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(54) Title: METHODS FOR TRANSMITTING A WAVEFORM HAVING A CONTROLLABLE ATTENUATION AND PROPAGATION VELOCITY

(57) Abstract: Systems and methods are described for transmitting a waveform having a controllable attenuation and propagation velocity. An exemplary method comprises: generating an exponential waveform, the exponential waveform (a) being characterized by the equation $V_{in} = De^{-ASD[t-vSD]}$, where D is a magnitude, V_{in} is a voltage, t is time, ASD is an attenuation coefficient, and v_{SD} is a propagation velocity; and (b) being truncated at a maximum value. An exemplary apparatus comprises: an exponential waveform generator; an input recorder coupled to an output of the exponential waveform generator; a transmission line under test coupled to the output of the exponential waveform generator; an output recorder coupled to the transmission line under test; an additional transmission line coupled to the transmission line under test; and a termination impedance coupled to the additional transmission line and to a ground.



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DESCRIPTION

METHODS FOR TRANSMITTING A WAVEFORM HAVING A CONTROLLABLE ATTENUATION AND PROPAGATION VELOCITY

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CROSS-REFERENCE(S) TO RELATED APPLICATION(S)

This application is a continuation-in-part of, and claims a benefit of priority under 35 U.S.C. 119(e) and/or 35 U.S.C. 120 from, copending U.S. Ser. No. 09/519,922, filed March 7, 2000, the entire contents of which are hereby expressly incorporated by reference for all
10 purposes.

BACKGROUND OF THE INVENTION

1. Field of the Invention

The invention relates generally to the field of signal transmission. More particularly, a
15 representative embodiment of the invention relates to transmission line testing and communication.

2. Discussion of the Related Art

Transmission lines, with their characteristic loss of signal as well as inherent time delay, may create problems in designing systems that employ a plurality of signals, which may be
20 subject to delay and distortion. Typical signals used to generate inputs to transmission lines generally exhibit delay or propagation times that are not easily determinable. The propagation velocity of these waves is also variable with displacement along the transmission line.

A Time Domain Reflectometer (TDR) is a test instrument used to find faults in transmission lines and to empirically estimate transmission line lengths and other parameters
25 characterizing the line, such as: inductance per unit length, capacitance per unit length, resistance per unit length and conductance per unit length.

An important measurement in TDR test technology is the time-of-flight (TOF) of a test pulse generated by the instrument and applied to the line. The time-of-flight may be measured by timing the passage of the pulse detected at two locations along the line. Along with a value of the

propagation speed of the pulse, time-of-flight measurements can allow one to obtain the distance between measurement points or, in the case of a reflected wave, the distance from the pulse launch point to the location of the impedance change causing the pulse to be reflected and returned.

5 A fundamental limitation in TDR technology is the poor accuracy of TOF measurements in lossy, dispersive transmission lines. The relatively high TDR accuracy of TOF values obtainable in short low loss, low dispersion transmissions lines is possible only because the propagating test pulses keep their shape and amplitude in tact over the distances they travel during TOF measurements. By contrast, in dispersive, lossy long transmission lines the test
10 pulses used in the art change shape, amplitude, and speed as they travel.

Further, it is difficult to provide high-speed communications in a lossy, frequency dependent transmission media. It would be advantageous to have a method to increase data transmission rates in such transmission lines.

Until now, the requirements of providing a method and/or apparatus for accurately
15 measuring times-of-flight, estimating line lengths and other parameters characterizing lossy, dispersive transmission lines, and providing high-speed communications via such transmission media have not been met. What is needed is a solution that addresses these requirements.

SUMMARY OF THE INVENTION

20 There is a need for the following embodiments. Of course, the invention is not limited to these embodiments.

According to an aspect of the invention, a method for transmitting a waveform having an essentially constant propagation velocity along a transmission line comprises: generating an exponential waveform, the exponential waveform (a) being characterized by the equation
25 $V_{in} = De^{-A_{SD}[x-v_{SD}t]}$, where D is a magnitude, V_{in} is a voltage, t is time, A_{SD} is an attenuation coefficient, and v_{SD} is a propagation velocity; and (b) being truncated at a maximum value. According to another aspect of the invention, an apparatus for determining an attenuation coefficient and a propagation velocity comprises: an exponential waveform generator; an input recorder coupled to an output of the exponential waveform generator; a transmission line under
30 test coupled to the output of the exponential waveform generator; an output recorder coupled to

the transmission line under test; an additional transmission line coupled to the transmission line under test; and a termination impedance coupled to the additional transmission line and to a ground.

These, and other, embodiments of the invention will be better appreciated and understood when considered in conjunction with the following description and the accompanying drawings. It should be understood, however, that the following description, while indicating various embodiments of the invention and numerous specific details thereof, is given by way of illustration and not of limitation. Many substitutions, modifications, additions and/or rearrangements may be made within the scope of the invention without departing from the spirit thereof, and the invention includes all such substitutions, modifications, additions and/or rearrangements.

BRIEF DESCRIPTION OF THE DRAWINGS

The drawings accompanying and forming part of this specification are included to depict certain aspects of the invention. A clearer conception of the invention, and of the components and operation of systems provided with the invention, will become more readily apparent by referring to the exemplary, and therefore nonlimiting, embodiments illustrated in the drawings, wherein like reference numerals (if they occur in more than one view) designate the same or similar elements. The invention may be better understood by reference to one or more of these drawings in combination with the description presented herein. It should be noted that the features illustrated in the drawings are not necessarily drawn to scale.

FIG. 1A is a graph of an SD waveform for $(x = 0, t)$, representing an embodiment of the invention.

FIG. 1B is a graph of an SD waveform in moving frame at $x' = \ell$ ($x = \ell, t'$), representing an embodiment of the invention.

FIG. 2 is block diagram of a setup to measure SD parameters, representing an embodiment of the invention.

FIG. 3 is a graph of voltage traces at $x = 0$ and $x = 1002$ ft of a 2004 ft transmission line (T1 cable) with $\alpha = 3.0 \times 10^6$ 1/sec indicating time-of-flight measured at 4V threshold, representing an embodiment of the invention.

FIG. 4A is a graph of SD threshold overlap regions for voltages at $x = 0$ and at $x = d$ (1002 ft), representing an embodiment of the invention.

FIG. 4B is a graph of measured constant threshold times-of-flight for 1002 ft T1 line with $\alpha = 3.0 \times 10^6$ 1/sec, representing an embodiment of the invention.

5 **FIG. 5A** is a graph of SD threshold overlap regions for voltages at $x = 0$ and 1/2 voltage across $R_t = \infty$, representing an embodiment of the invention.

FIG. 5B is a graph of measured constant threshold times-of-flight for 2004 ft T1 line with $\alpha = 3.0 \times 10^6$ 1/sec, representing an embodiment of the invention.

10 **FIG. 6A** is a graph of an input pulse applied to a T1 cable to empirically determine a transfer function, representing an embodiment of the invention.

FIG. 6B is a graph of a pulse measured at 1002 for a T1 cable, representing an embodiment of the invention.

FIG. 7A is a graph of a Power Spectral Density of an input pulse applied to a T1 cable, representing an embodiment of the invention.

15 **FIG. 7B** is a graph of a Power Spectral Density of a pulse measured at 1002 for a T1 cable, representing an embodiment of the invention.

FIG. 8 is a graph of an empirically derived transfer function for a 1002 ft T1 cable, representing an embodiment of the invention.

20 **FIG. 9** is a graph of a measured initial waveform with simulated 1002 ft waveform for T1 cable and constant threshold times-of-flight, representing an embodiment of the invention.

FIG. 10 is a graph of a difference between measured and simulated 1002 ft voltage traces, representing an embodiment of the invention.

25 **FIG. 11** is a graph of a simulated initial waveform with 1002 ft waveform for T1 cable, $\alpha = 1 \times 10^5$ sec⁻¹, and constant threshold times-of-flight, representing an embodiment of the invention.

