Experimental Characterization of Transmission Lines in Thin-Film Multichip Modules

Steven Lipa, Michael B. Steer, *Senior Member IEEE*, Andreas C. Cangellaris, *Member IEEE*, and Paul D. Franzon, *Member IEEE*

***Abstract*—Signal propagation on transmission lines fabricated in thin polyimide films on silicon substrates is investigated. Series resistive and shunt conductive losses are separated and it is shown that the effective dielectric loss is much higher than that expected from bulk material properties.**

***Index Terms*—Lossy transmission lines, thin-film MCM, polyimide dielectric.**

I. Introduction

THE design of controlled impedance interconnects in thin-film multichip modules (MCM-D's) and the subsequent circuit-level simulation of the package requires accurate modeling of microstrip and stripline transmission lines. Current practice is to use established planar transmission line models developed for conventional microwave circuit analysis. These models have been experimentally verified for various families of lines fabricated on ceramic substrates which have isotropic dielectric properties. In contrast, polymer thin-films have in-plane/out-of-plane dielectric anisotropy which can result in leaky quasi-TEM mode propagation on thin-film interconnects [1]–[4]. With lines having uniform cross-sections, the frequency dependent per-unit R, L, and C of a line can be calculated using any of a large number of electromagnetic simulation tools. The conductance term, G, is often ignored or is calculated using the fields determined for a lossless line and the dielectric loss tangent of bulk substrate material.

The purpose of the work described here is to investigate the validity of such models for transmission lines in a specific MCM-D using Hitachi PIX-L112 polyimide dielectric. In particular, the effects of loose dimensional control and of the dielectric anisotropy of thin-films on loss were of interest. In this work, the dielectric loss characteristics are found using a through-reflect-line- (TRL-) like measurement procedure combined with calculation of the lines' resistance per unit length. The anomalously high effective dielectric loss obtained has important implications for the modeling of interconnects in MCM-D's based on bulk material properties.

Manuscript received October 22, 1993; revised October, 1995. This paper was presented in part at the IEEE 2nd Topical Meeting on Electrical Performance of Electronic Packaging, Monterey, CA, October 20–22, 1993. This work was supported in part by the National Science Foundation through Grant MIPS-9017054.

S. Lipa, M. B. Steer, and P. D. Franzon are with the Picosecond Digital Systems Laboratory, Department of Electrical and Computer Engineering, North Carolina State University, Raleigh, NC 27695-7911 USA.

A. C. Cangellaris is with the Department of Electrical and Computer Engineering, University of Arizona, Tucson, AZ 85721 USA.

Publisher Item Identifier S 1070-9886(96)00896-7.

The TRL-like procedure used here is the through-line (TL) technique [5] utilizing direct capacitance measurement, but, in principle, any TRL-like technique could be used.

II. Background

The experimental characterization of planar microwave interconnects and discontinuities first requires experimental modeling of the fixtures interfacing the coaxial measurement system to the planar interconnects. In the case of surface microstrip transmission lines and discontinuities, the state of the art for determining the two-port model of a fixture is represented by the multiple line TRL-like procedure of Marks and Williams [6]. In this technique the propagation constant, γ, of various lengths of a line is found as a by-product of a conventional TRL procedure [7]. Combining this with the shunt admittance $(G + \jmath\omega C)$ per unit length of the line yields the characteristic impedance, Z_0, of the line, the series impedance $(R + \jmath\omega L)$ per unit length, and subsequently the fixture parameters. The per unit admittance of the line is approximated as a capacitance determined as a low-frequency asymptote of microwave measurements or by direct measurement using a capacitance meter. Invariably, G is treated as being negligible, although, if it can be independently determined, it can be used to obtain a better estimate of R while having an insignificant effect on the L determination. Alternatively, as is shown below, evaluation of R enables G to be derived.

In terms of per-unit parameters

$$\gamma = \sqrt{(R + \jmath\omega L)(G + \jmath\omega C)} \tag{1}$$

and on squaring

$$\gamma^2 = RG - \omega^2 LC + \jmath\omega(RC + LG). \tag{2}$$

Partitioning this into real and imaginary parts and eliminating L leads to a quadratic equation for G

$$G^2 - G\frac{\mathrm{Re}(\gamma^2)}{R} + (\omega C)^2 - \frac{\omega C}{R}\mathrm{Im}(\gamma^2) = 0. \tag{3}$$

While the actual variation of C with frequency is not precisely known, it is likely that this variation is rather small. In this work, the value obtained by direct capacitance measurement of the transmission line at 1 MHz is used [8]. The variation of R with frequency was analytically determined using the combined finite-element/integral equation method described in [9]. Thus it is possible to solve for G as a function of frequency.

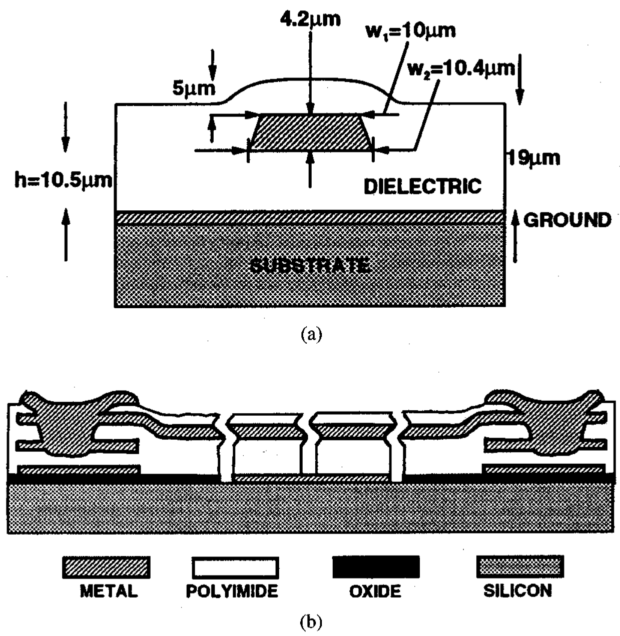


Fig. 1. SEM micrograph tracing of a 10 μm second layer metal (SLM) line: (a) cross-section and (b) longitudinal section.

III. Results and Discussion

The structures considered here are aluminum embedded microstrip transmission lines built on a silicon-substrate MCM using Hitachi PIX-L112 polyimide dielectric. Sections of a representative structure are shown in Fig. 1. The MCM contains both first (FLM) and second (SLM) layer metal lines with widths (w_1) of 10, 16, and 32 μm ($w_2 = w_1 + 0.4\mu$m). The three FLM lines are designated 1A, 1B, and 1C, respectively, and the SLM lines are similarly labeled 2A, 2B, and 2C. The differences between the FLM and SLM interconnects are that the line-to-ground-plane spacing is smaller (5 μm versus 10.5 μm) and the FLM lines have thicker dielectric above them.

The capacitance of the lines at 1 MHz was measured directly, using a Hewlett-Packard 4280A capacitance meter. The capacitances were:

- 1A 138 fF/mm;
- 1B 171 fF/mm;
- 1C 257 fF/mm;
- 2A 82.4 fF/mm;
- 2B 102 fF/mm;
- 2C 156 fF/mm.

A finite-element/integral equation analysis [9] was performed to calculate the variation of R with frequency for these lines. R was scaled by up to $\pm 15\%$ for each line so that the calculated DC resistance corresponded to that measured for that line. This scaling accounts for the tolerances on the measured cross-sectional dimensions of the lines and, in part, the effect of surface roughness. The results of this numerical analysis are shown in Fig. 2.

S-parameter data was taken using a Hewlett-Packard 8510B network analyzer over the frequency range of 45 MHz to 18 GHz, using Model 40 ground-signal-ground probes from GGB Industries with a TRL-like procedure [5] used to extract

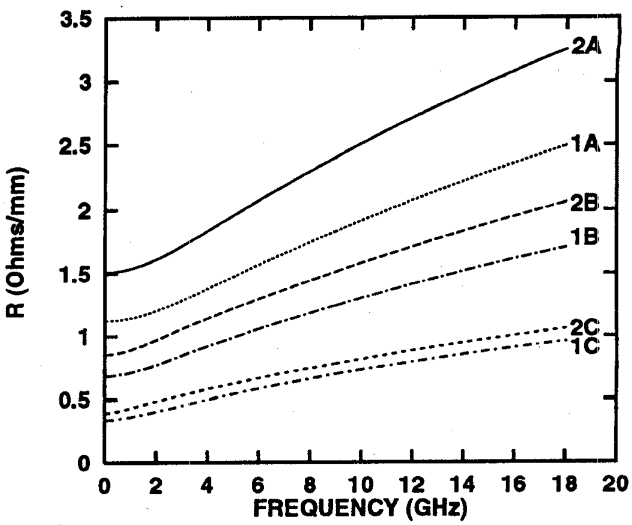


Fig. 2. Calculated R versus frequency.

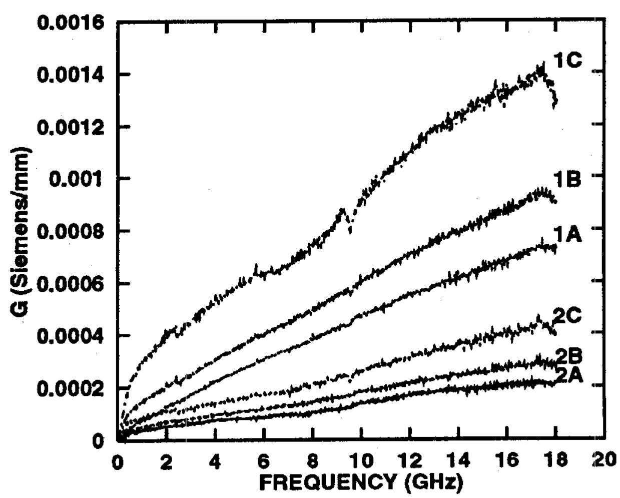


Fig. 3. Derived G versus frequency.

γ. Subsequently, the substrates were baked at 423 K for one hour, stored in a dessicator, and periodically remeasured over a period of several months with no discernible differences. Substituting R and C and these results into (3) leads to G as plotted in Fig. 3. $G \ll \omega C$ for all lines over the entire frequency range and therefore one might conclude that losses due to conductance are negligible. G accounts for the effective dielectric loss which can be seen by rewriting (1) as

$$\gamma = \jmath\omega\sqrt{LC}\left(1 - \jmath\frac{R}{\omega L}\right)^{\frac{1}{2}}\left(1 - \jmath\frac{G}{\omega C}\right)^{\frac{1}{2}} \tag{4}$$

and approximating the terms in parentheses by the first two terms of their power series expansions since $R \ll \omega L$ and $G \ll \omega C$ except at very low frequencies. The real part of the resulting equation is an approximation of the attenuation constant $\alpha = \alpha_C + \alpha_D$ where $\alpha_C = \frac{1}{2}R\sqrt{C/L}$ is due to conductor loss and $\alpha_D = \frac{1}{2}G\sqrt{L/C}$ represents the loss in the dielectric. The quantities α_C and α_D also include the effects of radiation loss.

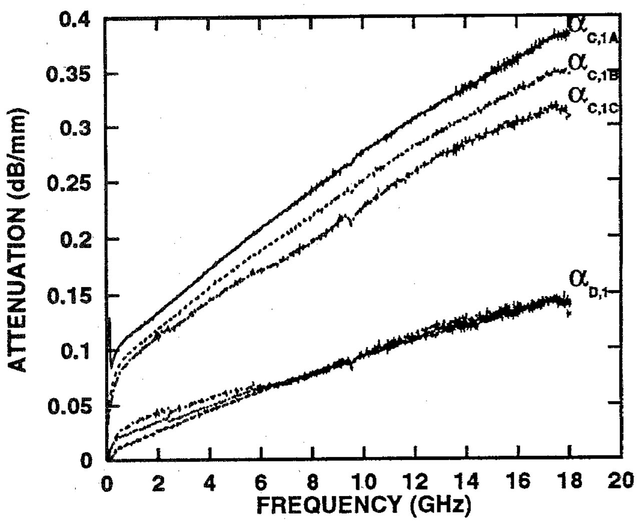


Fig. 4. Attenuation due to dielectric for first layer metal lines: $\alpha_{C,1A}$, $\alpha_{C,1B}$, and $\alpha_{C,1C}$ are the conductive attenuation constants for the 10 μm, 16 μm, and 32 μm width lines respectively; and $\alpha_{D,1}$ is the dielectric attenuation constant.

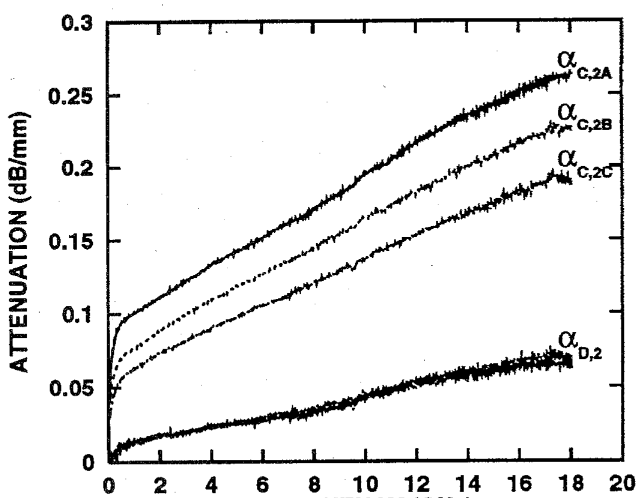


Fig. 5. Attenuation due to dielectric for second layer metal lines: $\alpha_{C,2A}$, $\alpha_{C,2B}$, and $\alpha_{C,2C}$ are the conductive attenuation constants for the 10 μm, 16 μm, and 32 μm width lines respectively; and $\alpha_{D,2}$ is the dielectric attenuation constant.

Plots of α_C and α_D for the FLM and SLM lines are shown in Figs. 4 and 5. Clearly α_D (which depends on G) makes up a significant proportion of the attenuation of each line even though $G \ll \omega C$. Significantly, the α_D's of the SLM and FLM lines coincide except at very low frequencies where the approximation to (4) breaks down. The dielectric loss constitutes a higher proportion of the total attenuation for the FLM lines than it does for the SLM lines, which is expected since more of the fields of the FLM lines are in the dielectric.

The behavior of α_D can be better understood by considering the relation between G and the dielectric loss tangent ($\tan\delta$) of a microstrip line [10, pp. 154–155]: $G/\omega C = k\tan\delta$. Here k is a constant called the filling factor ($0 \leq k \leq 1$), that depends on the ratio of the energy in the dielectric to the total energy with $k = 1$ indicating that the entire medium is dielectric. This relation was used to calculate the effective loss tangent for each of the lines under study. A straight-line fit to the G versus f data was made for each line in the range of 2 GHz to 16 GHz yielding $k\tan\delta$ of 0.046 for 1A; 0.046 for 1B; 0.046 for 1C; 0.021 for 2A; 0.024 for 2B; and 0.022 for 2C. This implies a minimum effective dielectric loss tangent of 0.046 as the maximum value of k is 1. The small nonzero linearly extrapolated G at DC is attributed to imprecise calculation of R, since the actual line cross section is not strictly trapezoidal, as well as to the breakdown of the separability approximation of α_C and α_D; see (4).

The above dielectric loss tangent is unusually high for polyimide if bulk properties alone are considered. To investigate this further, the dielectric loss tangent of the dielectric used was determined using a Hewlett Packard 4275A LCR meter (with 0.1 fF and 10 nS resolution) and the probing structure shown in Fig. 6. This structure was electrically modeled by the circuit shown in Fig. 7, where R_S is the contact resistance which must be found before the parallel RC equivalent circuit of the dielectric region between the ground plane and the bottom surface of the mercury can be determined. R_S is comprised of the resistance of the probing fixture, mercury contact resistance, and contact resistance between the ground probe to the ground plane. R_S was determined using two impedance measurements at different frequencies as follows. At frequency f_1 the impedance of the equivalent circuit is

$$\begin{aligned} Z_1 &= R_1 + jX_1 \\ &= R_S + \frac{1/G}{1+(\omega C/G)^2} - j\frac{\omega C/G^2}{1+(\omega C/G)^2} \end{aligned} \tag{5}$$

and so

$$R_1 = R_S + \frac{1/(\omega G_0)}{1+(C/G_0)^2} \tag{6}$$

where it is assumed that $G_0 = G/\omega$ is frequency independent. Multiplying (6) by f_1 and subtracting from it a similar expression at f_2 multiplied by f_2 leads to

$$R_S = \frac{f_1 R_1 - f_2 R_2}{f_1 - f_2}. \tag{7}$$

Measurements were made at a number of frequencies in the range 100 kHz to 10 MHz consistently yielding $R_S = 20\Omega$ and an out-of-plane loss tangent ($\approx G/\omega C$) of 0.006. This compares favorably with measurements on similar polyimides which implied an out-of-plane loss tangent of 0.005 [11]. (The in-plane loss tangent in [11] was 0.004.) It is clear, however, that the effective dielectric loss tangent of polyimide interconnects considered here far exceeds that of bulk polyimide. This excess loss could be due to radiation resulting from dimensional irregularities, or the effect of dielectric anisotropy.

IV. Conclusion

It has been shown that it is possible to determine the dielectric conductance per unit length of embedded microstrips

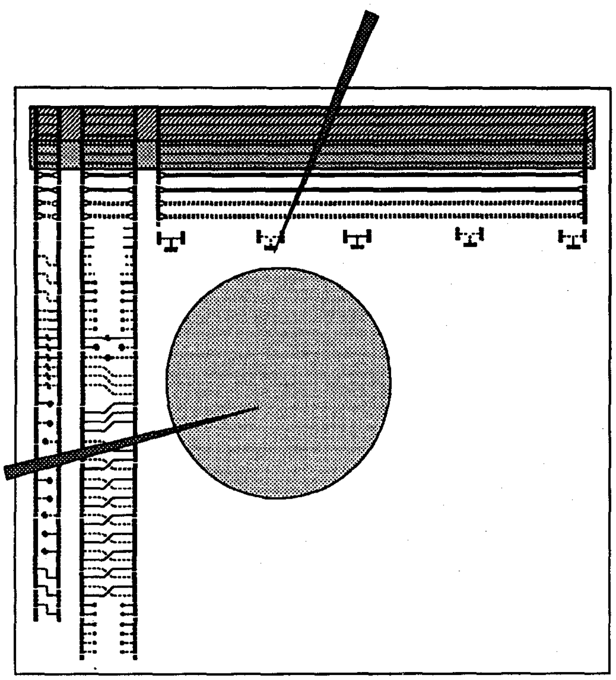


Fig. 6. Top view of low-frequency $\tan\delta$ experiment. The needle probes were electrically isolated from the micropositioners. One was immersed in a mercury droplet, and the other made ground contact through a 50 μm square surface pad. Also shown are the interconnect test structures.

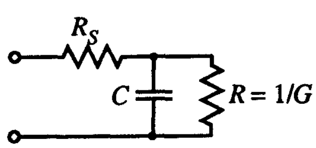


Fig. 7. Circuit model for finding contact resistance R_S and parallel equivalent circuit model of the dielectric between the ground plane and the bottom surface of the mercury.

on MCM substrates using a combination of measurements and analysis. In spite of the fact that the conductance is small compared to the line capacitance, it does contribute significantly to the attenuation of the line. The effective dielectric loss tangent of the polyimide far exceeds that of bulk polyimide. This has major implications for the modeling of interconnects in electrical simulators. At this stage, it appears that electrical modeling of narrow thin-film interconnects in MCM-D's must be calibrated to measurements rather than being based solely on geometrical considerations and bulk material properties.

REFERENCES

[1] A. A. Oliner, "Leakage from higher modes on microstrip line with application to antennas," *Radio Sci.*, vol. 22, no. 6. pp. 907–912, Nov. 1987.

[2] W. P. Harokopus and P. B. Katehi, "Radiation properties of microstrip discontinuities," in *IEEE Antennas Propag. Symp. Digest*, pp. 26–30, June 1989.

[3] T. S. Horng, S. C. Wu, H. Y. Yang, N. G. Alexopoulos, and P. B. Katehi, "A generalized method for the distinction of radiation and surface wave losses in microstrip discontinuities," in *IEEE MTT-S Int. Microwave Symp. Digest*, May 1990, pp. 1055–1058.

[4] W. P. Harokopus, P. B. Katehi, W. Y. Ali-Ahmad, and G. M. Rebeiz, "Surface wave excitation from open microstrip discontinuities," *IEEE Trans. Microwave Theory Tech.*, vol. 39, pp. 1098–1107, July 1991.

[5] M. B. Steer, S. B. Goldberg, G. Rinne, P. D. Franzon, I. Turlik, and J. S. Kasten, "Introducing the through-line deembedding procedure," in *1992 IEEE MTT-S Int. Microwave Symp. Digest*, June 1992, pp. 1455–1458.

[6] R. B. Marks, "A multiline method of network analyzer calibration," *IEEE Trans, Microwave Theory Tech.*, vol. 39, pp. 1205–1215, July 1991.

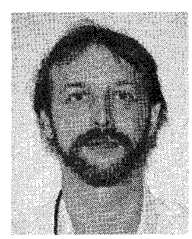
[7] R. B. Marks and D. F. Williams, "Characteristic impedance determination using propagation constant measurement," *IEEE Microwave Guided Wave Lett.*, vol. 1, pp. 141–143, June 1991.

[8] S. Lipa, M. B. Steer, A. S. Morris, and P. D. Franzon, "Comparison of methods for determining the capacitance of planar transmission lines with application to multichip module characterization," *IEEE Trans. Comp., Hybrids, Manufact. Technol.*, June 1993

[9] T. E. Van Deventer, P. B. Katehi, and A. C. Cangellaris, "An integral equation method for the evaluation of conductor and dielectric losses in high-frequency interconnects," *IEEE Trans. Microwave Theory Tech.*, vol. 37, no. 12, pp. 1964–1972, Dec. 1989.

[10] R. E. Collin, *Foundations for Microwave Engineering.* New York: McGraw-Hill, 1992.

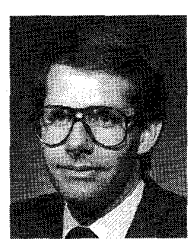
[11] M. Ree, K.-J. Chen, D. P. Kirby, N. Katzenellenbogen, and D. Grishkowsky, "Anisotropic properties of high-temperature polyimide thin films: dielectric and thermal-expansion behaviors," *J. Appl. Phys.*, vol. 72, iss. 5, pp. 2014–2021, Sept. 1, 1992.



Steven Lipa received the B.S. degree in electrical engineering from the University of Virginia, Charlottesville, VA, in 1980, and the M.S. degree in electrical engineering from North Carolina State University in 1993.

He is currently a research assistant at the Engineering Research Laboratory at North Carolina State University, where he is pursuing the Ph.D. in electrical engineering.

Mr. Lipa's current research is in the area of signal integrity and interconnect analysis of large-scale digital systems.



Michael B. Steer (S'76–M'78–SM'90) received the B.E. and Ph.D. degrees in electrical engineering from the University of Queensland, Brisbane, Australia, in 1978 and 1983, respectively.

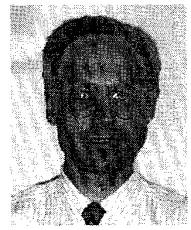
Currently he is director of the Engineering Research Laboratory (ERL), co-director of the High Frequency Electronics Laboratory, and Associate Professor of Electrical and Computer Engineering at North Carolina State University. One aspect of his research involves the experimental characterization, simulation, and computer aided design of high speed digital systems including interconnect simulation, the behavioral model development of digital drivers and receivers incorporating simultaneous switching noise, and the characterization of multichip modules. His continuing interests are in the simulation and modeling of advanced packaging, and in the computer aided analysis and design of nonlinear microwave circuits and systems, with contributions in steady-state simulation of microwave analog circuits, parameter extraction using simulated annealing techniques, microwave measurements, and simulation of millimeter-wave quasioptical power combining systems. The power combining work includes field-theoretic modeling of quasioptical systems and transient simulation of power combiners with multiple coupled oscillators. He has authored or co-authored more than 100 papers on these topics.

Dr. Steer is a 1987 Presidential Young Investigator.

Andreas C. Cangellaris (M'86) received the Diploma in electrical engineering from the Aristotelian University of Thessaloniki, Greece, in 1981, and the M.S. and Ph.D. degrees in electrical engineering from the University of California, Berkeley, in 1983 and 1985, respectively.

From 1985 to 1987, he was a Senior Research Engineer in the Electronics Department at General Motors Research Laboratories, Warren, MI. From 1987 to 1992, he was Assistant Professor for the Department of Electrical and Computer Engineering at the University of Arizona. He is currently Associate Professor in this department. He is author or co-author of over 30 papers in the areas of computational electromagnetics, microwave engineering, and modeling and simulation of high-speed interconnects. His expertise and research interests are in applied and computational electromagnetics, high-speed electronic packaging, and microwave engineering.

Dr. Cangellaris is an associate member of the International Union of Radio Science (URSI).



Paul D. Franzon (S'85–M'88) received the B.Sc. degree in physics and mathematics in 1983, the B.E. degree with First Class Honors in electrical engineering in 1984, and the Ph.D. degree in electrical engineering in 1989, all from the University of Adelaide, Adelaide, Australia.

He is currently an Associate Professor in the Department of Electrical and Computer Engineering at North Carolina State University. He has over eight years experience in electronic systems design and design methodology research and development. During that time, in addition to his current position, he has worked at AT&T Bell Laboratories in Holmdel, NJ, at the Australian Defense Science and Technology Organization, as a founding member of a successful Australian technology start up company, and as a consultant to industry, including technical advisory board positions. His current research interests include design sciences/methodology for high speed packaging and interconnect, for high speed and low power chip design and the application of Micro Electro Mechanical Machines to electronic systems. In the past, he has worked on problems and projects in wafer-scale integration, IC yield modeling, VLSI chip design and communications systems design. He has published over 50 articles and reports. He is also the co-editor and author of a book about multichip module technologies. His teaching interests focus on microelectronic systems building including package and interconnect design, circuit design, processor design and the gaining of hands-on systems experience for students.

Dr. Franzon is a member of the ACM and ISHM. He serves as the Chairman of the Education Committee for the National IEEE-CHMT Society. In 1993, he received an NSF Young Investigator's Award. In 1996, he will be the Technical Program Chair at the IEEE MultiChip Module Conference.