Comparison of Methods for Determining the Capacitance of Planar Transmission Lines with Application to Multichip Module Characterization

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***Abstract*—Electrical characterization of transmission lines and discontinuities in multichip modules (MCM's) and other planar structure requires the use of a calibration procedure to establish reference measurement planes within the structure at the ports of the device under test (DUT). The preferred method of calibration uses one-port measurements of an arbitrary but accurately repeatable reflection standard and measurements of two or more lengths of line together with the per unit capacitance of the line. Several methods for evaluating the capacitance of transmission lines are compared here as the overall accuracy of electrical characterization is directly dependent on this determination.**

I. Introduction

EXPERIMENTAL high frequency electrical characterization of transmission lines and discontinuities in multichip modules (MCM's) and other planar structures requires the use of controlled impedance measurement probes. These probes effect a low reflection transition from a coaxial-based automatic network analyzer system to the strip transmission line system of the device under test (DUT). Calibration of the measurement system to the probe tips is commonly achieved using a standard calibration substrate with known reference standards fabricated in a coplanar waveguide or microstrip on alumina or sapphire. However, MCM-D's and many other packaging structures rarely utilize such surface interconnects, see, for example, the longitudinal section of the MCM line shown in Fig. 1. Thus the effect of the transition from the probe tips, ports 1 and 2 in Fig. 1, to the DUT ports, ports 1′ and 2′ must be removed from measurements for accurate subsequent characterization of interconnects and discontinuities. This procedure of establishing the internal reference planes is referred to as *in situ* calibration, and this requirement means that standard calibration substrates cannot be used. Instead separate calibration structures must be provided on the DUT substrate or a similarly fabricated one.

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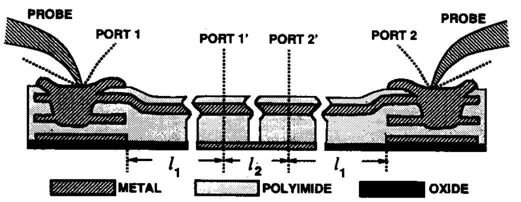

Fig. 1. Longitudinal section of MCM transmission line test structure constructed by tracing several scanning electron microscope photographs. Here l_1 is 0.498 mm and l_2 is 7.024 mm for the line section and zero for the through section. The structure was built on a silicon substrate and the oxide layer under the probe pads extends 50 μm from the center of the pad. The probes are model 40 Picoprobes from GGB Industries.

While three known loads are sufficient to establish the measurement reference plane, the preferred approach is to employ the through-reflect-line (TRL) method or similar procedure. In these techniques, two or more lengths of line are used instead of lumped standards, as the dispersion of the lines at high frequencies is much less than that of lumped reference standards. A special property of TRL-like procedures is that the propagation constant of the line can be determined from microwave [1] or time-domain transmission [2] measurements. When this is combined with the per unit length conductance G and capacitance C of the line, its complex propagation constant can be calculated [3]–[5]. Typically, however, G is assumed to be zero as it is generally small.

Several approaches have been proposed for the determination of C of a transmission line from microwave measurements [2], [5]–[7]. In [6], Marks and Williams use one or other of the two methods presented in [8] for determining the capacitance of a line from microwave measurements. One of these uses extrapolation of the microwave measurements of a line plus dc measured resistance to determine C. A second method uses microwave measurements of a terminated microwave transmission line to determine the proportionality factor in the TRL procedure and thus the capacitance of the line. In [7], Goldberg *et al.* use the high frequency asymptotic behavior of skin effect to determine a high frequency approximate of C. In the work of Deutsch *et al.* [2], direct measurement of capacitance by a capacitance meter is utilized.

The purpose of this paper is to compare the techniques for determining the capacitance of embedded interconnects in an MCM-D MCM. The error in estimating this capacitance is generally the greatest contributor to the overall error in determining the impedance of interconnects and discontinuities. Using *in situ* calibration we report microwave characterization of various MCM interconnects fabricated using an alumina on polyimide MCM-D process.