FIG. 12 is a graph of a simulated initial waveform with 1002 ft waveform for T1 cable, $\alpha = 1 \times 10^7$ sec⁻¹, and constant threshold times-of-flight, representing an embodiment of the invention.

30 **FIG. 13A** is a graph of SD parameters V_{SD} determined by simulation with empirical transfer function, representing an embodiment of the invention.

FIG. 13B is a graph of SD parameters A_{SD} determined by simulation with empirical transfer function, representing an embodiment of the invention.

FIG. 14 is block diagram of a setup to determine $Z_{SD}(\alpha)$, representing an embodiment of the invention.

5 **FIG. 15A** is a graph of voltage traces measured at $d = 1002$ ft with varying termination resistances, $\alpha = 1.5 \times 10^6$ 1/sec, representing an embodiment of the invention.

FIG. 15B is a graph of ratios of output signal amplitudes with known termination traces and one-half of line output signal amplitude for open termination trace, $\alpha = 1.5 \times 10^6$ 1/sec, representing an embodiment of the invention.

10 **FIG. 16** is a graph of experimentally determined SD impedances for a T1 cable as a function of α , representing an embodiment of the invention.

FIG. 17 is a graph of SD voltage traces and the natural logarithm of SD voltage traces measured at input, ~6kft, ~12kft and ~18kft along 24 AWG twisted wire pairs connected in series, representing an embodiment of the invention.

15 **FIG. 18** is a graph used to detect end of SD region demonstrating sensitivity to noise and speed of response for the filters at 12-kft., representing an embodiment of the invention.

FIG. 19 is a graph of a variation of $\frac{1}{\sqrt{LC}}$ as a fraction of the speed of light in a vacuum with α , representing an embodiment of the invention.

20 **FIG. 20** is a graph of an input waveform attenuated and time shifted for 11,573 ft (Top), 12,219 ft (Middle) and 12,859 ft (Bottom) shown with waveform measured at ~12 kft, representing an embodiment of the invention.

FIG. 21 is a graph of a standard deviation of $V_{\text{Predicted}} - V_{\text{Measured}}$ in the SD region vs. estimated distance in region of correlation, representing an embodiment of the invention.

25 **FIG. 22A** is a graph of $V_{\text{Predicted}} - V_{\text{Measured}}$ and minimum variance of $V_{\text{Predicted}} - V_{\text{Measured}}$ at estimated distance of 12,216 ft, representing an embodiment of the invention.

FIG. 22B is a graph of $V_{\text{Predicted}} - V_{\text{Measured}}$ and variance of $V_{\text{Predicted}} - V_{\text{Measured}}$ at estimated distance of 12,254 ft, representing an embodiment of the invention.

30 **FIG. 23** is a graph of voltage traces measured a ~6, ~12, and ~18 kft of 24 AWG BKMA 50 cable and applied waveform attenuated and shifted in time according to SD prediction, representing an embodiment of the invention.

FIG. 24 is a graph of voltage trace measured at input showing applied truncated SD pulse and wave reflected from open termination at 1002 ft, representing an embodiment of the invention.

FIG. 25 is a graph of voltage trace measured at input showing wave reflected from open termination at 1002 ft and applied wave attenuated and shifted forward in time for best fit resulting in 2004 ft estimated total distance traveled, representing an embodiment of the invention.

FIG. 26A is a graph of SD TDR threshold overlap regions for voltages at $x = 0$ for 1002 ft open terminated T1 line, representing an embodiment of the invention.

FIG. 26B is a graph of measured constant threshold TDR times-of-flight for 1002 ft T1 line with $\alpha = 6.7 \times 10^6$ 1/sec resulting in total estimated distance traveled by the reflected wave of 2010 ft., representing an embodiment of the invention.

FIG. 27 is a graph of voltage trace measured at input showing applied truncated SD pulse and wave reflected from short circuit termination at 1002 ft., representing an embodiment of the invention.

FIG. 28 is a graph of a voltage trace measured at input inverted, showing wave reflected from short circuit termination at 1002 ft inverted and the non-inverted applied input wave attenuated and shifted forward in time for best fit resulting in 2004 ft estimated total distance traveled, representing an embodiment of the invention.

FIG. 29A is a graph of SD TDR threshold overlap regions for voltages at $x = 0$ for 1002 ft short circuit terminated T1 line, representing an embodiment of the invention.

FIG. 29B is a graph of measured constant threshold TDR times-of-flight for 1002 ft T1 line with $\alpha = 6.7 \times 10^6$ 1/sec resulting in total estimated distance traveled of 2008 ft., representing an embodiment of the invention.

FIG. 30 is a graph of a voltage pulse applied to 1002 ft T1 cable (dotted line), and to resistor with value Z_o , representing an embodiment of the invention.

FIG. 31 is another graph of the voltage pulse applied to 1002 ft T1 cable (dotted line), and to resistor with value Z_o , representing an embodiment of the invention.

FIG. 32 is a graph of a transfer function between the difference of waveform applied to the twisted wire pair and the waveform applied to Z_o , representing an embodiment of the invention.

FIG. 33 is a graph of simulated applied SD waveform with $\alpha = 6.7 \times 10^6$ 1/sec, and reflected waveform calculated with transfer function, representing an embodiment of the invention.

FIG. 34 is a graph of simulated voltage traces showing wave reflected from open termination at 1002 ft and applied wave attenuated and shifted forward in time for best fit resulting in 2004 ft estimated total distance traveled, representing an embodiment of the invention.

FIG. 35A is a graph of simulated SD TDR threshold overlap regions for voltages at $x = 0$ for 1002 ft open terminated T1 line, representing an embodiment of the invention.

FIG. 35B is a graph of simulated constant threshold TDR times-of-flight for 1002 ft T1 line with $\alpha = 6.7 \times 10^6$ 1/sec resulting in total estimated distance traveled of 2005 ft, representing an embodiment of the invention.

FIG. 36A is a graph of an uncompensated SD waveform applied to a 2004-ft twisted wire pair transmission line, representing an embodiment of the invention.

FIG. 36B is a graph of an uncompensated SD waveform measured 100-Ohm termination on a 2004-ft twisted wire pair transmission line, representing an embodiment of the invention.

FIG. 37A is a graph of a compensated SD waveform applied to a 2004-ft twisted wire pair transmission line, representing an embodiment of the invention.

FIG. 37B is a graph of a compensated SD waveform measured 100-Ohm termination on a 2004-ft twisted wire pair transmission line, representing an embodiment of the invention.

FIG. 38 is a graph of a series of four symbols measured at input to a 2004-ft twisted wire pair transmission line, representing an embodiment of the invention.

FIG. 39 is a graph of a series of four symbols measured at 100-Ohm termination at 2004-ft on twisted wire pair transmission line, representing an embodiment of the invention.

FIG. 40 is a natural logarithmic graph of voltages measured at 100-ohm termination at 2004-ft for each symbol period showing a least mean square fit of linear region, representing an embodiment of the invention.

FIG. 41 is a graph of $X_i = e^{(\alpha_0 + (i-1) \cdot \Delta\alpha) \cdot t}$, representing an embodiment of the invention.

FIG. 42 is a graph of an orthonormal vector set $Y = [Y_1, Y_2, Y_3, Y_4, Y_5]$, representing an embodiment of the invention.

FIG. 43 is a graph of non-orthogonal waveforms $Y' = [Y_1, Y_2, Y_3, Y_4, Y_5] \cdot \sin\left(\frac{\pi}{7 \times 10^{-6}} \cdot t\right)$, representing an embodiment of the invention.

5 **FIG. 44** is a graph of dispersed waveforms Y' at the receiver at 8000 ft, representing an embodiment of the invention.

FIG. 45 is a graph of a compensating Pulse C , at Input and at 8000 ft, representing an embodiment of the invention.

10 **FIG. 46** is a graph of component functions at input with compensating pulses, $Z = Y' + b \cdot C$, representing an embodiment of the invention.

FIG. 47 is a graph of component functions Z at 8 kft, representing an embodiment of the invention.

FIG. 48 is a graph of linear independent waveforms, S , transmitted to generate orthogonal waveforms at 8 kft, representing an embodiment of the invention.

15 **FIG. 49** is a graph of orthogonal waveforms, S , received at 8 kft, representing an embodiment of the invention.

FIG. 50 is a flow diagram of an orthonormal set, S , generation algorithm, representing an embodiment of the invention.

20 **FIG. 51** is a graph of three successive transmitted symbols, Q , with five bits encoded on each of four orthonormal pulses of S , representing an embodiment of the invention.

FIG. 52 is a graph of three successive symbols, Q , at the input to the receiver at 8kft with five bits encoded on each of four orthonormal pulses of S , representing an embodiment of the invention.

25 **FIG. 53** is a graph of histogram of signal error, $a_{\text{expected}} - a_{\text{detected}}$, for one second, or 1.67×10^6 bits, of data transmitted with -140 dBm AWGN present, representing an embodiment of the invention.

DESCRIPTION OF PREFERRED EMBODIMENTS

The invention and the various features and advantageous details thereof are explained more fully with reference to the nonlimiting embodiments that are illustrated in the accompanying drawings and detailed in the following description. Descriptions of well known starting materials, processing techniques, components and equipment are omitted so as not to unnecessarily obscure the invention in detail. It should be understood, however, that the detailed description and the specific examples, while indicating specific embodiments of the invention, are given by way of illustration only and not by way of limitation. Various substitutions, modifications, additions and/or rearrangements within the spirit and/or scope of the underlying inventive concept will become apparent to those skilled in the art from this disclosure.

Within this application several publications are referenced by Arabic numerals within parentheses or brackets. Full citations for these, and other, publications may be found at the end of the specification immediately preceding the claims after the section heading References. The disclosures of all these publications in their entireties are hereby expressly incorporated by reference herein for the purpose of indicating the background of the invention and illustrating the state of the art.

The below-referenced allowed U.S. Patent Application discloses embodiments that were satisfactory for the purposes for which they are intended. The entire contents of U.S. Patent Application 09/519,922, filed March 7th, 2000, is hereby expressly incorporated by reference herein for all purposes.

In general, the context of the invention can include signal transmission. The context of the invention can include a method and/or apparatus for measuring and estimating transmission line parameters. The context of the invention can also include a method and/or apparatus for performing high-speed communication over a lossy, dispersive transmission line.

A practical application of the invention that has value within the technological arts is in time domain reflectometry. Further, the invention is useful in conjunction with lossy transmission lines. The invention is also useful in conjunction with frequency dependent transmission lines. Another practical application of the invention that has value within the

technological arts is high-speed communications. There are virtually innumerable other uses for the invention. In fact, any application involving the transmission of a signal may benefit.

A method for transmitting a waveform having a controllable attenuation and propagation velocity, representing an embodiment of the invention, can be cost effective and advantageous.

5 The invention improves quality and/or reduces costs compared to previous approaches.

The invention includes a method and/or apparatus for utilizing a truncated waveform, coined a Speedy Delivery (SD) waveform by the inventors. An analysis of SD propagation in coupled lossy transmission lines is presented and practical considerations associated with truncating the SD waveforms are addressed. Parameters used to describe the propagation of the
10 SD waveform are defined and techniques for determining their values are presented. These parameters include the Speedy Delivery propagation velocity, v_{SD} , the Speedy Delivery attenuation coefficient, A_{SD} , and the Speedy Delivery impedance, Z_{SD} . These parameters may depend on properties of a transmission media as well as an exponential coefficient α (SD input signal parameter).

15 The invention describes SD signal propagation in coupled transmission lines. An embodiment involving two transmission lines is illustrated. More complex configurations that include a larger number of coupled lines may be readily developed by one of ordinary skill in the art in light of this disclosure. Additionally, the behavior of the propagation of a slowly varying envelope of an optical pulse containing a truncated SD signal in a single mode communication
20 fiber is described.

The invention teaches a method for using SD test pulses to obtain empirical estimates of the SD signal propagation parameters, A_{SD} and v_{SD} , in lossy transmission lines, including those with frequency dependent parameters. Once the value of v_{SD} is determined for a calibration cable length, constant threshold time of flight measurements can be used to measure distances to
25 discontinuities in similar cables. The calibration procedure can be repeated experimentally for each exponential coefficient α .

The invention provides a method for utilizing a wideband, non-SD waveform in order to develop an empirical transfer function for a calibration length of a given transmission media. This transfer function may be used to simulate the transmission of SD waveforms with a wide
30 range of values for the exponential coefficient α . The SD parameters $A_{SD}(\alpha)$ and $v_{SD}(\alpha)$ can be determined from these simulated waveforms. The inclusion of a simulation step in the calibration

process can significantly reduce the experimental measurements needed to determine the values of $A_{SD}(\alpha)$ and $v_{SD}(\alpha)$ for a variety of values for α . The SD behavior of a media can then be determined by simulation, without the need for an SD waveform generator.

The invention teaches a method for determining the SD impedance of a transmission line.

5 It also illustrates the predicable variation of Z_{SD} as a function of α . This predictability can allow a designer to control the transmission line SD impedance by appropriately selecting the parameter α .

The invention also describes an alternate precision distance measurement method using truncated SD test pulses with high loss, long lines. As the test pulses travel down these lines, the
10 attenuation can be so extensive that their amplitudes are too small for common threshold crossing time of flight measurements to be feasible. The invention includes a method for predicting an attenuation and overcoming this difficulty, extracting accurate length measurements for long lossy lines.

The invention demonstrates the utility of the SD test waveform in accurately determining
15 distances to impedance discontinuities. Examples show how to use SD wave propagation as the basis for an accurate time domain reflectometry (TDR) unit. This unit may include an SD waveform generator, a means of coupling with the transmission media, a means of measuring the applied and reflected waveforms, a display, memory storage, and computational ability to analyze and interpret the SD waveforms. The invention can also include a computer. The
20 invention can include a method which can be applied to modify currently available TDR units to increase their accuracy. This modification may be a TDR software modification resulting in a nonrecurring unit manufacturing cost. A standard TDR waveform may be used to determine an empirical transfer function for the line under test, and this transfer function may then be used to simulate the propagation of the SD waveform. This process allows all the advantages of SD to be
25 incorporated into currently available units with only a modification of existing computational algorithms, creating a virtual SD TDR unit.

The truncated SD test pulse approach can also be applied to the propagation of acoustic waves in sonar and geophysical applications. Reflected acoustic signals can be used to accurately determine the location of underwater objects and to determine locations of geophysical strata
30 boundaries. Empirical transfer functions can be determined and used to simulate SD pulse propagation eliminating the need to physically generate complex acoustic pulses.

The invention provides an SD waveform that has several propagation properties in lossy, frequency dependent transmission media, which make it suitable for enhancing a data transmission scheme.

The invention provides a method to increase the number of bits transmitted per pulse by varying the value of α to encode data on truncated SD waveforms. The example provided utilizes SD pulses with four different α 's sequentially transmitted on a 2kft cable. These SD pulses are received at the end of the cable with the exponential constants undistorted. The four pulses may be transmitted as positive or negative signals resulting in eight possible states. Therefore, each pulse represents three bits of transmitted information in each symbol period. This strategy can be used with short, low loss transmission lines.

The invention also teaches the use of SD propagation properties to generate a set of pulses consisting of linear combinations of truncated SD waveforms that are orthogonal at the receiver. Data can be encoded on these pulses by varying their amplitude, and these amplitude-modulated pulses may be transmitted simultaneously on a transmission line. The individual pulse amplitudes are computed at the receiver taking advantage of their orthogonality. This method provides improved noise immunity, and can be used with longer, higher loss transmission lines.

The invention also teaches the use of SD waveforms in digital circuits, specifically, on-chip clock circuits. It provides a method for utilizing truncated SD clock waveforms to reduce clock skew.

The propagation of a Speedy Delivery (SD) waveform in a transmission line can be described by the equation $V(x,t) = De^{-A_{SD}[x-v_{SD}t]}$. The associated boundary condition at the input to the line is $V(x=0,t) = De^{\alpha t}$, wherein A_{SD} is the exponential coefficient describing the attenuation of the SD signal as a function of distance along the line at a fixed time instant, and v_{SD} is the SD signal propagation velocity. These quantities are also related by the simple equation, $A_{SD}(\alpha) \cdot v_{SD}(\alpha) = \alpha$. The parameters A_{SD} and v_{SD} of the propagating wave and the line boundary condition are all functions of α , and may be determined experimentally for a transmission line.

For the case of two balanced twisted wire pair transmission lines, the differential voltage can be equally divided between the two wires and the current on each wire in the pair is equal in magnitude but opposite in direction. In this case, the telegrapher's equations can be written:

$$\frac{\partial}{\partial x} \mathbf{V} = \mathbf{Z} \cdot \mathbf{I}$$

$$\frac{\partial}{\partial x} \mathbf{I} = \mathbf{Y} \cdot \mathbf{V}$$

$$\frac{\partial}{\partial x} \begin{bmatrix} V_1(x, t) \\ V_2(x, t) \end{bmatrix} = \begin{bmatrix} R_1 + L_1 \frac{\partial}{\partial t} & M_{12} \frac{\partial}{\partial t} \\ M_{21} \frac{\partial}{\partial t} & R_2 + L_2 \frac{\partial}{\partial t} \end{bmatrix} \cdot \begin{bmatrix} I_1(x, t) \\ I_2(x, t) \end{bmatrix}.$$

$$\frac{\partial}{\partial x} \begin{bmatrix} I_1(x, t) \\ I_2(x, t) \end{bmatrix} = \begin{bmatrix} G_1 + C_1 \frac{\partial}{\partial t} & C_{12} \frac{\partial}{\partial t} \\ C_{21} \frac{\partial}{\partial t} & G_2 + C_2 \frac{\partial}{\partial t} \end{bmatrix} \cdot \begin{bmatrix} V_1(x, t) \\ V_2(x, t) \end{bmatrix}$$

- 5 This analysis includes the net cross capacitance, C_{ij} , and the net cross inductances, M_{ij} . The cross resistance and conductance terms are neglected. These equations are then Laplace transformed into the complex frequency domain.

$$\frac{\partial}{\partial x} \begin{bmatrix} V_1(x, s) \\ V_2(x, s) \end{bmatrix} = \begin{bmatrix} \bar{R}_1(s) + s\bar{L}_1(s) & s\bar{M}_{12}(s) \\ s\bar{M}_{21}(s) & \bar{R}_2(s) + s\bar{L}_2(s) \end{bmatrix} \cdot \begin{bmatrix} I_1(x, s) \\ I_2(x, s) \end{bmatrix}$$

$$\frac{\partial}{\partial x} \begin{bmatrix} I_1(x, s) \\ I_2(x, s) \end{bmatrix} = \begin{bmatrix} \bar{G}_1(s) + s\bar{C}_1(s) & s\bar{C}_{12}(s) \\ s\bar{C}_{21}(s) & \bar{G}_2(s) + s\bar{C}_2(s) \end{bmatrix} \cdot \begin{bmatrix} V_1(x, s) \\ V_2(x, s) \end{bmatrix}$$

Taking the partial derivative of the top equation with respect to x , and substituting the bottom

- 10 equation for the $\frac{\partial}{\partial x} \begin{bmatrix} I_1(x, s) \\ I_2(x, s) \end{bmatrix}$ term, gives the result:

$$\frac{\partial^2}{\partial x^2} \mathbf{V}(x, s) = \bar{\mathbf{Z}}(s)\bar{\mathbf{Y}}(s) \cdot \mathbf{V}(x, s)$$

$$\frac{\partial^2}{\partial x^2} \begin{bmatrix} V_1(x, s) \\ V_2(x, s) \end{bmatrix} = \begin{bmatrix} \bar{R}_1(s) + s\bar{L}_1(s) & s\bar{M}_{12}(s) \\ s\bar{M}_{21}(s) & \bar{R}_2(s) + s\bar{L}_2(s) \end{bmatrix} \cdot \begin{bmatrix} \bar{G}_1(s) + s\bar{C}_1(s) & s\bar{C}_{12}(s) \\ s\bar{C}_{21}(s) & \bar{G}_2(s) + s\bar{C}_2(s) \end{bmatrix} \cdot \begin{bmatrix} V_1(x, s) \\ V_2(x, s) \end{bmatrix}$$

or,

$$\frac{\partial^2}{\partial x^2} \mathbf{V}(x,s) = \bar{\Gamma}(s)^2 \cdot \mathbf{V}(x,s)$$

$$\frac{\partial^2}{\partial x^2} \begin{bmatrix} V_1(x,s) \\ V_2(x,s) \end{bmatrix} = \begin{bmatrix} (\bar{R}_1(s) + s\bar{L}_1(s))(\bar{G}_1(s) + s\bar{C}_1(s)) & s^2\bar{M}_{12}(s)\bar{C}_{21}(s) \\ s^2\bar{M}_{21}(s)\bar{C}_{12}(s) & (\bar{R}_2(s) + s\bar{L}_2(s))(\bar{G}_2(s) + s\bar{C}_2(s)) \end{bmatrix} \cdot \begin{bmatrix} V_1(x,s) \\ V_2(x,s) \end{bmatrix}$$

This system of equations has the solution:

$$\mathbf{V}(x,s) = \mathbf{e}^{-\bar{\Gamma}(s) \cdot x} \cdot \mathbf{A} + \mathbf{e}^{+\bar{\Gamma}(s) \cdot x} \cdot \mathbf{B}$$

Applying the boundary conditions for a semi-infinite pair of transmission lines, $V(x=0,s)$ at $x=0$, and $V(x \rightarrow \infty, s) = 0$ gives:

$$\mathbf{V}(x,s) = \mathbf{e}^{-\bar{\Gamma}(s) \cdot x} \cdot \mathbf{V}(x=0,s).$$

Because $\bar{\Gamma}$ is a complex symmetric matrix, it can be written $\bar{\Gamma} = (\bar{\mathbf{Z}}\bar{\mathbf{Y}})^{1/2} = \bar{\mathbf{P}}\bar{\gamma}\bar{\mathbf{P}}^{-1}$, where $\bar{\mathbf{P}}$ is the eigenvector matrix of the $\bar{\mathbf{Z}}\bar{\mathbf{Y}}$ product and $\bar{\gamma}$ is the diagonal eigenvalue matrix of $\bar{\Gamma}$. Therefore, we can write $\mathbf{e}^{-\bar{\Gamma}(s) \cdot x} = \mathbf{P}(s) \cdot \mathbf{e}^{-\bar{\gamma}(s) \cdot x} \cdot \mathbf{P}(s)^{-1}$. This gives the result:

$$10 \quad \mathbf{V}(x,s) = \mathbf{P}(s) \cdot \mathbf{e}^{-\bar{\gamma}(s) \cdot x} \cdot \mathbf{P}(s)^{-1} \cdot \mathbf{V}(x=0,s)$$

For the case where the applied boundary conditions at $x=0$ are SD waveforms,

$$V_1(x=0,t) = D_1 e^{\alpha_1 t}, \quad V_2(x=0,t) = D_2 e^{\alpha_2 t}, \quad \text{the SD solution is:}$$

$$\mathbf{V}(x,t) = \mathbf{P}(\alpha_1) \cdot \mathbf{e}^{-\gamma(\alpha_1) \cdot x} \cdot \mathbf{P}^{-1}(\alpha_1) \cdot \begin{bmatrix} D_1 \cdot e^{\alpha_1 \cdot t} \\ 0 \end{bmatrix} + \mathbf{P}(\alpha_2) \cdot \mathbf{e}^{-\gamma(\alpha_2) \cdot x} \cdot \mathbf{P}^{-1}(\alpha_2) \cdot \begin{bmatrix} 0 \\ D_2 \cdot e^{\alpha_2 \cdot t} \end{bmatrix},$$

where $\mathbf{P}(\alpha_i)$ is a real matrix.

$$15 \quad \mathbf{V}(x,t) = \begin{bmatrix} P_{11}(\alpha_1) & P_{12}(\alpha_1) \\ P_{21}(\alpha_1) & P_{22}(\alpha_1) \end{bmatrix} \cdot \begin{bmatrix} e^{-\gamma_1(\alpha_1) \cdot x} & 0 \\ 0 & e^{-\gamma_2(\alpha_1) \cdot x} \end{bmatrix} \cdot \begin{bmatrix} P_{11}(\alpha_1) & P_{12}(\alpha_1) \\ P_{21}(\alpha_1) & P_{22}(\alpha_1) \end{bmatrix}^{-1} \cdot \begin{bmatrix} D_1 \cdot e^{\alpha_1 \cdot t} \\ 0 \end{bmatrix}$$

$$+ \begin{bmatrix} P_{11}(\alpha_2) & P_{12}(\alpha_2) \\ P_{21}(\alpha_2) & P_{22}(\alpha_2) \end{bmatrix} \cdot \begin{bmatrix} e^{-\gamma_1(\alpha_2) \cdot x} & 0 \\ 0 & e^{-\gamma_2(\alpha_2) \cdot x} \end{bmatrix} \cdot \begin{bmatrix} P_{11}(\alpha_2) & P_{12}(\alpha_2) \\ P_{21}(\alpha_2) & P_{22}(\alpha_2) \end{bmatrix}^{-1} \cdot \begin{bmatrix} 0 \\ D_2 \cdot e^{\alpha_2 \cdot t} \end{bmatrix}$$

If we designate \mathbf{P}^{-1} as \mathbf{Q} , we can write:

$$\mathbf{V}(x,t) = \begin{bmatrix} P_{11}(\alpha_1) & P_{12}(\alpha_1) \\ P_{21}(\alpha_1) & P_{22}(\alpha_1) \end{bmatrix} \cdot \begin{bmatrix} e^{-\gamma_1(\alpha_1) \cdot x} & 0 \\ 0 & e^{-\gamma_2(\alpha_1) \cdot x} \end{bmatrix} \cdot \begin{bmatrix} Q_{11}(\alpha_1) & Q_{12}(\alpha_1) \\ Q_{21}(\alpha_1) & Q_{22}(\alpha_1) \end{bmatrix} \cdot \begin{bmatrix} D_1 \cdot e^{\alpha_1 \cdot t} \\ 0 \end{bmatrix} \\ + \begin{bmatrix} P_{11}(\alpha_2) & P_{12}(\alpha_2) \\ P_{21}(\alpha_2) & P_{22}(\alpha_2) \end{bmatrix} \cdot \begin{bmatrix} e^{-\gamma_1(\alpha_2) \cdot x} & 0 \\ 0 & e^{-\gamma_2(\alpha_2) \cdot x} \end{bmatrix} \cdot \begin{bmatrix} Q_{11}(\alpha_2) & Q_{12}(\alpha_2) \\ Q_{21}(\alpha_2) & Q_{22}(\alpha_2) \end{bmatrix} \cdot \begin{bmatrix} 0 \\ D_2 \cdot e^{\alpha_2 \cdot t} \end{bmatrix}$$

or,

$$\mathbf{V}(x,t) = \mathbf{P}(\alpha_1) \cdot \begin{bmatrix} Q_{11}(\alpha_1) \cdot e^{-\gamma_1(\alpha_1) \cdot x} \\ Q_{21}(\alpha_1) \cdot e^{-\gamma_2(\alpha_1) \cdot x} \end{bmatrix} \cdot D_1 \cdot e^{\alpha_1 \cdot t} \\ + \mathbf{P}(\alpha_2) \cdot \begin{bmatrix} Q_{12}(\alpha_2) \cdot e^{-\gamma_1(\alpha_2) \cdot x} \\ Q_{22}(\alpha_2) \cdot e^{-\gamma_2(\alpha_2) \cdot x} \end{bmatrix} \cdot D_2 \cdot e^{\alpha_2 \cdot t}$$

or,

$$\mathbf{V}(x,t) = \begin{bmatrix} Q_{11}(\alpha_1)P_{11}(\alpha_1)e^{-\gamma_1(\alpha_1) \cdot x} + Q_{21}(\alpha_1)P_{12}(\alpha_1)e^{-\gamma_2(\alpha_1) \cdot x} \\ Q_{11}(\alpha_1)P_{21}(\alpha_1)e^{-\gamma_1(\alpha_1) \cdot x} + Q_{21}(\alpha_1)P_{22}(\alpha_1)e^{-\gamma_2(\alpha_1) \cdot x} \end{bmatrix} \cdot D_1 \cdot e^{\alpha_1 \cdot t} \\ + \begin{bmatrix} Q_{12}(\alpha_2)P_{11}(\alpha_2)e^{-\gamma_1(\alpha_2) \cdot x} + Q_{22}(\alpha_2)P_{12}(\alpha_2)e^{-\gamma_2(\alpha_2) \cdot x} \\ Q_{12}(\alpha_2)P_{21}(\alpha_2)e^{-\gamma_1(\alpha_2) \cdot x} + Q_{22}(\alpha_2)P_{22}(\alpha_2)e^{-\gamma_2(\alpha_2) \cdot x} \end{bmatrix} \cdot D_2 \cdot e^{\alpha_2 \cdot t}$$

The real constants associated with the eigenvector matrix, P, its inverse, $P^{-1} = Q$, and the SD parameters D can be written:

$$\begin{aligned} A_{11} &= D_1 P_{11}(\alpha_1) Q_{11}(\alpha_1) \\ A_{12} &= D_1 P_{12}(\alpha_1) Q_{21}(\alpha_1) \\ A_{21} &= D_2 P_{11}(\alpha_2) Q_{12}(\alpha_2) \\ A_{22} &= D_2 P_{12}(\alpha_2) Q_{22}(\alpha_2) \\ B_{11} &= D_1 P_{21}(\alpha_1) Q_{11}(\alpha_1) \\ B_{12} &= D_1 P_{22}(\alpha_1) Q_{21}(\alpha_1) \\ B_{21} &= D_2 P_{21}(\alpha_2) Q_{12}(\alpha_2) \\ B_{22} &= D_2 P_{22}(\alpha_2) Q_{22}(\alpha_2) \end{aligned}$$

10 The signals on transmission lines will be:

$$\begin{aligned} V_1(x,t) &= A_{11} \cdot e^{-\gamma_1(\alpha_1) \cdot x} e^{\alpha_1 \cdot t} + A_{21} \cdot e^{-\gamma_2(\alpha_1) \cdot x} e^{\alpha_1 \cdot t} + A_{12} \cdot e^{-\gamma_1(\alpha_2) \cdot x} e^{\alpha_2 \cdot t} + A_{22} \cdot e^{-\gamma_2(\alpha_2) \cdot x} e^{\alpha_2 \cdot t} \\ V_2(x,t) &= B_{11} \cdot e^{-\gamma_1(\alpha_1) \cdot x} e^{\alpha_1 \cdot t} + B_{21} \cdot e^{-\gamma_2(\alpha_1) \cdot x} e^{\alpha_1 \cdot t} + B_{12} \cdot e^{-\gamma_1(\alpha_2) \cdot x} e^{\alpha_2 \cdot t} + B_{22} \cdot e^{-\gamma_2(\alpha_2) \cdot x} e^{\alpha_2 \cdot t} \end{aligned}$$

or,

$$V_1(x, t) = A_{11} \cdot e^{-A_{SD11}(x-v_{SD11}t)} + A_{21} \cdot e^{-A_{SD21}(x-v_{SD21}t)} \\ + A_{12} \cdot e^{-A_{SD12}(x-v_{SD12}t)} + A_{22} \cdot e^{-A_{SD22}(x-v_{SD22}t)}$$

$$V_2(x, t) = B_{11} \cdot e^{-A_{SD11}(x-v_{SD11}t)} + B_{21} \cdot e^{-A_{SD21}(x-v_{SD21}t)} \\ + B_{12} \cdot e^{-A_{SD12}(x-v_{SD12}t)} + B_{22} \cdot e^{-A_{SD22}(x-v_{SD22}t)}$$

Where:

$$A_{SD11} = \gamma_1(\alpha_1), A_{SD12} = \gamma_1(\alpha_2), A_{SD21} = \gamma_2(\alpha_1), \text{ and } A_{SD22} = \gamma_2(\alpha_2).$$

$$5 \quad v_{SD11} = \frac{\alpha_1}{\gamma_1(\alpha_1)}, v_{SD12} = \frac{\alpha_2}{\gamma_1(\alpha_2)}, v_{SD21} = \frac{\alpha_1}{\gamma_2(\alpha_1)}, \text{ and } v_{SD22} = \frac{\alpha_2}{\gamma_2(\alpha_2)}.$$

Thus, there are SD attenuations and velocities associated with each eigenvalue, γ_i , of the product \overline{ZY} evaluated at each α .

Because positive exponential waveforms continually increase, practical considerations may create the need to truncate or limit the waveform at some level determined by the specific application. Truncation may be accomplished by several methods, as is known in the art.

The result of this truncation of the input of a transmission media is that the propagating exponential (SD) signal is also limited in magnitude as it travels in the media, and in lossy media, this maximum amplitude decreases with distance traveled exhibiting an attenuating behavior.

15 The propagating (SD) signal is described in the coordinate frame, (x, t) , by $V(x, t) = De^{-A_{SD}(\alpha)[x-v_{SD}(\alpha)t]} = De^{\alpha \cdot t} e^{-A_{SD}(\alpha)x}$, where $A_{SD}(\alpha)v_{SD}(\alpha) = \alpha$.

Referring to **FIG. 1A**, a graph of an SD waveform for $(x=0, t)$ is depicted. The truncated input waveform at $x=0$ reaching a limit at $t=t_i$ is shown.

In the case of a four parameter (LCRG) transmission line, we have the expressions

$$20 \quad A_{SD}(\alpha) = \sqrt{LC\alpha^2 + (RC + LG)\alpha + RG}, \text{ and } v_{SD}(\alpha) = \frac{\alpha}{\sqrt{LC\alpha^2 + (RC + LG)\alpha + RG}}.$$

The attenuation of the traveling SD wave may be viewed from the perspective of a moving reference frame traveling at the speed $\frac{1}{\sqrt{LC}}$.

Introducing the new coordinates:

$$\begin{aligned} t' &= t - x\sqrt{LC} \\ x' &= x \end{aligned}$$

Then

$$V(x', t') = D e^{-A_{SD}(\alpha)} \cdot [x' - v_{SD}(\alpha) \cdot (t' + x' \sqrt{LC})],$$

5 or

$$V(x', t') = D e^{\alpha \cdot t'} e^{-A_{SD}(\alpha)} \cdot [1 - v_{SD}(\alpha) \cdot \sqrt{LC}] x'.$$

If we view the traveling wave after traveling a distance, $x = x' = \ell$, in the transmission media, then

$$V(\ell, t') = D e^{\alpha \cdot t'} e^{-A_{SD}(\alpha)} \cdot [1 - v_{SD}(\alpha) \cdot \sqrt{LC}] \ell.$$

10

Setting $t' = t_i$, yields the maximum value of the truncated SD waveform at $x = \ell$, and for any value $0 \leq t_j \leq t_i$, the value of the signal at the associated point t_j in the moving reference frame at $x = \ell$ is:

$$V(\ell, t_j) = D e^{\alpha \cdot t_j} e^{-A_{SD}(\alpha)} \cdot [1 - v_{SD}(\alpha) \cdot \sqrt{LC}] \ell$$

15

Referring to **FIG. 1B**, a graph of an SD waveform in moving frame at $x' = \ell$ ($x = \ell, t'$) is depicted. The attenuation of this truncated waveform is defined by the ratio:

$$\frac{V(\ell, t_j)}{V(0, t_j)} = \frac{D e^{\alpha \cdot t_j} e^{-A_{SD}(\alpha)} \cdot [1 - v_{SD}(\alpha) \cdot \sqrt{LC}] \ell}{D e^{\alpha \cdot t_j}}$$

20

$$= e^{-A_{SD}(\alpha) \cdot [1 - v_{SD}(\alpha) \cdot \sqrt{LC}] \ell}$$

This attenuation of the signal at a point t_j of a truncated SD waveform traveling in a transmission line with frequency dependent parameters whose Laplace transforms are $\bar{L}(s)$, $\bar{C}(s)$, $\bar{R}(s)$, and $\bar{G}(s)$, is:

$$\frac{V(\ell, t_j)}{V(0, t_j)} = e^{-\bar{A}_{SD}(\alpha) \cdot [1 - \bar{v}_{SD}(\alpha) \cdot \sqrt{\bar{L}\bar{C}(\alpha)}] \ell}$$

25

where:

$$\bar{A}_{SD}(\alpha) = \sqrt{\overline{LC}(\alpha) \cdot \alpha^2 + (\overline{RC}(\alpha) + \overline{LG}(\alpha)) \cdot \alpha + \overline{RG}(\alpha)}, \text{ and}$$

$$\bar{v}_{SD}(\alpha) = \frac{\alpha}{\sqrt{\overline{LC}(\alpha) \cdot \alpha^2 + (\overline{RC}(\alpha) + \overline{LG}(\alpha)) \cdot \alpha + \overline{RG}(\alpha)}}.$$

Furthermore, in this transmission media with frequency dependent parameters, if the
 5 input exponential pulse is rapidly closed when the exponential amplitude truncation limit is reached, then the end of the exponential region of the leading edge of the transmitted pulse is further reduced in amplitude and rounded by the chromatic dispersion.

Coupled transmission lines can be treated in a similar fashion. The equations governing SD propagation on one of two coupled transmission lines is:

$$10 \quad V_1(x, t) = A_{11} \cdot e^{-A_{SD11}(x - v_{SD11}t)} + A_{21} \cdot e^{-A_{SD21}(x - v_{SD21}t)} + A_{12} \cdot e^{-A_{SD12}(x - v_{SD12}t)} + A_{22} \cdot e^{-A_{SD22}(x - v_{SD22}t)}.$$

Applying the substitution: $t' = t - x\sqrt{\overline{LC}}$, $x' = x$, gives the result:

$$\frac{V_1(x' = \ell, t' = t_i)}{V_1(x = 0, t = t_i)} = \frac{A_{11} \cdot e^{\alpha_1 t_i} \cdot e^{-A_{SD11}(1 - v_{SD11} \sqrt{\overline{LC}})\ell} + A_{21} \cdot e^{\alpha_1 t_i} \cdot e^{-A_{SD21}(1 - v_{SD21} \sqrt{\overline{LC}})\ell}}{A_{11} \cdot e^{\alpha_1 t_i} + A_{21} \cdot e^{\alpha_1 t_i} + A_{12} \cdot e^{\alpha_2 t_i} + A_{22} \cdot e^{\alpha_2 t_i}} + \frac{A_{12} \cdot e^{\alpha_2 t_i} \cdot e^{-A_{SD12}(1 - v_{SD12} \sqrt{\overline{LC}})\ell} + A_{22} \cdot e^{\alpha_2 t_i} \cdot e^{-A_{SD22}(1 - v_{SD22} \sqrt{\overline{LC}})\ell}}{A_{11} \cdot e^{\alpha_1 t_i} + A_{21} \cdot e^{\alpha_1 t_i} + A_{12} \cdot e^{\alpha_2 t_i} + A_{22} \cdot e^{\alpha_2 t_i}}$$

Which can be written:

15

$$\frac{V_1(\ell, t_j)}{V_1(0, t_j)} = \frac{\left(A_{11} \cdot e^{-A_{SD11}(1 - v_{SD11} \sqrt{\overline{LC}})\ell} + A_{21} \cdot e^{-A_{SD21}(1 - v_{SD21} \sqrt{\overline{LC}})\ell} \right)}{(A_{11} + A_{21}) + (A_{12} + A_{22}) \cdot e^{(\alpha_2 - \alpha_1)t_i}} + \frac{\left(A_{12} \cdot e^{-A_{SD12}(1 - v_{SD12} \sqrt{\overline{LC}})\ell} + A_{22} \cdot e^{-A_{SD22}(1 - v_{SD22} \sqrt{\overline{LC}})\ell} \right)}{(A_{11} + A_{21}) \cdot e^{(\alpha_1 - \alpha_2)t_i} + (A_{12} + A_{22})}$$

Another example includes of the evolution of a slowly varying envelope, $E(x, t)$, of an optical pulse in a single mode communication fiber. The Schrodinger partial differential equation [3],

$$5 \quad \frac{\partial E}{\partial x} + \beta_1 \frac{\partial E}{\partial t} - \frac{j}{2} \beta_2 \frac{\partial^2 E}{\partial t^2} - \frac{1}{6} \beta_3 \frac{\partial^3 E}{\partial t^3} = -\frac{\gamma}{2} E,$$

describes the evolution of the shape of the propagating pulse envelope undergoing chromatic dispersion in fiber. Chromatic dispersion occurs because the mode-propagation constant, $\beta(\omega)$, is a nonlinear function of the angular frequency ω , where

10

$$\beta_n = \left[\frac{\partial^n \beta}{\partial \omega^n} \right]_{\omega=\omega_0}$$

and ω_0 is the frequency of the light being modulated in the fiber.

ω_0 / β_0 is the phase velocity of the pulse. $1 / \beta_1$ is the group velocity, β_2 is the group velocity dispersion (GVD) parameter which causes the pulse to broaden as it propagates in the fiber. β_3 is the third-order dispersion (TOD) parameter [4]

The higher order derivatives of β with respect to ω are assumed negligible. However, they may be added in these analyses by anyone familiar with the art. The fiber attenuation is represented by the parameter γ .

20

The SD solution of the linear Schrodinger equation with loss ($\gamma \neq 0$) is:

$$E(x, t) = D e^{\left[\frac{t}{T_0} - \left(\frac{\beta_1}{T_0} - \frac{\beta_3}{6T_0^3} + \frac{\gamma}{2} \right) \cdot x \right]} e^{j \left(\frac{\beta_2 \cdot x}{2T_0^2} \right)}.$$

The boundary condition at the input to the fiber at $x = 0$ is $E(0, t) = D e^{\frac{t}{T_0}}$, (see FIG. 1A with $\alpha = \frac{1}{T_0}$). The receiver detector in a fiber optic communication network responds to the

square of the magnitude $|E(x, t)|^2$ of the propagating pulse envelope. In this case

$$|E(x, t)|^2 = D^2 e^{-2 \cdot A(T_0) [x - v_{sd}(T_0) \cdot t]},$$

5 where the velocity of propagation of a truncated SD leading edge of the envelope is

$$v_{sd}(T_0) = \frac{1}{\left(\beta_1 - \frac{\beta_3}{6T_0^2} + \frac{T_0 \gamma}{2} \right)}, \text{ and } A(T_0) = \left(\frac{\beta_1}{T_0} - \frac{\beta_3}{6T_0^3} + \frac{\gamma}{2} \right).$$

Note that $v_{sd}(T_0) \cdot A(T_0) = \frac{1}{T_0}$.

10

Transforming to a moving reference frame traveling at the group velocity $\frac{1}{\beta_1}$,

$$\begin{aligned} t' &= t - \beta_1 x \\ x' &= x \end{aligned}$$

The SD propagation speed in this moving reference frame is

15
$$\bar{v}_{sd}(T_0)_{\text{in } x', t' \text{ frame}} = \frac{1}{T_0 \left(\frac{\gamma}{2} - \frac{\beta_3}{6T_0^3} \right)},$$

The attenuation of the truncated SD portion of the leading edge of the pulse envelope is

$$\frac{|E(\ell, t_j)|^2}{|E(0, t_j)|^2} = e^{-\left(\frac{\gamma}{2} - \frac{\beta_3}{6T_0^3} \right) \ell}.$$

This result implies that if T_0 can be made small enough, this attenuation may be reduced. Higher order terms of $\beta(\omega)$, which were not included in this particular analysis, may become more significant as T_0 becomes smaller.

5

EXAMPLES

Specific embodiments of the invention will now be further described by the following, nonlimiting examples which will serve to illustrate in some detail various features. The following examples are included to facilitate an understanding of ways in which the invention may be practiced. It should be appreciated that the examples which follow represent embodiments
10 discovered to function well in the practice of the invention, and thus can be considered to constitute preferred modes for the practice of the invention. However, it should be appreciated that many changes can be made in the exemplary embodiments which are disclosed while still obtaining like or similar result without departing from the spirit and scope of the invention. Accordingly, the examples should not be construed as limiting the scope of the invention.

15

Example 1

Method for determining SD waveform parameters, A_{SD} and v_{SD}

This example teaches how SD test pulses may be used to obtain empirical estimates of the transmission line parameters, A_{SD} and v_{SD} , that describe the propagation of SD waveforms in
20 lossy, dispersive lines, including those with frequency dependent parameters. The numerical values of A_{SD} and v_{SD} as a function of α can be determined empirically.

Referring to **FIG. 2**, an experimental set up that may be used to determine these parameters for a copper, twisted wire pair transmission line is depicted. The test length of the transmission line, d , and the value of α for the applied SD input signal are measured. An
25 additional length of transmission line that is identical to the line under test is attached as shown and terminated with an open circuit. In this example the cables are two 24 AWG, individually shielded, twisted wire pairs in a 1002 ft coil of T1 cable.

The values of D , α and the duration of the SD signal were chosen to prevent the occurrence of reflections at the measurement point, d , during the initial time of flight of the SD
30 signal propagating in the test line. The additional length of transmission line also ensures that all

reflections occurring in the test line are delayed until well after the propagating SD waveform measurement at d is complete.

The SD signal time of flight can be directly measured by timing the propagating SD waveform crossing of constant voltage thresholds (**FIG. 3**). The value of the SD signal propagation velocity in the line, v_{SD} , can be calculated using the relationship,

5 $TOF = (t_f - t_o) = v_{sd}d$. The times of flight for these waveforms shown again in **FIG. 4A** at thresholds from 1.2 volts to 4.8 volts are plotted in **FIG. 4B**. The average TOF for the threshold overlap region shown in **FIG. 4.3b** is 1,716 ns. This TOF results in

$$v_{SD} = \frac{d}{TOF_{AVG}} = \frac{1002\text{ft}}{1,716 \times 10^{-9}\text{sec}} = 5.839 \times 10^8 \frac{\text{ft}}{\text{sec}}.$$

10 The SD attenuation coefficient, A_{SD} , can be calculated from α and v_{sd} using the relationship $A_{SD} \cdot v_{SD} = \alpha$. The calculated A_{SD} obtained from the waveforms shown in **FIG.**

4.3a is $A_{SD} = \frac{\alpha}{v_{SD}} = \frac{3.0 \times 10^6 \frac{1}{\text{sec}}}{5.839 \times 10^8 \frac{\text{ft}}{\text{sec}}} = 5.138 \times 10^{-3} \frac{1}{\text{ft}}.$

Once this velocity is known for a calibrated length of the cable, this type of threshold crossing TOF measurement can be used to determine the unknown length of another sample of
15 the same transmission line.

Referring to **FIG. 5A**, the measured input voltage of the two twisted wire pairs in a T1 line connected in series is depicted (see **FIG. 2**) and one-half of the voltage measured at the end of the open terminated second twisted wire pair in the T1 line ($R_t = \infty$ in **FIG. 2**). The average TOF for the threshold overlap region shown in **FIG. 5B** is 3,418 ns. This time of flight together
20 with the previously calibrated v_{SD} , results in an estimated distance of 1,996 ft, or 99.6% of the actual length (2004 ft).

FIG. 5A also indicates a potential difficulty in longer lines caused by the attenuation of the truncated SD pulse. The region of SD threshold overlap grows smaller as the pulse propagates, and is attenuated over longer distances. The accuracy of this measurement can be
25 improved, and the difficulty caused by the reduced region of threshold overlap can be overcome for very long transmission lines using the TOF measurement technique discussed in example 4.

The high accuracy of this method results from the use of an average of TOF measurements taken over a range of signal threshold amplitudes. It is feasible to use the average of these values to improve the accuracy of the measurement of TOF since all points on the leading SD edge of the waveform travel at the same speed. This is in contrast with the current time domain reflectometry art in which a measurement of the time of emergence of the dispersing test pulse is attempted. This is equivalent to attempting to accurately measure the threshold crossing time of the pulse at a zero threshold level when the slope of the pulse is also nearly zero.

Example 2

Extracting SD Parameters From Empirical Transfer Function

The method described in example 1 can be limiting in that it should be experimentally repeated for each exponential coefficient α . This may require a lengthy laboratory calibration time for each cable type. Simulation can be included in the calibration process to significantly reduce the experimental measurements needed to determine the values of A_{SD} and v_{SD} . The use of simulation in this process requires a transfer function that describes the response of the specific type of cable. The α 's analyzed for the T1 cable range from $1 \times 10^5 \text{ sec}^{-1}$ to $1 \times 10^7 \text{ sec}^{-1}$. This range of α 's corresponds to the frequency band of 16 kHz to 1.6 MHz, so the transfer function must be accurate for this range. The transfer function of the cable may available from the manufacturer or can be determined empirically by applying a known pulse and measuring the applied cable input waveform and the response waveform at a known (calibration) distance along the cable. The transfer function is then calculated from the ratio of the fast Fourier transforms

(FFT) of the input-output pulses:
$$H_{\text{est}}(j\omega) = \frac{\text{FFT}\left(V_{\text{Measured at } \ell}\right)}{\text{FFT}\left(V_{\text{Applied}}\right)}$$
. Once the transfer function

is determined, the propagation of SD waveforms with a series of different values for α can be simulated and the times of flight and SD parameters, A_{SD} and v_{SD} , determined.

This method can be demonstrated for the T1 cable used in example 1. The end of the total length has been terminated with a 100-Ohm resistor (R_t in FIG. 4.1) to minimize reflections at the 1002 ft measurement point. FIG. 6A shows a measured input pulse applied to the T1. FIG.

6B shows the output pulse measured at 1002 ft. FIGS. 7A-B show the Power Spectral Density of the measured pulses. The range of α 's is indicated to show that there is sufficient energy in the frequency band of interest to empirically estimate an accurate transfer function. The magnitude and the phase of the transfer function determined from the measured signals are shown in FIG. 8.

5 Once the transfer function is known, a waveform with any α in the indicated range of interest can be simulated and the value of A_{SD} and v_{SD} determined. The values of A_{SD} and v_{SD} of the measured SD waveforms of section three provide a test of the transfer function for an α of $3 \times 10^6 \text{ sec}^{-1}$. FIG. 9 shows the time of flight plots determined using the input signal measured in section two, and the 1002 ft waveform predicted by simulation using the empirical transfer
 10 function. The simulation TOF result (FIG. 9) of 1,714 nsec is almost identical to the experimentally measured (FIG. 5B) TOF of 1,716 nsec. FIG. 10 shows the difference between the measured waveform at 1002 ft and the simulated waveform at 1002 ft (FIG. 9, Top). During the SD propagation region the difference is on the order of $\pm 20 \text{ mV}$.

This process is performed for the range of α 's by simulating the initial SD waveforms and
 15 using the transfer function to predict the waveforms at 1002 ft. In all cases, the simulated initial waveforms were closed with a 1 μsec ramp to reduce their high frequency components. FIGS. 11-12 show the simulated input and 1002 ft curves for the endpoint α 's of $1 \times 10^5 \text{ sec}^{-1}$ and $1 \times 10^7 \text{ sec}^{-1}$ along with their respective constant threshold times of flight curves. The values of v_{SD} and A_{SD} from simulation are presented in FIGS 13A-B, illustrated in Table I for the range of α 's, as
 20 set forth below (* values in parenthesis obtained using methods of example 1, with measured input and output SD signals along actual T1 Line):

Table I
 v_{SD} and A_{SD} Determined By Simulation with Empirical Transfer Function

| Alpha (sec^{-1}) | 1002 Simulation TOF (μsec) | Simulation v_{SD} ($\times 10^8 \text{ ft/sec}$) | Simulation A_{SD} ($\times 10^{-3} \text{ 1/ft}$) |
|--------------------------------|---|---|--|
| 1×10^5 | 4.487 | 2.233 | 0.448 |
| 2×10^5 | 3.068 | 3.266 | 0.612 |
| 3×10^5 | 2.608 | 3.842 | 0.781 |

| Alpha (sec ⁻¹) | 1002 Simulation TOF (μsec) | Simulation v _{SD} (x10 ⁸ ft/sec) | Simulation A _{SD} (x10 ⁻³ 1/ft) |
|-------------------------------|----------------------------------|---|--|
| 4x10 ⁵ | 2.382 | 4.207 | 0.951 |
| 5x10 ⁵ | 2.271 | 4.412 | 1.133 |
| 6x10 ⁵ | 2.172 | 4.613 | 1.301 |
| 7x10 ⁵ | 2.103 | 4.765 | 1.469 |
| 8x10 ⁵ | 2.040 | 4.912 | 1.629 |
| 9x10 ⁵ | 1.999 | 5.013 | 1.796 |
| 1x10 ⁶ | 1.973 | 5.079 | 1.969 |
| 2x10 ⁶ | 1.792 | 5.592 | 3.577 |
| 3x10 ⁶ | 1.714 (1.716*) | 5.849 (5.839*) | 5.129 (5.138*) |
| 4x10 ⁶ | 1.666 | 6.014 | 6.651 |
| 5x10 ⁶ | 1.634 | 6.132 | 8.154 |
| 6x10 ⁶ | 1.610 | 6.224 | 9.641 |
| 7x10 ⁶ | 1.592 | 6.294 | 11.122 |
| 8x10 ⁶ | 1.577 | 6.354 | 12.591 |
| 9x10 ⁶ | 1.565 | 6.403 | 14.057 |
| 1x10 ⁷ | 1.555 | 6.444 | 15.519 |

Example 3

Measurement of SD Transmission Line Impedance

- 5 The SD line impedance, $Z_{SD}(\alpha)$, is a real number which may be determined for the α 's of interest. If the Laplace transform of the frequency dependent transmission line parameters $\bar{R}(s)$, $\bar{L}(s)$, $C(s)$, and $\bar{G}(s)$, are known, then $Z_{SD}(\alpha)$ may be computed

$$\text{from, } Z_{SD} = \sqrt{\frac{\alpha \bar{L}(\alpha) + \bar{R}(\alpha)}{\alpha \bar{C}(\alpha) + \bar{G}(\alpha)}}.$$

The SD line impedance, $Z_{SD}(\alpha)$, may also be experimentally determined for a test transmission line by measuring the SD portion of the voltage waveform across various known termination impedances and computing the value $Z_{SD}(\alpha)$ from the measurements. The measured SD waveform at a termination, $V_{SD}(R_t)$, consists of the sum of the incident SD waveform, V_{SD}^+ , and a reflected SD wave. The reflected wave is the product of the SD reflection coefficient, Γ , and the incident wave. The measured SD signal at the termination will be $V_{SD}(R_t) = (1 + \Gamma_t) \cdot V_{SD}^+ = \frac{2 \cdot R_t}{R_t + Z_{SD}} \cdot V_{SD}^+$. