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MODELING, SIMULATION, AND MEASUREMENT CONSIDERATIONS OF HIGH-SPEED DIGITAL BUSES

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Abstract

With the ever increasing clock speed of computers, the quality of digital signals becomes important. The propagation time through the interconnecting elements (PCB traces, wires, cables, connectors) is no longer negligible compared to the delays in the semiconductor devices, and transition times of digital devices may often be less than the propagation time along the interconnecting path. The proper matching conditions of these interconnections are essential. The paper gives an insight to some of the major limiting factors of high-speed buses. It is shown that on loaded backplanes, the propagation delay is a function of the device's trigger threshold. The stub-ringing phenomenon is also discussed. For single, capacitively loaded stubs, the ringing frequency is shown to be lower than what one expects from the widely used quasi-lumped models. A simple dual-stub 'butterfly' model is introduced to predict the ringing of backplanes. This model predicts the stub ringing in high-impedance states of TTL buses or high-level states of BTL buses. It is shown that the damping factor of the oscillation depends on the ratio of stub spacing and stub length, and is a very weak function of the termination resistance used at the end of backplane.

I. Introduction

Transmission lines

Characteristic impedance (Z_0) and unloaded propagation delay (t_{pd}) of a transmission line are determined by the unit resistance (R_0), inductance (L_0), conductance (G_0), and capacitance (C_0), as well as the physical length (l) of transmission line (Fig. 1). In lossy case, the characteristic impedance and propagation function (γ) are complex, and depend on frequency (ω):

$$Z_0 = \sqrt{\frac{(R_0 + j\omega L_0)}{(G_0 + j\omega C_0)}} \qquad (1)$$

$$\gamma = \sqrt{(R_0 + j\omega L_0)(G_0 + j\omega C_0)} \qquad (2)$$

For lossless transmission lines, the resistance and conductance are zero, this yields frequency independent, resistive characteristic impedance, and imaginary propagation function:

$$Z_0 = \sqrt{\frac{L_0}{C_0}} \qquad (3)$$

$$\gamma = j\omega\sqrt{L_0 C_0} \qquad (4)$$

$$t_{pd} = l\sqrt{L_0 C_0} = \sqrt{LC} \qquad (5)$$

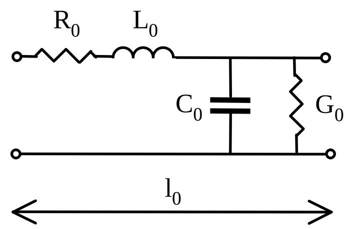

Fig. 1. Lumped equivalent of an l_0 *segment of a transmission line.* R_0, L_0, G_0, *and* C_0 *are the resistance, inductance, conductance, and capacitance per unit length (*l_0*). The total length of line is l.*

The lossless transmission line is nondispersive, *i.e.*, the time-domain shape of a wave travelling down a homogeneous line does not vary with location [1]. This holds true for the circuit of Fig. 1. in the passband of the equivalent lowpass filter.

At each discontinuity of the transmission line a portion of incident wave is reflected (see Fig. 2.). The reflection coefficient (Γ) is the ratio of reflected and incident waves. Γ can be expressed in terms of characteristic impedance of transmission line and impedance of connected device:

$$\Gamma = \frac{Z - Z_0}{Z + Z_0} \qquad (6)$$

$|\Gamma| = 1$, when Z reactive, short, or open.

Note that the reflection coefficient may be different at a point of transmission line looking in different directions. In Fig. 2a, the source reflection coefficient Γ_S, and load reflection coefficient Γ_L are obtained by substituting Z_S, and Z_L into (6), respectively.

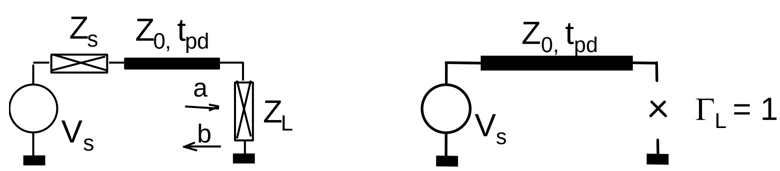

Fig. 2. Definition of reflection coefficient for $\Gamma_L = b/a$ *(left), and illustration of total reflection caused by an extreme (open) termination (right).*

<u>Noise Sources in Digital Systems</u>

In high-speed digital systems there are several contributors to the overall noise. A typical data-communication path has three major elements: signal source, propagation medium, and receiver. Each of these elements may generate noise.

In the logic circuit functioning as transmitter (signal source), noise is the result of ground bounce and crosstalk in leadframes.

In the communication path, along the printed-circuit-board traces, cables, and connectors, noise is the result of crosstalk among traces, wires, connector pins, ground shift of the system, and reflections at discontinuities. Crosstalk in multiwire interconnections and in packages are widely covered in the literature: [3] provides a fundamental description of crosstalk in multiwire systems, [7] describes efficient simulation methods for lossy transmission lines, while [6] gives closed-form solution for the crosstalk of lossless lines loaded by linear non-reactive components. Crosstalk on printed-circuit boards are treated in [4], high-speed integrated circuit packaging are analyzed in [5].

During the subsequent discussions transmission lines are assumed to be lossless and free of interaction from other lines in the bus, *i.e.*, crosstalk is neglected.

Providing matched termination is a major problem on backplanes, where the configuration (number and location of plug-in boards) may vary during use. Duration of 'steps' due to reflections on the backplane is determined by the total two-way propagation delay between discontinuities, while the actual value of reflection coefficient varies with configuration, but usually does not get very close to unity, or total reflection.

Stub ringing is also the result of signal reflections, but as opposed to backplane reflections, stub ringing is inherent in every system based on a backplane with plug-in boards, regardless of the value of backplane termination. Stubs are unterminated or capacitively loaded transmission lines consisting of the connectors and attached PCB wiring of plug-in boards. The ringing is the result of the total reflection on the reactive, or open termination. The damping factor of ringing is low. Stubs are usually short compared to the total length of bus, this results in a ringing frequency which is typically much higher than the frequency of multiple backplane reflections.

Signal propagation along a transmission line strongly depends on the loads connected to various points of the line. If the transition time of the signal is longer than the two-way propagation delay between discontinuities, signal reflections appear as a change in waveshape, *e.g.*, an increase in the transition time of signal. A lumped parameter model is used in [2] to obtain the performance of a loaded multiplexed bus.

In the receiver, noise may be added from the supply rail of terminations, and from instabilities of the receiver itself. In edge-triggered devices metastability can occur under unfortunate timing circumstances. In every high-gain device, self oscillation of the cell may occur if the input is biased (*e.g.*, due to reflections) near to the trip point.

In Section II, the effect of discrete capacitive loads is discussed. Section III describes reactive-loaded stubs, while Section IV gives a simple 'butterfly' model to analyse backplane stub ringing. Section V contains measurement considerations.

II. Transmission Line with Discrete Capacitive Loads

Buses are typically composed of transmission line(s) with several loads. If the bus is routed *e.g.*, on a printed-circuit board, and its loads are connected in a series manner without stubs and/or connectors, the bus can be modelled as shown in Fig. 3. Except when considering glitches, only one transmitter on the bus is active at a time. The loads are located randomly along the transmission line. Loads between which the propagation delay is much less than the signal rise time, can be modelled as a single lumped load. The transmission line is described by its characteristic impedance and unloaded propagation delay, its loss is neglected.

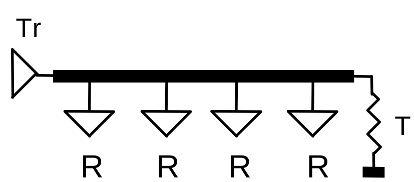

Fig. 3. Functional model of a single bus trace with multiple loads. Tr: transmitter, R: receiver, T: termination.

In situations when the transmitter is located somewhere in the middle of the bus, the propagation path can be considered as a source driving two separate bus sections. To exclude the interaction between the two bus sections, we assume a source impedance which is significantly less than the line impedance. Note that this is not true when the active state of the driver represents high impedance, namely the high-impedance state of TTL and BTL drivers.

State-of-the-art receivers represent a capacitive load, the resistive part of the load is neglected. This leads us to the equivalent circuit of loaded bus trace shown in Fig. 4.

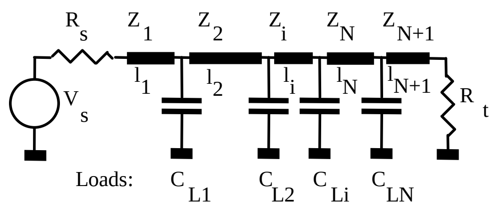

Fig. 4. Equivalent circuit of a single bus trace with multiple loads.

The series-placed resistor R_S may represent the source impedance of driver (usually neglected), or an intentional series termination resistor, either internal or external to the driver logic device.

The loaded line impedance, and propagation delay of the *i*th segment of the loaded lossless transmission line is approximated by

$$Z_0^{'} = \sqrt{\frac{L_i}{(C_i + C_{Li})}} = \frac{Z_0}{\sqrt{(1+\frac{C_{Li}}{C_i})}} \qquad (7)$$

$$t_{pd}^{'} = \sqrt{L_i(C_i + C_{Li})} = t_{pd}\sqrt{(1+\frac{C_{Li}}{C_i})} \qquad (8)$$

where C_i, and L_i are the transmission-line capacitance and inductance for the length of l_i. Note that due to loading, characteristic impedance goes down, propagation delay goes up by the same proportion determined by the ratio of C_{Li}/C_i.

We have to remember though that the unit inductance and unit capacitance of transmission line represent distributed components, while the loads in the model are discrete capacitances. Therefore the loaded transmission line of Fig. 4. can not be matched any more in the strict sense, in other words there is no such value of the termination resistor which would equal the input impedance regardless of frequency.

For a simplified model, the characteristic impedances and the physical lengths of the transmission-line segments, as well as the load capacitances are the same:

$$Z_1 = Z_2 = Z_i = Z_N = Z_0 \qquad (9)$$

$$l_1 = l_2 = l_i = l_N = l \qquad (10)$$

$$C_{L1} = C_{L2} = C_{Li} = C_{LN} = C_L \qquad (11)$$

It is not possible to obtain a closed-form analytical solution to the time-domain response of this loaded line. Characteristics of the loaded bus trace were therefore analysed by numerical analysis. The circuit is shown in Fig. 5.

Numerical values of Fig. 5. may represent a typical printed-circuit board trace with equally spaced loads (transceivers). Note that the load capacitance at each segment is three times larger than the unit capacitance of the transmission-line segment, therefore the square-root correction terms in (7) and (8) equals 2, resulting in $Z_0^{'} = 30\ \Omega$, and $t_{pd}^{'} = 0.36$ ns per segment. The $R_t = 30\ \Omega$? termination resistance 'matches' the loaded line impedance. Fig. 6. illustrates the time-domain waveform of the circuit at some of the 30 nodes. The ringing clearly shows that wideband matching of the discretely loaded transmission line is not possible.

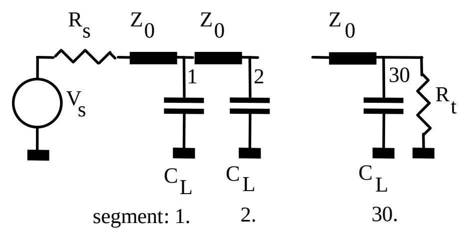

Fig. 5. Circuit schematic of the loaded transmission-line model used for numerical analysis. Number of segments: 30. $Z_0 = 60\ \Omega$ $t_{pd} = 0.18$ *ns / segment* ($C_0 = 3$ *pF / segment,* $L_0 = 10.88$ *nH / segment),* $R_t = 30\ \Omega$, $C_L = 9$ *pF. The source voltage is a piecewise linear step waveform with 10 ps risetime and 1 V amplitude.*

Further analysis of the transient response reveals a gradual degradation of signal rise time (Fig. 7) and extra delay (Fig. 8) along the loaded line. It can be seen from the waveforms of Fig. 6 that the linear input ramp is distorted, and the actual waveform varies from node to node. Rise-time figures of Fig. 7 are therefore given in a normalised manner. It shows 0%-to-100% rise times of equivalent linear ramps with slopes equal to that of the actual waveform of the particular node at 0.5 V, *i.e.*, at 50% of the steady-state value.

