule of thumb of stub delay $= \frac{T_{RAMP}}{5}$, then the maximum stub length is

$$\text{Stub length} = \frac{(T_{RAMP}/5)}{2 * \tau_0} .$$

7. Bibliography

[B1] IEEE Std 896.1-1987, IEEE Standard Backplane Bus Specification for Multiprocessor Architectures: Futurebus (ANSI).[3]

[B2] Amemiya, H. "Time Domain Analysis of Multiple Parallel Transmission Lines," *RCA Review,* vol. 28, June 1967, pp. 241–276.

[B3] Borrill, Paul L. "Microprocessing System—Techniques and the Application to Space Research," Ph.D Thesis, University College, London, March 1986.

[B4] Botos, Bob. "Designer's Guide to: RLC Measurements," Hewlett Packard Co. Reprinted from the June 5, 1979 issue of *EDN.*

[B5] Catt, I. "Crosstalk (Noise) in Digital Systems," *IEEE Transactions on Digital Computers,* EC-16, Dec. 1967, pp. 743–763.

[B6] DeFalco, J. A. "Reflections and Crosstalk in Logic Circuit Interconnections," *IEEE Spectrum,* vol. 7, July 1970, pp. 44–50.

[B7] Dike, Charles E. "Equivalent Circuits Model Subtle Traits of Advanced CMOS," *EDN,* Apr. 14, 1988.

[B8] Doubrava, Laudie. "Proper Handling of Voltage Spikes Safeguards Circuit Design," EDN, March 5, 1979, pp. 83–87.

[B9] Epand, Donald and Ken Liddane. "Selecting Capacitors Property," *Electronic Design 13,* June 21, 1977.

[B10] Feller, A., H. R. Kaupp, and J. J. Digiacomo. "Crosstalk and Reflections in High Speed Digital Systems," *Proceedings AFIPS Fall Joint Computer Conference,* 1965, pp. 511–525.

[B11] HP Application Note 62. "Time Domain Reflectometry," Hewlett-Packard Co, 1964, 1988.

[B12] Kraus, John D. and Keith R. Carver. "Electromagnetics," McGraw-Hill, Inc., 1973.

[B13] Magnuson, Philip C. "Transmission Lines and Wave Propagation." 2nd ed. Oregon State University Book Stores, 1987.

[B14] Matick, Richard E. *Transmission Lines for Digital and Communications Networks.* McGraw Hill, 1969.

[B15] Schaum's Outline Series. "Theory and Problems of Feedback and Control Systems."

[3]IEEE publications can be obtained from the Institute of Electrical and Electronics Engineers, Inc., Service Center, 445 Hoes Lane, P.O. Box 1331, Piscataway, NJ 08855-1331, 1-800-678-4333.