II. Background

In terms of its per unit length parameters, a transmission line is described by its propagation constant

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} = \frac{j\omega}{c} \sqrt{\mu_{r,\text{eff}} \epsilon_{r,\text{eff}}} \quad (1)$$

and characteristic impedance

$$Z_C = \sqrt{\frac{R + j\omega L}{G + j\omega C}} = Z_0 \sqrt{\frac{\mu_{r,\text{eff}}}{\epsilon_{r,\text{eff}}}} \quad (2)$$

where $\omega = 2\pi f$, $\mu_{r,\text{eff}}$ and $\epsilon_{r,\text{eff}}$ are the effective relative permeability and permittivity, respectively, f is frequency, R and L are the unit resistance and inductance of the line, and the free space impedance of the line with ideal conductors is

$$Z_0 = \sqrt{\frac{L_0}{C_0}} = \frac{1}{C_0 c}. \quad (3)$$

For nonmagnetic media

$$\mu_{r,\text{eff}} = \frac{R + j\omega L}{j\omega L_0} \quad (4)$$

and

$$\epsilon_{r,\text{eff}} = \frac{G + j\omega C}{j\omega C_0} \quad (5)$$

where L_0 and C_0 are the per unit inductance and capacitance of the line in the absence of dielectric. Combining (2), (3), and (5) the frequency dependent characteristic impedance is related to the propagation constant by

$$Z_C(f) = -j \frac{\gamma(f)}{\omega \epsilon_{r,\text{eff}} C_0}. \quad (6)$$

Thus since γ can be determined from measurement of a through and line as a by-product of the conventional TRL calibration procedure [1], Z_C can be found once $\epsilon_{r,\text{eff}}$ is determined. In the absence of dielectric loss, $\epsilon_{r,\text{eff}} = C/\omega C_0$ and (6) becomes [6]

$$Z_C(f) = -j \frac{\gamma(f)}{\omega C}. \quad (7)$$

III. Capacitance Determination

A. Method A. Direct Measurement

While direct measurement of the line capacitance and conductance seems the obvious approach, it is complicated by the test line being invariably short as little real estate can generally be devoted to a single set of test structures and their relatively large probe pads. This is particularly so if a standard set of test structures is to be included as a test coupon on every wafer. Two lengths of line are required to remove the capacitive and conductive effects of transitions to the line, and practically, these lines must be those used in one of the TRL-like microwave calibration procedures. Thus the lines will generally be less than a centimeter long and have capacitances of a picofarad or less. Since measurements must be made at a low enough frequency to ensure that the lines have negligible electrical length and so can be treated as lumped capacitances, the required measurements are at the limits of performance of commercial instruments.

B. Method B. High Frequency Approximation

A high frequency estimate of the line capacitance can be obtained by using the assumed frequency dependence of the per unit components of the line [9]: C and G/ω are approximately frequency independent, R has a dc component of R_{dc} and a skin effect component R_S which has an assumed ω^δ dependence where $\delta$is around 0.5 when the skin effect is fully established, and $L = L_0 + L_{\text{int}}$. L_{int} is the internal conductor inductance of the line due to current internal to the conductors and is asymptotically zero at high frequencies as the skin effect is fully established. Consequently:

$$\mu_{r,\text{eff}} = \frac{R + j\omega(L_0 + L_{\text{int}})}{j\omega L_0} \quad (8)$$

and

$$\epsilon_{r,\text{eff}} = \frac{G + j\omega C}{j\omega C_0}. \quad (9)$$

Since R increases sublinearly because of the skin effect, $R \ll \omega L$ at a high frequency, f_H, where $L \approx L_0$ and so

$$\lim_{f \to f_H} \mu_{r,\text{eff}}(f) = 1 - \frac{jR}{\omega L_0} \quad (10)$$

Thus

$$\epsilon_{r,\text{eff}} = \lim_{f \to f_H} \hat{\epsilon}_{r,\text{eff}}(f) \quad (11)$$

where the intermediate dielectric constant

$$\hat{\epsilon}_{r,\text{eff}}(f) = \frac{-\gamma^2(f) c^2}{\omega^2 [1 - jR/(\omega L_0)]} \quad (12)$$

is obtained by substituting for $\mu_{r,\text{eff}}$ in (1) and rearranging. Combining (9) and (12):

$$C - j\frac{G}{\omega} = \frac{-c^2 C_0}{\omega^2} \lim_{f \to f_H} \left\{ \frac{\gamma^2(f)}{[1 - jR/(\omega L_0)]} \right\}. \quad (13)$$

Since $R \ll \omega L_0$ at high frequencies:

$$C - j\frac{G}{\omega} \approx \frac{-c^2 C_0}{\omega^2} \lim_{f \to f_H} \{\gamma^2(f)[1 + jR/(\omega L_0)]\}. \quad (14)$$

That is

$$C \approx \frac{-c^2 C_0}{\omega^2} \lim_{f \to f_H} \left\{ \mathrm{Re}[\gamma^2(f)] - \frac{\mathrm{Im}[\gamma^2(f)]R}{\omega L_0} \right\} \quad (15)$$

and

$$\frac{G}{\omega} \approx \frac{c^2 C_0}{\omega^2} \lim_{f \to f_H} \left\{ \mathrm{Im}[\gamma^2(f)] + \frac{\mathrm{Re}[\gamma^2(f)]R}{\omega L_0} \right\}. \quad (16)$$

With the frequency dependence assumed previously at high frequencies, $\mathrm{Im}[\gamma^2(f)]$ and $\mathrm{Re}[\gamma^2(f)] \sim O(\omega^2)$, and $R \sim O(\omega^\alpha)$ with $\alpha \sim 0.5$. Thus since $\mathrm{Im}[\gamma^2(f)]$ is always positive and $\mathrm{Re}[\gamma^2(f)]$ is negative at high frequencies:

$$C \geq \frac{c^2 C_0}{\omega^2} \lim_{f \to f_H} \mathrm{Re}[\gamma^2(f)] \quad (17)$$

and

$$\frac{G}{\omega} \leq \frac{-c^2 C_0}{\omega^2} \lim_{f \to f_H} \mathrm{Im}[\gamma^2(f)]. \quad (18)$$

In summary, the basic assumptions behind (17) and (18) are that the skin effect is fully established and that the skin effect resistance increases sublinearly with frequency. At sufficiently high a frequency

$$C \approx \frac{c^2 C_0}{\omega^2} \mathrm{Re}[\gamma^2(f)] \quad (19)$$

and

$$\frac{G}{\omega} \approx \frac{c^2 C_0}{\omega^2} \mathrm{Im}[\gamma^2(f)]. \quad (20)$$

C. Method C. Low Frequency Extrapolation (R_{dc} Method)

Williams and Marks [8] presented two methods for the determination of the capacitance of transmission lines using low frequency network parameter measurements. The first of these methods uses the measured propagation constant γ: squaring (1):

$$\gamma^2 = (R_{\mathrm{dc}} + R_S)G - \omega^2 LC + j\omega((R_{\mathrm{dc}} + R_S)C + LG)$$

and rearranging

$$(R_{\mathrm{dc}} + R_S)C + LG = \mathrm{Im}(\gamma^2)/\omega. \quad (21)$$

Ignoring dielectric loss and R_S, which is reasonable at low frequencies

$$C = \lim_{f \to f_0} \frac{\mathrm{Im}(\gamma^2)}{R_{\mathrm{dc}}\omega}. \quad (22)$$

D. Method D. Low Frequency Estimate ($R_{\mathrm{load,dc}}$ Method)

The second technique[1] presented in [8] determines the capacitance of a line by noting that Z_C is identical to the reference impedance, Z_{ref}, of the TRL measurement procedure. For a small lumped load resistor, $R_{\mathrm{load,dc}}$, at low frequencies [8]:

$$Z_{\mathrm{ref}}\left(\frac{1+\Gamma_{\mathrm{load}}}{1-\Gamma_{\mathrm{load}}}\right) \equiv Z_C\left(\frac{1+\Gamma_{\mathrm{load}}}{1-\Gamma_{\mathrm{load}}}\right) = Z_{\mathrm{load}} \approx R_{\mathrm{load,dc}} \quad (23)$$

where Γ_{load} is the measured reflection coefficient of the load referred to Z_{ref}. Substituting (23) in

$$\gamma/Z_C = G + j\omega C \quad (24)$$

results in

$$C\left[1 - j\left(\frac{G}{\omega C}\right)\right] \approx \frac{\gamma}{j\omega R_{\mathrm{load,dc}}}\left(\frac{1+\Gamma_{\mathrm{load}}}{1-\Gamma_{\mathrm{load}}}\right). \quad (25)$$

That is

$$C \approx \frac{\mathrm{Im}[\gamma]}{\omega R_{\mathrm{load,dc}}}\left(\frac{1+\Gamma_{\mathrm{load}}}{1-\Gamma_{\mathrm{load}}}\right) \quad (26)$$

and

$$\frac{G}{\omega} \approx \frac{\mathrm{Re}[\gamma]}{\omega R_{\mathrm{load,dc}}}\left(\frac{1+\Gamma_{\mathrm{load}}}{1-\Gamma_{\mathrm{load}}}\right). \quad (27)$$

IV. Results and Discussion

The four approaches to *in situ* capacitance determination are compared in Fig. 2 for the MCM shown in Figs. 1 and 3. In Fig. 2 the curve labels correspond to the methods discussed in the previous section. The exception is curve C^F which is a curve fit to curve C assuming a $\sqrt{\omega}$ dependence of the skin effect component of the line resistance. This curve indicates a dc intercept of 81.5 pF/m for method C. This compares favorably to the direct measurement (method A) of 82.4 pF/m. However, the low frequency extrapolated value of C is dependent on the frequency range over which (22) is fitted (see Table I) owing principally to the measurement scatter and uncertainties as to the actual frequency dependent behavior of R_S. The high frequency asymptotic estimation of C, method B, and the low frequency $R_{\mathrm{load,dc}}$ method, curve D, also compare favorably to the other C determinations. The results are summarized in Table I.

The capacitance meter used (an HP4280A precision LCR meter) did not allow the conductive line loss to be determined. However, error tolerancing establishes an upper bound on G of 30 μS/m at 1 MHz. The high frequency asymptote method (B) yields an upper bound on the line conductance of 48 μS/m. No indication of G could be obtained from the low frequency extrapolation methods (C and D) because of the scatter of the measured data.

The errors of methods A, C, and D will reduce as the test line lengths increase, but the error of method B will

[1] Earlier comments made by one of the authors about this technique being incorrect [10] are wrong [11].

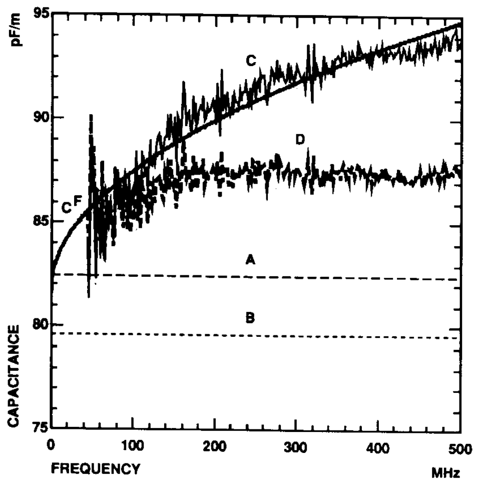

Fig. 2. Capacitance approximations: A: 1-MHz capacitance measurement (method A); B: high frequency asymptotic capacitance (method B); C: measured $\mathrm{Im}(\gamma^2)/\omega$ (method C); (C^F) curve fitted to C; and D: capacitance calculated using microwave measurements and dc load resistance (method D).

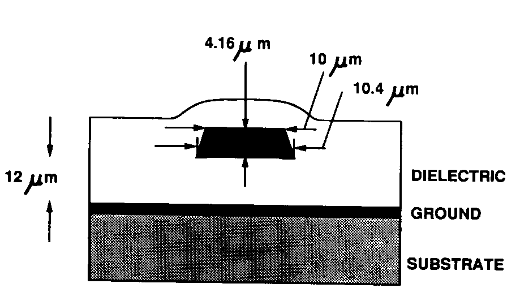

Fig. 3. Embedded microstrip cross section.

TABLE I
CAPACITANCES DETERMINED USING VARIOUS MEASUREMENT TECHNIQUES. THE ERROR BUDGETS ARE GIVEN IN TABLE II.