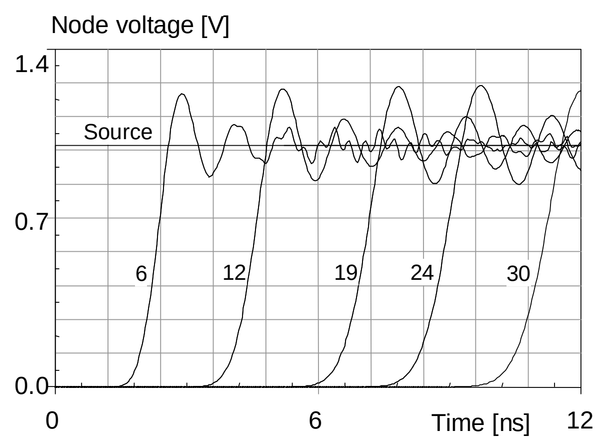

Fig. 6. Time-domain response of a series of capacitively loaded transmission lines as shown in Fig. 5. Source voltage is a 1-V linear ramp with 10 ps rise time. Load capacitance is three times the inherent capacitance of transmission line. $R_s = 0$, $R_t = Z_0'$.

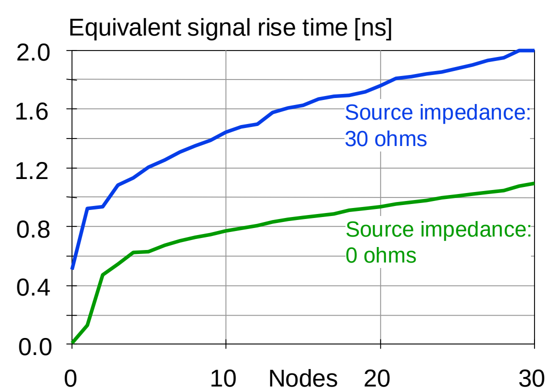

Fig. 7. Rise-time degradation along a transmission line with uniformly spaced capacitive loads.

There is a significant difference with the source resistance, too. Simulation results with $R_S = 30\ \Omega$ represent a series-matched line, while $R_S = 0\ \Omega$ corresponds to low-ohmic TTL drive.

The extra delay due to discrete capacitive loads should not be neglected when timing is critical. It is important to note that the extra delay shown in Fig. 8. is referenced to the 50% steady-state value. Lower device trip point will result in shorter delay.

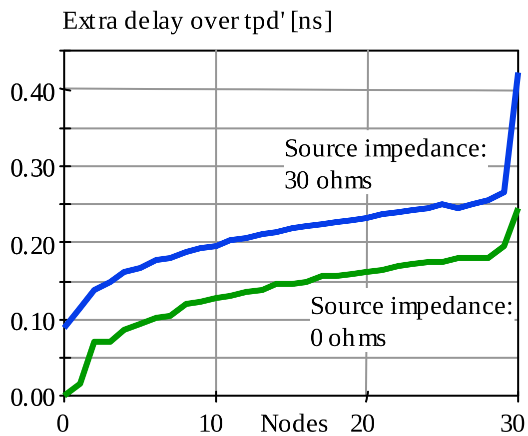

Fig. 8. Extra delay in excess of t_{pd}' (calculated according to (6)) of a capacitively loaded transmission line along the nodes with and without series matching resistor. The delay is measured at 50% of the steady-state response.

The rise-time degradation is important when calculating input skew of devices, and/or when receivers with different trip points are attached to the line.

III. Reactive-loaded Stubs

It is well known to logic designers that the time-domain response at the open end of the circuit in Fig. 2b to an ideal step excitation is an infinite stream of square waves with a period of $T = 4\ t_{pd}$. This phenomenon in digital circuits is called stub ringing, the peak-to-peak amplitude of oscillation is $2\ V_S$. Damping factor of the oscillation is zero whenever source impedance is zero and far-end termination is open.

In the frequency domain, the fundamental frequency of stub ringing corresponds to the first zero of input impedance. The input impedance of a transmission line is given by:

$$Z_{in} = Z_0 \frac{Z_L + jZ_0 tg(\phi)}{Z_0 + jZ_L tg(\phi)} \qquad (12)$$

where Z_{in} is the input impedance of the transmission line terminated by an impedance of Z_L. Φ is the electrical length of the transmission line:

$$\phi = \omega t_{pd} \qquad (13)$$

where $\omega = 2\pi f$, and f is the frequency of observation. With capacitive loads, (12) results in:

$$Z_{in} = jZ_0 \frac{\frac{\tau_L}{t_{pd}} \phi\ tg(\phi) - 1}{\frac{\tau_L}{t_{pd}} \phi + tg(\phi)} \qquad (14)$$

where $\tau_L = Z_0 C_L$. (15)

If $C_L = 0$, (14) yields the simple form:

$$Z_{in} = \frac{Z_0}{j\omega C} \quad (16)$$

Comparing (13) and (16), it is easy to see that zeros of (16) are at the integer multiple of $1/(4t_{pd})$, what is according to our expectations. Similarly, with a nonzero capacitive load, we would expect that the fundamental frequency of stub ringing is given by $1/(4t_{pd}')$. However, solving the nominator of (14) to find its roots reveals that the ringing frequency is lower than $1/4t_{pd}'$. The physical cause of this phenomenon is well known in microwave engineering as the increase of the transmission line's electrical length when the line is capacitively loaded. Fig. 9 shows the stub ringing frequency normalised to $1/4t_{pd}'$, as a function of load capacitance normalized to the total inherent capacitance of transmission line. Data shown in Fig. 9 was verified by time-domain simulation of ringing frequency.

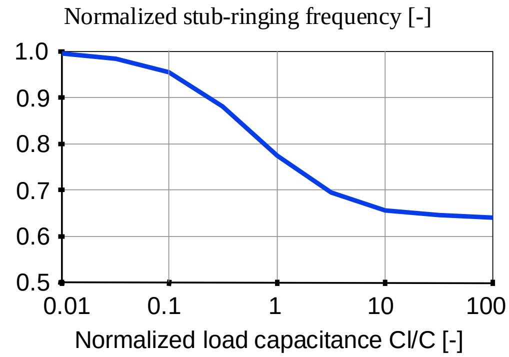

Fig. 9. Ringing frequency of a capacitively loaded single stub as a function of load capacitance. Ringing frequency is normalised to $4t_{pd}'$, load capacitance is normalised to the total inherent transmission-line capacitance of stub.

Stub ringing is not a problem even if the waveform crosses the threshold (trip point) of the attached devices as long as the period of the ringing is so short that the attached device(s) can not respond to it. In this respect, lower ringing frequency is worse, therefore the correction factor shown in Fig. 9 should be considered in every critical application.

Unfortunately logic devices are not adequately characterized in terms of transient suppression. This fact underlines that stubs in high-speed buses should be short, and stub-load capacitances should be small in order to increase the stub-ringing frequency.

IV. Transmission Line with Stub Loads

Bus lines with plug-in boards will necessarily have stubs due to connectors and printed-circuit-board traces going to loads. A simplified backplane schematic is shown in Fig. 10.

Even if the backplane trace (transmission line) is properly terminated to avoid reflections on the main path, the stubs do not (and should not) have matched terminations. In a typical data-transfer situation, one of the transceivers is active, terminating the stub with low impedance (in both states of TTL, and in low state of BTL), all other transceivers represent capacitive loads. Except the short duration of transitions of transmitters, the nonlinearities of the system can be neglected. (This is not true though when at one or more of the receivers or at the termination(s) intentional or not intentional clipping of the waveform occurs). In the linearized model, the overall transient response will be the sum of signal reflections (if any) on the backplane trace, and stub-ringing waveforms from the various combinations of stubs.

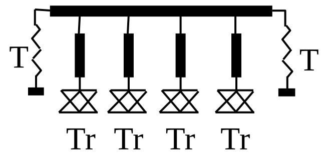

Fig. 10. Simplified schematic of a backplane trace with stubs. T: termination, Tr: transceiver.

Schematic of Fig. 10 can be further simplified to find the basic model of backplane stub ringing. Fig. 11 shows the backplane trace with its terminations and only two (arbitrarily) selected stubs. Response of a full backplane configuration will be the superposition of the individual responses of such segments. For the sake of simplicity, the two stubs (including the load capacitances) are assumed to be equal. It is easy to see that the "butterfly" structure of Fig. 11.b is prone to ringing, regardless of the value of termination, since with the two parts of the structure being the same, the resulting standing-wave pattern has a null at the center node. The ringing frequency is the stub-ringing frequency of one of the loaded stubs, as given in Section III.

The transient response of the more realistic PI model of Fig. 11.a depends on the backplane-trace length between the stubs.

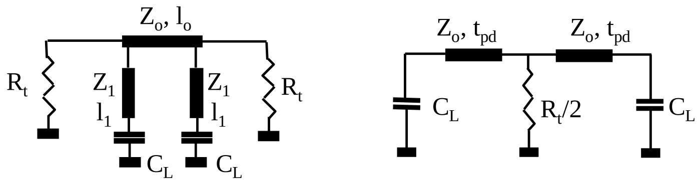

Fig. 11. Schematics of basic linearized models for backplane stub ringing: PI model on left, butterfly model on right.

V. Measurement Considerations

When measuring high-speed logic circuits, care should be taken to use a measuring system with adequate bandwidth, and at the same time to avoid excess loading of the circuit.

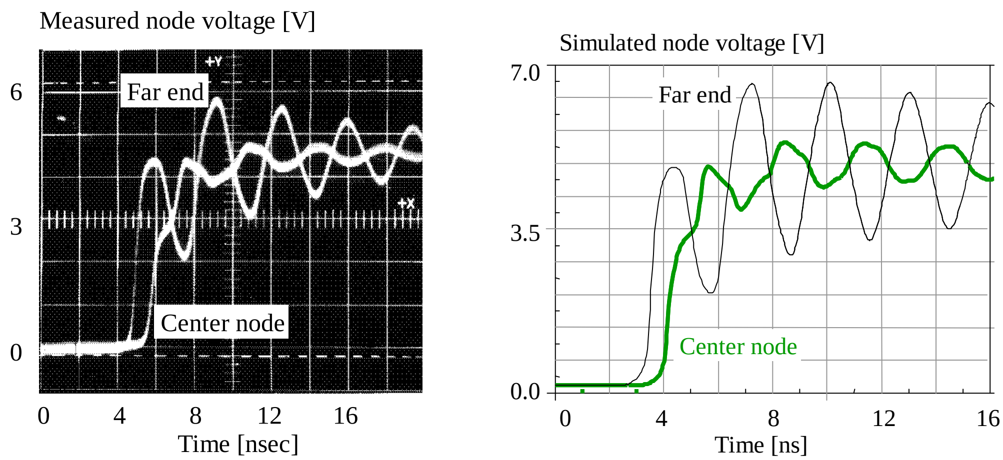

Fig. 12. Simulation and measurement results of a butterfly dual-stub segment of the backplane model: 2 x 5.6" 50-ohm stubs, 50-ohm termination connected to +5V. Drive signal: low-state to high-impedance state at one end of the butterfly. Waveforms are shown at the center node (smooth) and at the open end (ringing). Horizontal scale is 2 ns per division for both graphs.

Passive high-impedance probes have an input capacitance in the order of 10 pF, which is unacceptably high for measuring short (1 to 2 inches) transmission-line segments. To avoid significant resistive loading, and at the same time minimize capacitive loading, passive low-impedance probes can be used. Probes with a simple resistive divider may have an input resistance of 5000 ohms with an input capacitance of less than 1 pF. In a 50-ohm transmission system, the input resistance will result in a reflection coefficient of 1%, which is negligible compared to other sources of reflections.

Fig. 12. shows the measurement and simulation waveforms of the 'butterfly' model of Fig. 11.b. Measurement waveforms were taken with an oscilloscope of 1 GHz bandwidth, with 5000-ohm passive probes. Source of excitation was one segment of an FCT245 octal transmitter. The transceiver was connected to one open end of the butterfly, and was driven from low state to high-impedance state. Termination at the center node of butterfly was 50 ohms. The measurement waveform of Fig. 12.a, and the simulated waveform of Fig. 12.b show good correspondance.

VI. Conclusions

Transmission-line effects in high-speed logic designs must be taken into account. On bus lines without connectors and stubs, the major source of signal degradation is the reflection from the individual (capacitive) loads. When propagation delay between loads is much less than signal rise time, a gradual degradation of signal occurs along the line. Data are presented on the degradation of signal rise time, as well as on the propagation delay in excess of the delay calculated from quasi-lumped models.

Capacitively loaded stubs generate ringing with low damping factor when driven from a low source impedance. The ringing frequency becomes lower with increasing capacitive load. Normalised ringing frequency is presented as a function of normalised load capacitance.

Backplanes necessarily have stubs in the form of connectors and connecting traces. Simple PI and butterfly models are given to predict stub ringing on loaded backplanes.

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