On the Modeling of Conductor and Substrate Losses in Multiconductor, Multidielectric Transmission Line Systems

Tawfik Rahal Arabi, *Member, IEEE*, Arthur T. Murphy, *Fellow, IEEE*,
Tapan K. Sarkar, *Senior Member, IEEE*, Roger F. Harrington, *Fellow, IEEE*,
and Antonije R. Djordjević

Abstract—**Most models used for the analysis of lossy, multiconductor, multidielectric transmission line systems are noncausal and fail to predict accurately the signal distortion on practical printed circuits. This paper reviews the method of analysis and assumptions made in these models and presents more accurate models. Finally, to illustrate the theory, numerical examples are presented.**

I. Introduction

It has been shown [1] that the TEM mode can exist in a lossy medium provided that the conductors are perfect and the medium is homogeneous. The quasi-TEM approximation, however, remains the dominant trend in the electronics industry for analyzing lossy multiconductor, multidielectric transmission line systems (MCMDTLS's). The reasons are that the quasi-TEM approximation remains valid for most practical transmission line structures and offers relative ease and low cost compared with a full-wave approach in obtaining the time-domain response of a MCMDTLS to arbitrary excitations. The equivalence between the TEM mode field equations and the transmission line equations allows the description of propagation in terms of the line circuit parameters or the medium characteristics, that is, its admittivity and impedivity [1]. Both approaches are used in this paper and interchanged whenever convenient. The frequency dependence of the circuit parameters can be directly derived from the frequency dependence of the medium characteristics. It will be shown that this dependence must be very accurately determined in order to analyze practical printed circuits. The usual way to compute the frequency-dependent circuit parameters of a MCMDTLS is to perform a 2-D analysis of the cross section of the circuit board [2]–[7]. Such analyses generally yield frequency-independent inductance and electrostatic coefficient matrices $[L]$ and $[C]$, a conductance matrix $[G]$ that varies linearly with the frequency and is associated with the dielectric loss, and a skin effect limited resistance matrix that varies with the square root of frequency and is associated with the conductor losses. In the next two sections we show that such models lead to large errors in the prediction of the electrical performance of MCMDTLS's. In Section II of this paper we solve for the time-domain response of a single, lossy, perfectly matched conductor above a ground plane and immersed in a perfect dielectric. We assume TEM propagation and study the effect of the dependence of $[R]$ and $[L]$ on the frequency. In Section III we solve for the same line except that we assume that the line is perfect while the dielectric is lossy. The TEM mode propagates in such lines and no errors result from this assumption. In Section IV, a brief generalization to MCMDTLS's is described and finally the conclusion is presented in Section V.

Manuscript received August 27, 1990; revised February 13, 1991. This work was supported by the Northeast Parallel Architectures Center at Syracuse University, by a research grant from E. I. DuPont De Nemours & Company to Syracuse University, and by a grant for supercomputer time from the Cornell National Science Foundation SuperComputing Facility.

T. Rahal Arabi is with the Experimental Station, DuPont Electronics, E. I. DuPont Nemours & Company, Wilmington, DE 19880, and with the Department of Electrical Engineering, Syracuse University, Syracuse, NY 13244.

A. T. Murphy is with the Sony Research Center, Yokohama, Japan, on leave from the Experimental Station, DuPont Electronics, E. I. DuPont De Nemours & Company, Wilmington, DE 19880.

T. K. Sarkar and R. F. Harrington are with the Department of Electrical Engineering, Syracuse University, Syracuse, NY 13244.

A. R. Djordjević is with the University of Belgrade, Belgrade, Yugoslavia.

IEEE Log Number 9100852.

II. Modeling the Conductor Losses

Theoretically speaking, the TEM propagation cannot exist on lossy conductors [1]; a component of the electric field is needed in the direction of propagation to support the current. For most practical purposes, however, the frequency-dependent losses are small and the TEM propagation is assumed to be valid. In this section we solve for the time-domain response of a single perfectly matched, lossy line above a ground plane and immersed in a homogeneous perfect dielectric. It is assumed that the resistance of the line is skin effect limited and that it varies with the square root of frequency [8]. It is thus given by $R = R_0\sqrt{\omega}$, where R_0 is determined by the geometrical and electrical data of the line, and ω is given in radians

0018-9480/91/0700-1090$01.00

per second. The transfer function of the line is thus given by

$$H(\omega) = e^{-j\omega\sqrt{LC(1 - jR_0/L\sqrt{\omega})}\,l}. \tag{1}$$

For R_0 small and particularly at high frequencies, the square root in (1) can be expanded and only the first two terms kept. The transfer function of the line can then be written as

$$H(\omega) \approx e^{-j\omega\sqrt{LC}\,l} e^{-R_0/2\sqrt{\omega C/L}\,l}. \tag{2}$$

The transfer function of the transmission line must satisfy the following constraints:

$$H(-\omega) = H^*(\omega) \tag{3}$$

$$h(t) = 0 \qquad \text{for } t < 0. \tag{4}$$

Equation (3) ensures that the impulse response of the system is real while (4) ensures that it is causal. It can be shown that (1) and (2) do not satisfy (3) or (4). The reason is to be found in the modeling of the inductance of the line. The internal inductance is inversely proportional to the square root of ω and is very small at high frequencies. If the frequency-dependent internal inductance of the line is included in the total inductance [9], then (3) reduces to

$$H(\omega) \approx e^{-j\omega\sqrt{LC}(1 + R_0/2L\sqrt{\omega})l} e^{-R_0/2\sqrt{C\omega/L}\,l} \tag{5}$$

Equation (5) satisfies both conditions (3) and (4). It differs from (2) by a nonlinear phase term. This term, resulting from the small variation of the line total inductance with the frequency, is very small and is negligible compared with the linear part of the phase ($R_0/2L\sqrt{\omega}$ versus 1); it is often neglected, particularly at high frequencies. This nonlinear phase term does not appreciably affect the time delay at the far end of the line. It is, however, the only phase term that contributes to the pulse degradation and thus significantly affects the pulse distortion at the far end of the line, particularly for very short rise times. This phase term must therefore be included. To illustrate this phenomenon, we solve for a single perfectly matched, lossy line above a ground plane and immersed in a perfect homogeneous dielectric. The geometrical and electrical data of the line are presented in Fig. 1, and the circuit configuration is shown in Fig. 2. The line frequency-dependent parameters are computed using a 2-D analysis of the system cross section [2]. These parameters are given by

$$L = 0.53\ \mu\text{H} \qquad C = 0.21\ \text{nF} \qquad R_0 = 0.6 \qquad R = R_0\sqrt{\omega}. \tag{6}$$

The time-domain response of the line to an arbitrary excitation function can be computed either by inverse Fourier transforming the product of the transfer function with the Fourier transform of the excitation function or, equivalently, by convolving the impulse response of the line with the excitation function. The first method, however, requires less computational effort and is used throughout this paper. Fig. 3 shows the impulse response

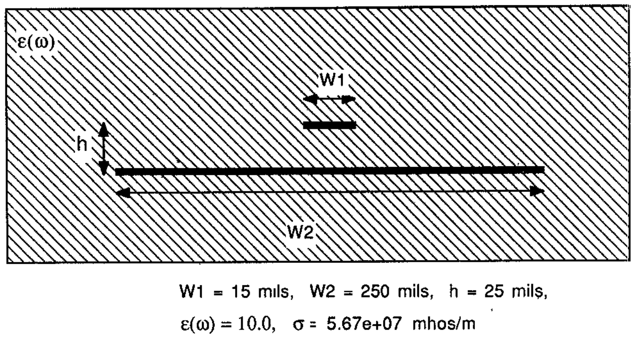


Fig. 1. The geometrical and electrical data of a single transmission line above ground plane and immersed in a homogeneous dielectric medium. The conductor's conductivity is a real constant σ and the dielectric permittivity $\epsilon(\omega)$ is a complex function of frequency.

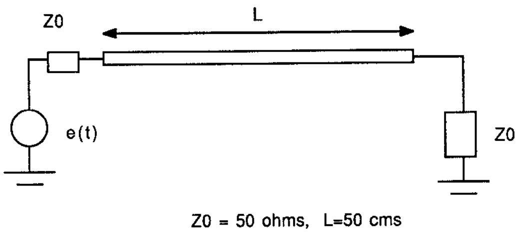


Fig. 2. The circuit configuration of the line in Fig. 1. The line is excited with an excitation $e(t)$ at the generator end and terminated by its characteristic impedance at both the generator and load ends.

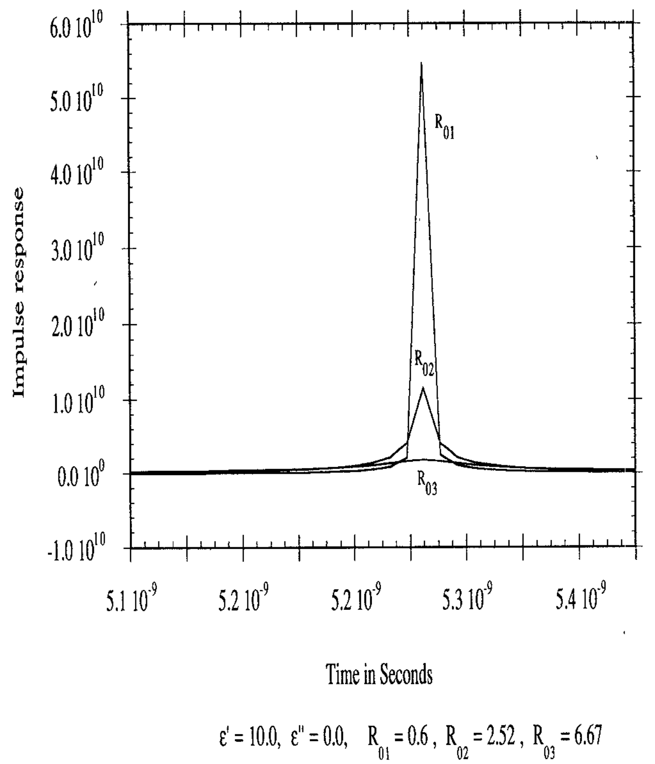


Fig. 3. The noncausal impulse response at the load end of the line in Figs. 1 and 2 when a delta excitation is applied at the generator end. The impulse response is shown for different values of the resistive loss R_{01}, R_{02}, and R_{03}.

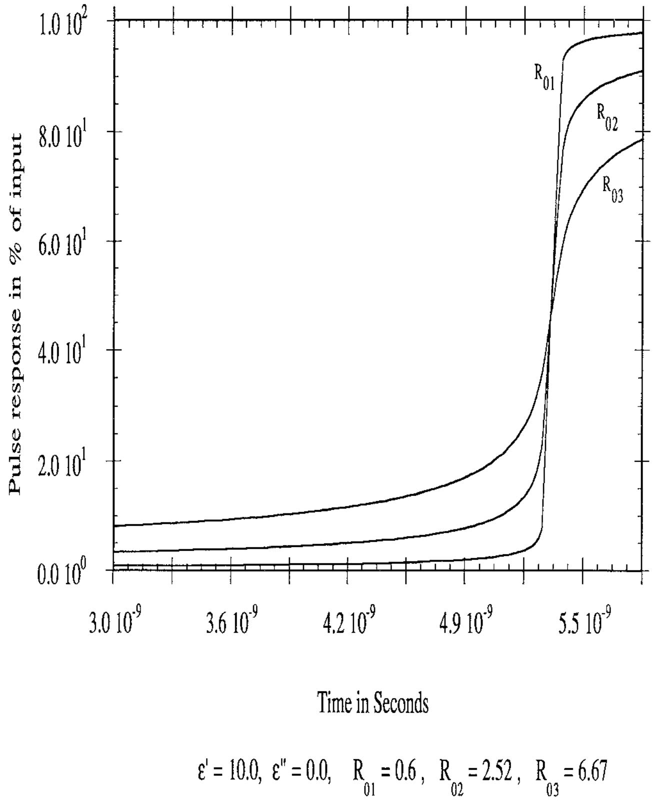


Fig. 4. The noncausal time-domain response of the line in Figs. 1 and 2 when a step of 50 ps rise time is applied at the generator end for different values of the resistive loss R_{01}, R_{02} and R_{03}. The output is expressed as a percentage of the input step.

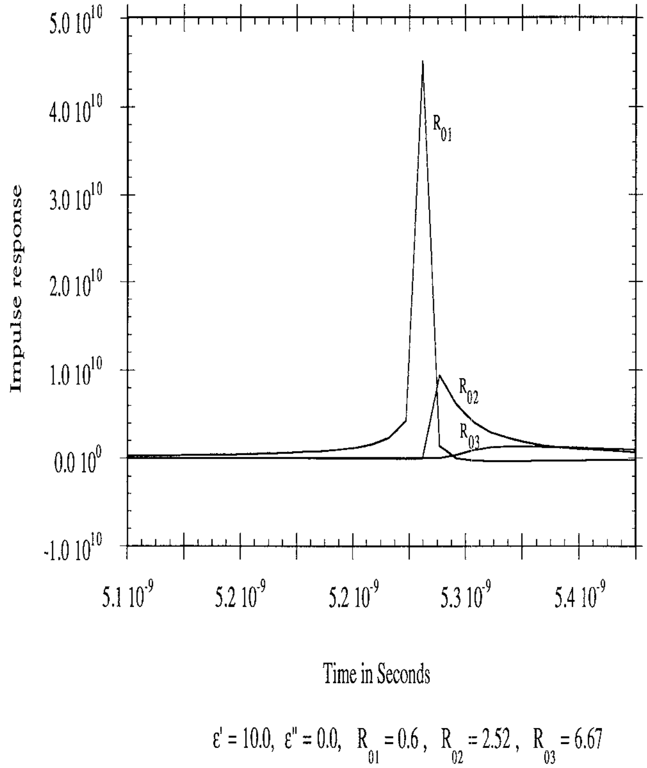


Fig. 5. The causal impulse response at the load end of the line in Figs. 1 and 2 when a delta excitation is applied at the generator end. The impulse response is shown for different values of the resistive loss R_{01}, R_{02}, and R_{03}.

of the line, computed using the noncausal transfer function of (2), at the load end when the generator end is excited with a delta excitation. The impulse response is real and noncausal. Fig. 4 shows the corresponding time-domain response to a step excitation of 50 ps rise time for different values of the resistive loss R_0. The output is plotted as a percentage of the excitation waveform. It is clear that the system is noncausal, and the velocity of propagation appears to be higher for lossier lines, a clearly nonphysical phenomenon. Fig. 5 shows the impulse response of the line using (5) this time. The impulse response is causal and is always zero prior to 5.27 ns, which is the time delay of the line if it was lossless. Finally Fig. 6 shows the corresponding pulse response of the line. It is clear that no voltage appears at the output prior to 5.27 ns, and the time delay increases slightly with an increase in the line loss. In Figs. 3, 4, 5, and 6, R_0 was artificially increased in order to show the effects of the losses on the impulse and pulse responses of the line. Table I, however, shows the output rise time and time delay for the actual line ($R_0 = 0.6$) for different values of the input rise time. All rise times are given as (10–90) rise times while the time delay is defined as the time when the signal reaches 10% of its value. It can be seen from this table that the error in the rise time degradation resulting from the use of the noncausal model of (2) is very large. It is small only when the input rise time is high enough so

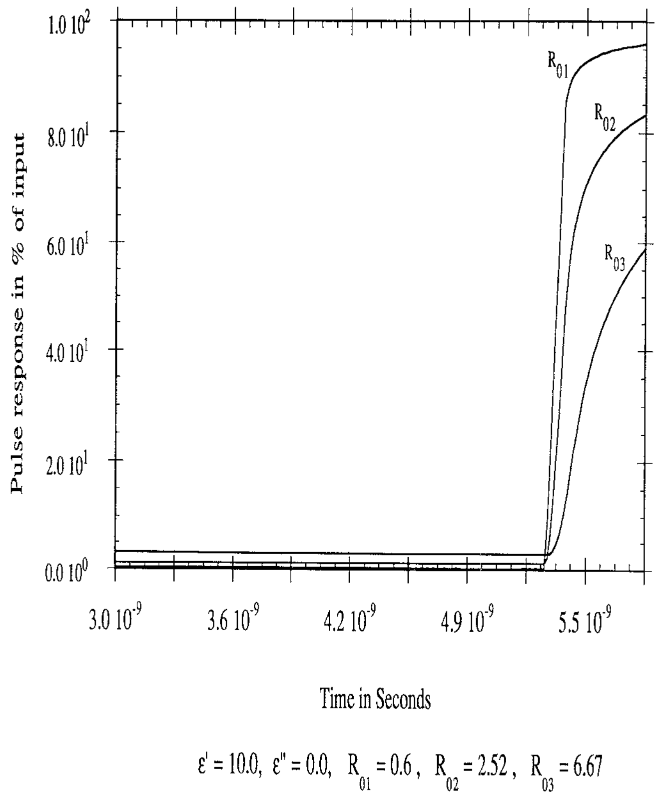


Fig. 6. The causal time-domain response of the line in Figs. 1 and 2 when a step of 50 ps rise time is applied at the generator end for different values of the resistive loss R_{01}, R_{02}, and R_{03}. The output is expressed as a percentage of the input step.

TABLE I
DIFFERENCE BETWEEN NONCAUSAL AND CAUSAL MODELS OF LINE RESISTIVE LOSSES IN PREDICTING THE LINE DELAY AND RISE TIME DEGRADATION AT THE LOAD END

Input Rise Time	RT_n	T_{0n}	RT_c	T_{0c}	Error in RT	Error in T_0
0 ps	40 ps	5.26 ns	90 ps	5.275 ns	89.5%	1.67%
45 ps	55 ps	5.275 ns	105 ps	5.285 ns	47.6%	0.2%
90 ps	95 ps	5.275 ns	130 ps	5.29 ns	26.92%	0.28%
225 ps	225 ps	5.29 ns	225 ps	5.31 ns	0%	0.37%

RT_n: rise time for the noncausal model.
T_{0n}: time delay using the noncausal model.
RT_c: rise time using the causal model.
T_{0c}: time delay using the causal model.
Error in $RT = (RT_c - RT_n)/RT_c$.
Error in $T_0 = (T_{0c} - T_{0n})/T_{0c}$.

that, for all practical purposes, the line can be considered lossless.

III. MODELING THE DIELECTRIC LOSSES

It has been shown that the TEM propagation can exist in a lossy homogeneous medium provided that the conductors are perfect [1]. The transverse electromagnetic fields have the form $(K(x,y)e^{-j\omega\sqrt{\mu(\omega)\epsilon(\omega)}})$, where $K(x,y)$ is independent of frequency. The transfer function of the line can be expressed in terms of the line circuit parameters or, equivalently, in terms of the medium characteristics [1]. For the purpose of this section it is conveniently written as

$$H(\omega) = e^{-j\omega\sqrt{\mu(\omega)\epsilon(\omega)}\,l}. \tag{7}$$

In this section it is assumed that the conductors are made of nonmagnetic matter and thus are characterized by the free-space magnetic permeability, $\mu(\omega) = \mu_0$. From (7) it follows that

$$\epsilon(\omega) = \frac{\text{pha}^2(H(\omega)) - \log^2|H(\omega)|}{\mu_0\omega^2} - 2j\frac{\log|H(\omega)|\,\text{pha}(H(\omega))}{\mu_0\omega^2} \tag{8}$$

where $\text{pha}(H(\omega))$ denotes the phase of the transfer function, and $|H(\omega)|$ its magnitude. From the above expression, it is obvious that the real and imaginary parts of $\epsilon(\omega)$ are related. If the real part, $\epsilon'(\omega)$, is a constant independent of frequency, then for the system to be causal, it follows that the imaginary part, $\epsilon''(\omega)$, must be zero or have a $(1/\omega)$ dependence. The modeling of the dielectric constant by $(\epsilon' - j\epsilon'')$ is therefore not valid. The real and imaginary parts of $\epsilon(\omega)$ are also related by the causality relations of Kramers and Kronig [10], [11]. The assumption $\epsilon(\omega) = \epsilon' - j\epsilon''$ leads to

$$H(\omega) = e^{-jk'l}e^{-k''l} \tag{9}$$

where

$$k' = \omega\sqrt{\mu\left(\frac{\epsilon' + \sqrt{\epsilon^{2\prime} + \epsilon^{2\prime\prime}}}{2}\right)}$$

$$k'' = \omega\epsilon''\sqrt{\frac{\mu_0}{2\left(\epsilon' + \sqrt{\epsilon^{2\prime} + \epsilon^{2\prime\prime}}\right)}}. \tag{10}$$

The transfer function thus has a real term and a linear phase term. The linear phase introduces a time shift while the real part characterizes the losses and distorts the propagating signal. The real part, however, describes a noncausal system. The velocity of propagation can be calculated from (10) as

$$v = \frac{\sqrt{2}}{\sqrt{\mu_0\left(\epsilon' + \sqrt{\epsilon^{2\prime} + \epsilon^{2\prime\prime}}\right)}} \tag{11}$$

The velocity of propagation does not change significantly with ϵ''. For $\epsilon'' = 0.25\epsilon'$ the time delay changes only by approximately 0.76% compared with the lossless line case ($\epsilon'' = 0$). The rise time, however, as with the lossy conductor, perfect dielectric case, changes drastically with both ϵ' and ϵ''. To illustrate this fact, we rewrite $\epsilon(\omega)$ as

$$\epsilon(\omega) = \epsilon' + \delta\epsilon'(\omega) - j\left[\epsilon'' + \Delta\epsilon''(\omega)\right] = \epsilon'(\omega) - j\epsilon''(\omega) \tag{12}$$

where ϵ' and ϵ'' are real constants and $\delta\epsilon'(\omega)$ and $\delta\epsilon''(\omega)$ are real functions of frequency. It is assumed that at all frequencies $\delta\epsilon'(\omega)$, $\epsilon''(\omega)$, and $\Delta\epsilon''(\omega)$ are all $\ll \epsilon'$. It is not assumed, however, that $\Delta\epsilon''(\omega) < \epsilon''$. The rationale for this is that while the first assumption holds for most practical materials in the frequency range of interests, $\Delta\epsilon''(\omega)$ may be of the order of magnitude of ϵ'' owing to resonances in $\epsilon''(\omega)$. Expanding the square root and keeping only the first two terms, the transfer function of the medium can be written as

$$H(\omega) \approx e^{-j\omega\sqrt{\mu_0\epsilon'}\,l}e^{-j\omega\sqrt{\mu_0}\,\delta\epsilon'(\omega)/2\sqrt{\epsilon'}\,l}e^{-\omega\sqrt{\mu_0}[\epsilon'' + \Delta\epsilon''(\omega)]/\sqrt{\epsilon'}\,l}. \tag{13}$$

The first term in the above transfer function represents a linear phase term; thus it introduces only a time shift and does not distort the pulse. The second term contributes only slightly to the velocity of propagation since it is negligible compared with the first term. It contributes, however, significantly to the pulse distortion since it is the only phase term that affects the distortion and since $\delta\epsilon'(\omega)$ may well be of the order of magnitude of

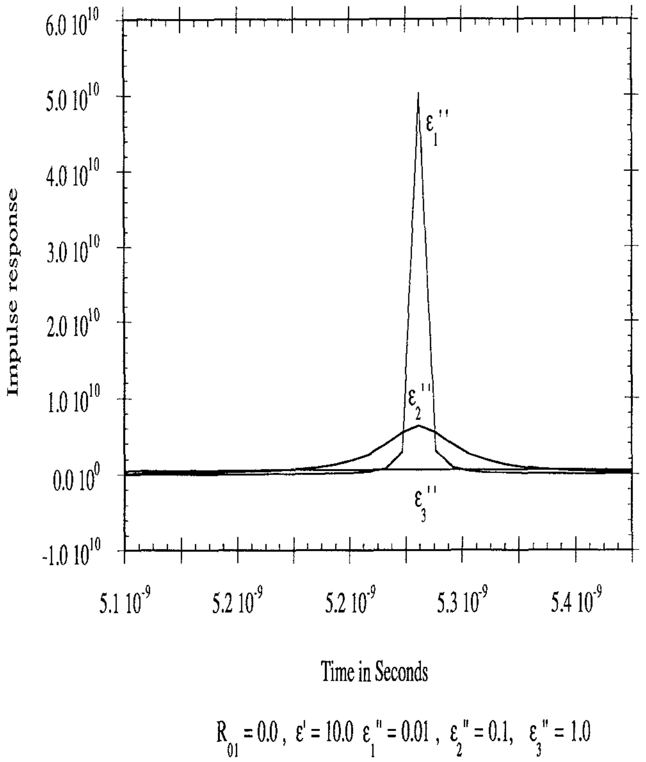


Fig. 7. The noncausal impulse response at the load end of the line in Figs. 1 and 2 when a delta excitation is applied at the generator end. The impulse response is shown for different values of the dielectric dissipation factor ϵ_1'', ϵ_2'', and ϵ_3''.

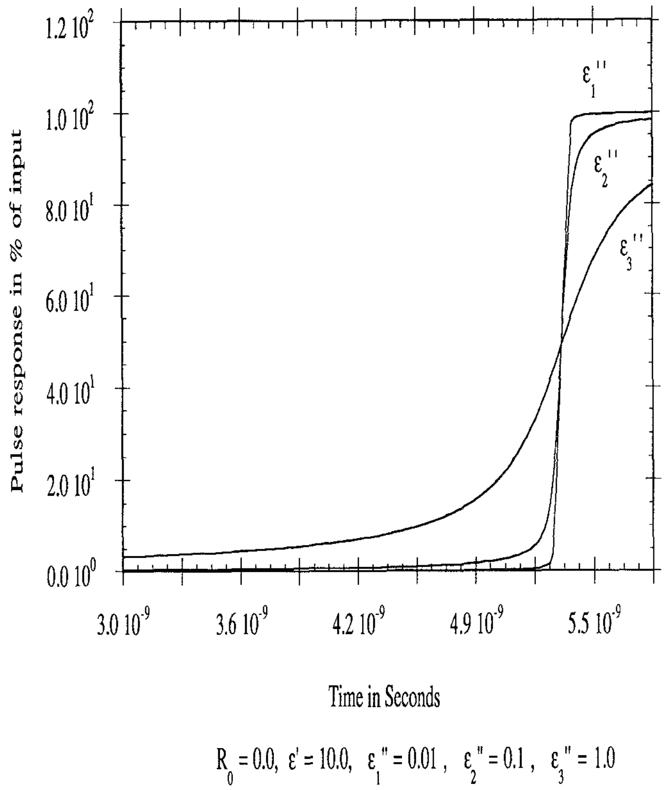


Fig. 8. The noncausal time-domain response of the line in Figs. 1 and 2 when a step of 50 ps rise time is applied at the generator end for different values of the dielectric dissipation factor ϵ_1'', ϵ_2'', and ϵ_3''. The output is expressed as a percentage of the input step.

$[\epsilon'' + \Delta\epsilon''(\omega)]$. The variation of the real part of the dielectric constant with frequency, though small, must thus be included in the modeling of lossy media since its effect may well be of the order of magnitude of the effect of $\epsilon''(\omega)$. In terms of transmission line parameters, the variation of $\epsilon'(\omega)$ with the frequency introduces a capacitance that varies with frequency. This variation is of the same order as that of $\epsilon'(\omega)$ while the variation of conductance with frequency is one degree higher than that of $\epsilon''(\omega)$ [1]. To illustrate, we solve for the same line of Fig. 1 except that the conductors are assumed perfect ($\sigma = \infty$) while the lossy dielectric is characterized by a complex dielectric permittivity of the form $\epsilon = \epsilon' - j\epsilon''$. Fig. 7 shows the impulse response of the line for various values of ϵ'' while Fig. 8 shows the corresponding time-domain response for a step input of 50 ps rise time. The time delay is approximately the same as that for the lossless line. In order to find the error in the rise time, however, the frequency-dependent dielectric permittivity must be accurately determined. This can be done only by measurements. Since no signal should appear at the far end prior to $T_0 = 5.275$ ns, the voltage prior to T_0 is due only to the error resulting from the noncausality of the medium. Thus by comparing Figs. 4 and 8 (for $t < T_0$) we can conclude that the error in Fig. 8 is of the order of magnitude of the error in Fig. 4 and thus that the error is large. To further illustrate the sensitivity of the signal distortion to small variations in $\epsilon(\omega)$ or, equivalently, $\mu(\omega)$, we compute the pulse response of the homogeneous and lossless medium

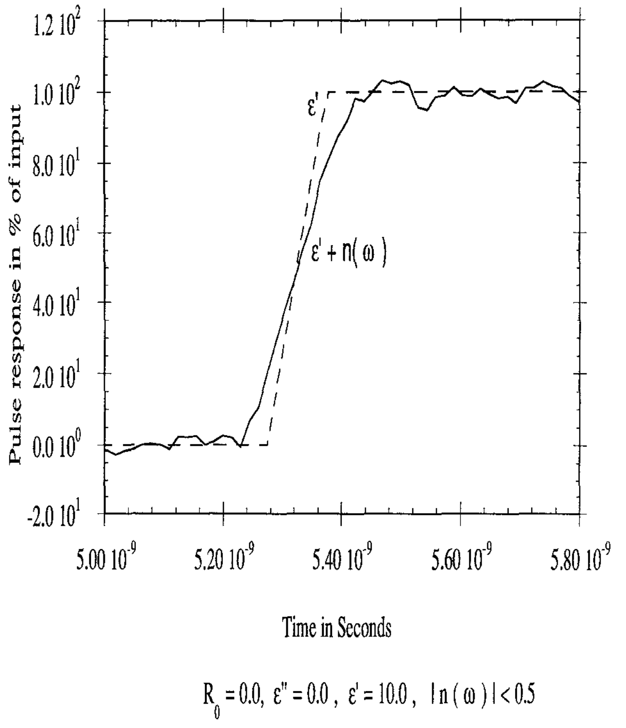


Fig. 9. The time-domain response of the line in Figs. 1 and 2 when a step of 50 ps rise time is applied at the generator end for different values of the real part of the dielectric constant $\epsilon'(\omega) = 10.0$ and $\epsilon'(\omega) = 10.0 + n(\omega)$, where $n(\omega)$ is a uniformly distributed random variable between -0.25 and 0.25.

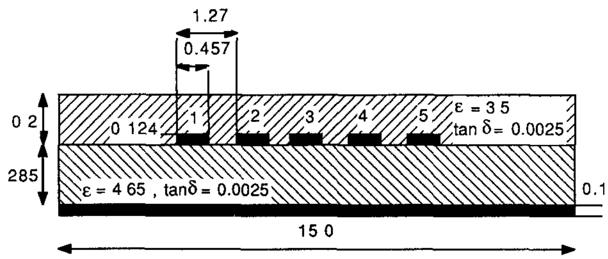


Fig. 10. The geometrical and electrical data of a multidielectric five-signal conductor line.

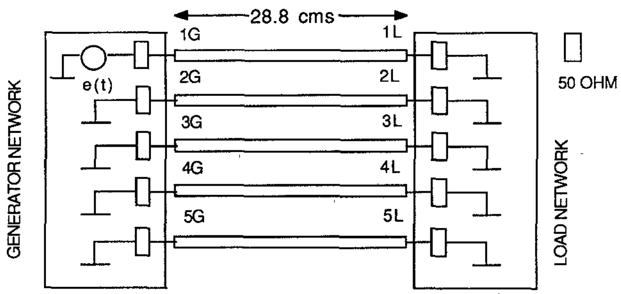


Fig. 11. The circuit configuration of the line in Fig. 10. The line is excited with an excitation $e(t)$ at the generator end of the first line and terminated by its characteristic impedance everywhere else.

given by (7) to a step of 50 ps rise time. In this example l is 50 cm and $\epsilon(\omega)$ is given by

$$\epsilon(\omega) = \epsilon' + n(\omega) \tag{14}$$

where $n(\omega)$ is a uniformly distributed random variable between $-0.025\epsilon'$ and $0.025\epsilon'$. It can be seen from Fig. 9 that a 5% variation in ϵ' causes a large error in the pulse response of the medium. The (10–90) output rise is 40 ps for the lossless case when $n(\omega) = 0$, whereas it is 150 ps when $n(\omega)$ is introduced. A small measurement error in $\epsilon(\omega)$ thus causes large errors in predicting the electrical performance of MCMDTLS's. Most measurement techniques in the literature, however, have such errors. Equation (13) provides a good way to measure ϵ', $\delta\epsilon'(\omega)$, and $\Delta\epsilon''(\omega)$ separately. Work is being done along these lines and will be reported soon.

IV. Treatment of Multiconductor, Multidielectric Transmission Lines

In an MCMDTLS, for an N-signal conductor line and one return line, there are N distinct modes that propagate on the line [12]–[14]. Each mode is decoupled from all other modes and has a different propagation constant. Figs. 12 and 13 show typical voltage waveform at the generator and load ends of the excited line of a typical five-conductor line. The line geometrical and electrical data are shown in Fig. 10 while the circuit configuration is shown in Fig. 11. For the purpose of this section, the line is treated as lossless. The voltage in this case is the sum of five independent modes. The propagation constants of the modes are determined from an eigenvalue analysis involving the line circuit parameters [12]. Each mode has

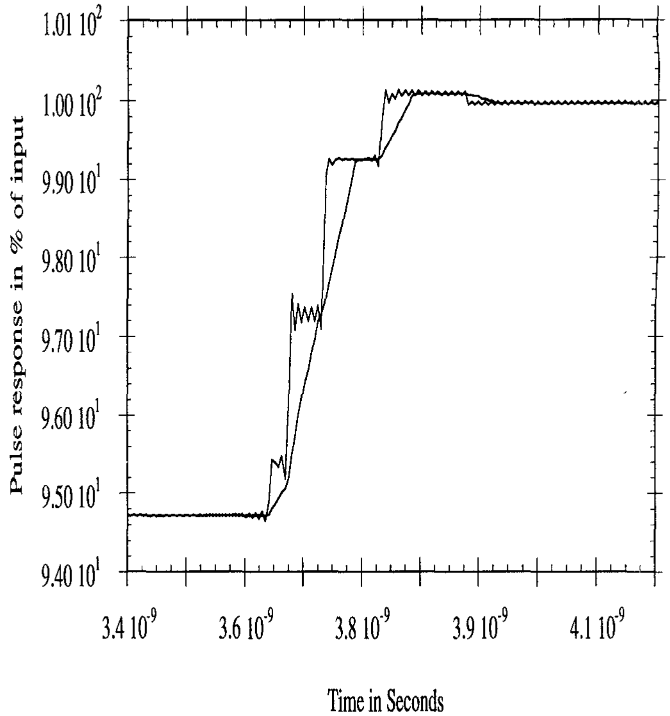


Fig. 12. The various modes excited at the generator end (1G) of the line presented in Figs. 10 and 11 when the line is excited with a step excitation of zero and 50 ps rise times at (1G). The ripples in the response to the zero rise time step are due to the FTT Gibbs phenomenon. The output is expressed as a percentage of the input signal.

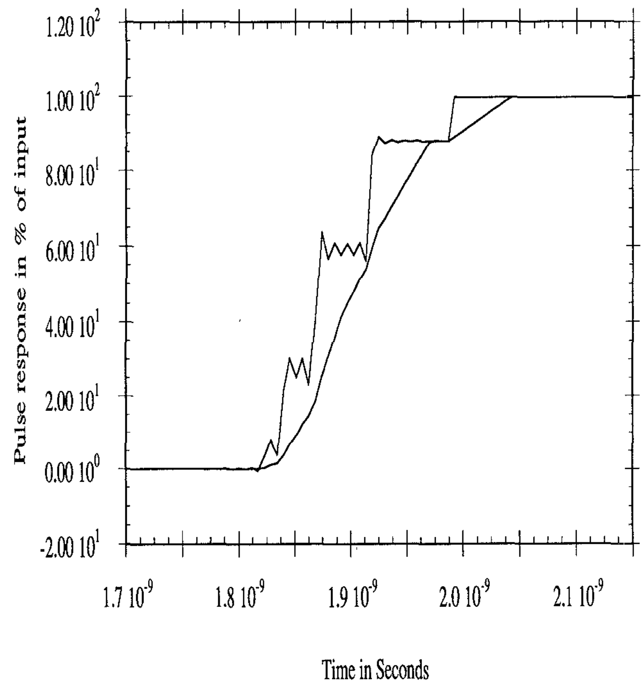


Fig. 13. The various modes excited at the load end (1L) of the line presented in Figs. 10 and 11 when the line is excited with a step excitation of zero and 50 ps rise times at (1G). The ripples in the response to the zero rise time step are due to the FTT Gibbs phenomenon. The output is expressed as a percentage of the input signal.

the same mathematical form as the single-line TEM mode and suffers the same kind of errors in the mode distortion, resulting in large errors in the crosstalk and rise time degradation. The reason, as in the single-line case, is in the modeling of the frequency-dependent inductance matrix and the frequency-dependent dielectric constant. The inductance matrix can be corrected as in the single-line case, while the variation of the dielectric constant results in capacitance and conductance matrices that vary with frequency which can be computed at each frequency in the same manner as described in [2]. Work is being done along these lines and will be promptly reported.

V. Conclusion

The traditional way of analyzing multiconductor, multidielectric transmission line networks leads to paradoxes. The models are noncausal and produce large errors in predicting the signal distortion at the excited lines and crosstalk at the quiet lines of the MCMDTLS. The reason is that a small nonlinear phase term is neglected in the transfer function of the MCMDTLS. A more accurate model, causal to within engineering accuracy, has been presented for the conductor losses and compared with the old ones and a newly developed model for the frequency-dependent dielectric constant has been presented. Experimental work is required, however, to check the accuracy of the new dielectric model.

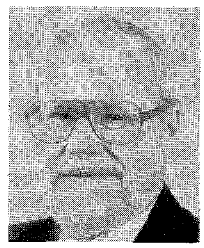
Acknowledgment

One of the authors would like to thank Dr. Fred Tesche for his valuable suggestions.

References

[1] R. F. Harrington, *Time-Harmonic Electromagnetic Fields*. New York: McGraw-Hill, 1961.
[2] T. R. Arabi, T. K. Sarkar, and A. R. Djordjevic, "Time and frequency domain characterization of multiconductor transmission lines," *J. Electromagnetics*, vol. 9, pp. 85–112, 1989.
[3] C. Wei, R. F. Harrington, J. R. Mautz, and T. K. Sarkar, "Multiconductor transmission lines in multilayered dielectric media," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 439–450, Apr. 1984.
[4] R. F. Harrington and C. Wei, "Losses on multiconductor transmission lines in multilayered dielectric media," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 705–710, July 1984.
[5] S. M. Rao, T. K. Sarkar, and R. F. Harrington, "The electrostatic field of conducting bodies in multiple dielectric media," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-32, pp. 1441–1448, 1984.
[6] J. Venkataraman, S. Rao, A. R. Djordjevic, T. K. Sarkar, and Y. Naiheng, "Analysis of arbitrarily oriented microstrip transmission lines in arbitrarily shaped dielectric media over a finite ground plane," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, Oct. 1985.
[7] A. R. Djordjevic, T. K. Sarkar, and S. M. Rao, "Analysis of finite conductivity cylindrical conductors excited by axially-independent TM electromagnetic field," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 960–966, Oct. 1985.
[8] A. R. Djordjevic, T. K. Sarkar, and R. F. Harrington, "Analysis of lossy transmission lines with arbitrary nonlinear networks," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 660–666, June 1986.
[9] R. E. Matick, *Transmission Lines For Digital and Communication Networks*. New York: McGraw-Hill, 1969.
[10] F. E. Gardiol, *Lossy Transmission Lines*. Norwood, MA: Artech House, 1987.
[11] F. E. Gardiol, "On the thermodynamic paradox in ferrite loaded waveguides," *Proc. IEEE*, vol. 55, pp. 1616–1617, 1976.
[12] A. R. Djordjevic and T. K. Sarkar, "Analysis of time response of lossy multiconductor transmission line networks," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-35, pp. 898–908, Oct. 1987.
[13] F.-Y. Chang, "Transient of lossless coupled transmission lines in a nonhomogeneous dielectric medium," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 616–626, Sept. 1970.
[14] K. D. Marx, "Propagation modes, equivalent circuits, and characteristic terminations for multiconductor transmission lines with inhomogeneous dielectrics," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-21, pp. 450–457, July 1973.

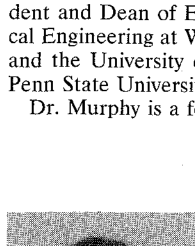
Tawfik Rahal Arabi (M'90) was born in Beirut, Lebanon, on December 3, 1964. He received the B.E.E. degree from the American University of Beirut, Beirut, Lebanon, in June 1985 and the M.S. degree from Syracuse University, Syracuse, NY, in August 1987. Currently he is working toward the Ph.D. degree in the Department of Electrical Engineering at Syracuse University. He is also a member of the electronic systems research group at Du Pont Electronics, Wilmington, DE, working in the area of high-speed digital signals and interconnects. His research focuses on numerical problems in electromagnetics, especially those in the electronic packaging area.



Arthur T. Murphy (S'49–A'54–M'54–SM'61–F'90) received the Ph.D. and M.S. degrees in electrical engineering from Carnegie Mellon University and the B.E.E. degree from Syracuse University.

He established an Electronic Systems Research Group which guides R&D programs by evaluating end-use performance of Du Pont Electronics products. He has expertise in applications in high speed electronic interconnections and electromagnetic interference. He has established a state-of-the art laboratory and a unique computer-aided design system, called ICONSIMsm, for interconnection simulation. This has been used for designing advanced interconnection assemblies, rigid and flexible printed wiring boards, and ceramic and polymeric packages. He also recently developed a unique thick-film UHF filter which eliminates electromagnetic emissions from computers. He is currently an exchange visiting scientist at the Sony Research Center, Yokohama, Japan, where he is developing high-speed/high-frequency gallium arsenide integrated circuit packages. Before joining Du Pont, 12 years ago, he was Brown Professor and Head of Mechanical Engineering at Carnegie Mellon University and previously held positions of Vice President and Dean of Engineering at Widener University, Head of Electrical Engineering at Wichita State University, Visiting Professor at M.I.T. and the University of Manchester (England), and Adjunct Lecturer at Penn State University.

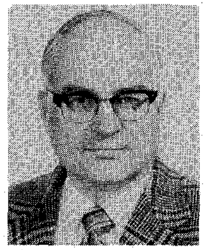
Dr. Murphy is a fellow of the AAAS.



Tapan K. Sarkar (S'69–M'76–SM'81) was born in Calcutta, India, on August 2, 1948. He received the B. Tech. degree from the Indian Institute of Technology, Kharagpur, India, in 1969, the M.Sc.E. degree from the University of New Brunswick, Fredericton, Canada, in 1971, and the M.S. and Ph.D. degrees from Syracuse University, Syracuse, NY, in 1975.

From 1975 to 1976 he was with the TACO Division of General Instruments Corporation. He was with the Rochester Institute of Technology, Rochester, NY, from 1976 to 1985. He was a Research Fellow at the Gordon McKay Laboratory, Harvard University, Cambridge, MA, from 1977 to 1978. He is now a Professor in the Department of Electrical and Computer Engineering at Syracuse University. His current research interests deal with the numerical solution of operator equations arising in electromagnetics and signal processing with application to system design. He has authored or coauthored over 154 journal articles and conference papers and has written chapters in eight books.

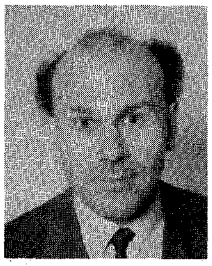
Dr. Sarkar is a registered professional engineer in the state of New York. He received the Best Paper Award of the IEEE TRANSACTIONS ON ELECTROMAGNETIC COMPATIBILITY in 1979. He also received one of the "best solution" awards in May 1977 at the Rome Air Development Center (RADC) Spectral Estimation Workshop. He was an Associate Editor for feature articles of the *IEEE Antennas and Propagation Society Newsletter* and the IEEE TRANSACTIONS ON ELECTROMAGNETIC COMPATIBILITY. He was the Technical Program Chairman for the 1988 IEEE Antennas and Propagation Society International Symposium and URSI Radio Science Meeting. Dr. Sarkar is an Associate Editor for the *Journal of Electromagnetic Waves and Applications* and is on the editorial board of the *International Journal of Microwave and Millimeter-Wave Computer Aided Engineering*. He has been appointed U.S. Research Council Representative to many URSI General Assemblies. He is also Chairman of the Intercommission Working Group of International URSI on Time Domain Metrology. He is a member of Sigma Xi and the International Union of Radio Science Commissions A and B.



Roger F. Harrington (S'84–A'53–M'57–SM'62–F'68) was born in Buffalo, NY, on December 24, 1925. He received the B.E.E. and M.E.E. degrees from Syracuse University, Syracuse, NY, in 1948 and 1950, respectively, and the Ph.D. degree from Ohio State University, Columbus, in 1952.

From 1945 to 1946, he served as an Instructor at the U.S. Naval Radio Materiel School, Dearborn, MI, and from 1948 to 1950 he was employed as an Instructor and Research Assistant at Syracuse University. While studying at Ohio State University, he served as a Research Fellow in the Antenna Laboratory. Since 1952, he has been on the faculty of Syracuse University, where he is presently Professor of Electrical Engineering. During the years 1959–1960, he was visiting Associate Professor at the University of Illinois, Urbana; in 1964 he was Visiting Professor at the University of California, Berkeley; and in 1969 he was Guest Professor at the Technical University of Denmark, Lyngby, Denmark.

Dr. Harrington is a member of Tau Beta Pi, Sigma Xi, and the American Association of University Professors.



Antonije R. Djordjević was born in Belgrade, Yugoslavia, in 1952. He received the B.Sc., M.Sc., and D.Sc. degrees from the University of Belgrade in 1975, 1977, and 1979, respectively.

In 1975 he joined the Department of Electrical Engineering, University of Belgrade, where at present he is an Associate Professor of Microwaves. From February 1983 until February 1984 he was with the Department of Electrical Engineering, Rochester Institute of Technology, Rochester, NY, as a Visiting Associate Professor. His research focuses on numerical problems in electromagnetics, especially those applied to antennas and microwave passive components. He is the coauthor of a monograph on wire antennas and of two textbooks.