Method	C (pF/m)	Error (pF/m)
A, Direct 1 MHz measurement	82.4	±1.5
B, High Frequency Asymptote	79.6	> ±4.1
C, R_{dc} method [8]		
45—400 MHz	80.3	> ±4.5
45—500 MHz	81.5	> ±4.5
45—800 MHz	83.7	> ±4.5
45—1000 MHz	84.3	> ±4.5
45—1245 MHz	84.3	> ±4.5
D, $R_{\mathrm{load,dc}}$ method [8]	85	> ±4.1

TABLE II
ERROR BUDGETS FOR CALCULATING THE TOTAL ROOT SUM SQUARED (RSS) ERROR OF THE VARIOUS CAPACITANCE AND CONDUCTANCE MEASUREMENTS

Method/Line Item	Notes	C Error (pF/m)
A, 1-MHz Capacitance measurement		
Capacitance measurement (Through) (192 ± 5 fF)	1	±0.7
Capacitance Measurement (Line) (771 ± 9 fF)	1	±1.3
Line length (Through) (996 ± 10 μm)	2	±0.1
Line length (Line) (8020 ± 10 μm)	2	±0.1
TOTAL RSS ERROR		±1.5
B, High Frequency Asymptote		
Propagation constant determiniation excluding line length (1%)	3	±0.8
Line length (Through) (996 ± 10 μm)	2	±0.1
Line length (Line) (8020 ± 10 μm)	2	±0.1
Cross-Sectional Dimension ±0.5 μm		±4.0
Extrapolation validity		?
TOTAL RSS ERROR		> ±4.1
C, Low Frequency Extrapolation R_{dc} method [8]		
Propagation constant determination excluding line length (1%)	3	±0.8
Line length (Through) (996 ± 10 μm)	2	±0.1
Line length (Line) (8020 ± 10 μm)	2	±0.1
Extrapolation validity (estimated)	4	> ±4.5
TOTAL RSS ERROR		> ±4.6
D, Low Frequency Extrapolation $R_{\mathrm{dc,load}}$ method [8]		
Propagation constant determination excluding line length (1%)	3	±0.8
Line length (Through) (996 ± 10 μm)	2	±0.1
Line length (Line (8020 ± 10 μm)	2	±0.1
Fixture and extrapolation error (estimated)	5	> ±4.5
TOTAL RSS ERROR		> ±4.6

[1]Based on accuracy of HP4280A measuring the capacitance of the through (192 ± 5 fF) and the line (771 ± 9 fF).
[2]Based on accuracy of measuring the through (996 ± 10 μm) and the line (8020 ± 10 μm).
[3]Estimated based on the quoted accuracy of the S parameter measurements and scatter in the calculated propagation constant data.
[4]Estimated based on scatter of low frequency data and dependence of extrapolated data on frequency range used.

change only slightly as the principle component of error is making the cross section measurements from scanning electron microscope photographs. While the four methods determine C within a ±3% spread, this translates to a corresponding ±3% spread in the calculation of the characteristic impedance of the line [7] and of the extracted impedances of discontinuities [12]. The direct capacitance method has the lowest error, and what is particularly important, of the four methods it is the only one traceable to capacitance standards and has a well-defined error tolerance. The major difficulty in assigning tolerances to the other methods is that while the extrapolations employed are theoretically sound, their error can only be estimated. Error budgets of the four methods are given in Table II and summarized in Table I.

The accuracy of the assumptions behind the high frequency asymptotic method, method B, can be examined more closely by inspecting the asymptotic behavior of the intermediate relative permittivity $\hat{\epsilon}_{r,\mathrm{eff}}$ shown in Fig. 4. Note that $\hat{\epsilon}_{r,\mathrm{eff}}$ is only approximately asymptotic at high frequencies. This is attributed to the effect of the high resistance of the MCM interconnect. By comparison, the high frequency behavior of $\hat{\epsilon}_{r,\mathrm{eff}}$ was found to be asymptotic for PCB interconnects which have much lower resistive loss, see [1].

The extrapolation validity errors of methods C and D is based on the observed scatter of the low frequency data. This scatter is not evident when a standard calibration substrate is used [6]. In this case the calibration structures are in a coplanar waveguide so that the transition from the probe to the measurement substrate is negligible, and the dimensional tolerancing is tight. In contrast, the MCM-D *in situ* calibration structures used here have poor dimensional tolerances and the

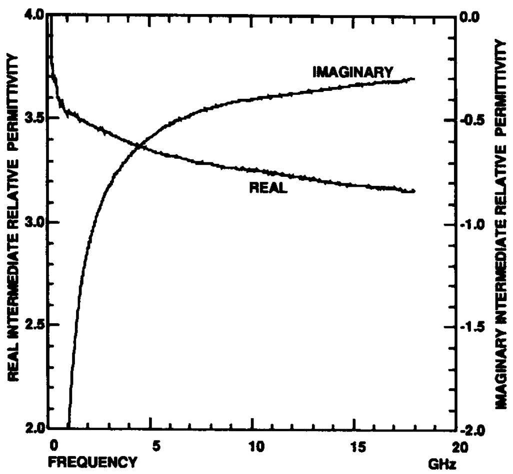

Fig. 4. Intermediate effective permittivity $\hat{\epsilon}_{r,\text{eff}}(f)$.

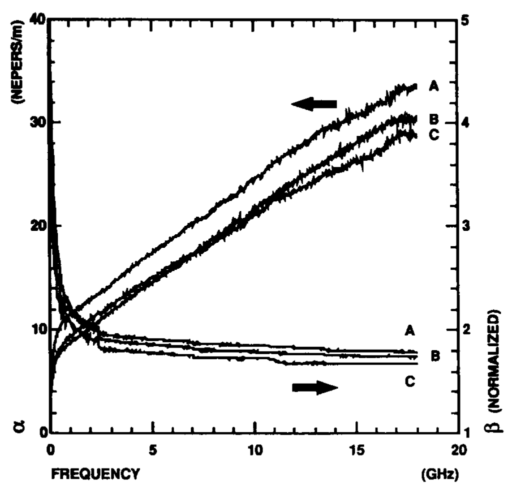

Fig. 5. Propagation constant of an aluminum on polyimide microstripline for three different linewidths: *A*: 10 μm; *B*: 16 μm; and *C*: 32 μm. The other dimensions are as given in Fig. 2.

effect of the transition is large, and at 1 MHz has a shunt capacitance of 96 fF and a series resistance of 0.33 Ω.

The propagation constant, γ, of several aluminum on polyimide lines of various widths are shown in Fig. 5 where the attenuation constant $\alpha = \text{Re}[\gamma]$ and normalized $\beta = \text{Im}[\gamma]/\beta_0$ are plotted. The constant $\beta_0 = \omega\sqrt{\mu_0\epsilon_0}$ is the free space propagation phase constant. Even at the highest frequencies the normalized β is not constant, indicating significant dispersion and putting into doubt the assumed frequency independence of C. The direct capacitance measurement was used in conjunction with the γ's (7), to obtain the Z_C shown in Fig. 6. The imaginary part of Z_C approaches a constant at high frequencies. This is not consistent with the usually accepted view that the conductance G is negligible. Indeed, G may not be solely due to dielectric loss as dimensional variations can excite radiation. Further developments of characterization techniques are required to properly measure and account for

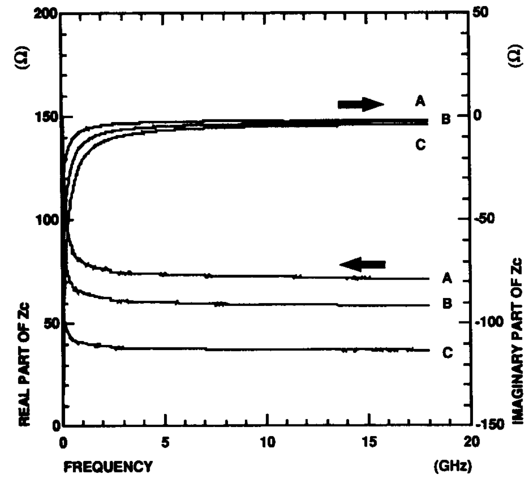

Fig. 6. Characteristic impedance of an aluminum on polyimide microstripline for three different linewidths: *A*: 10 μm; *B*: 16 μm; and *C*: 32 μm. The other dimensions are as given in Fig. 2.

finite G and possible small frequency dependence of C.

V. Conclusion

Four methods for determining the capacitance of the line were considered along with the assumptions behind the methods and their sources of error. It was determined that the constant capacitance and negligible conductance assumptions for interconnects on MCM-D's may not be valid. Further work is required to permit more accurate *in situ* calibration for the characterization of embedded interconnects and discontinuities.

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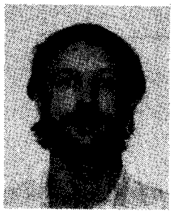

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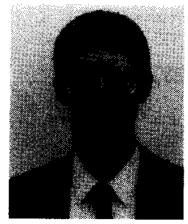

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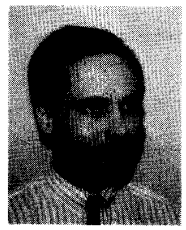

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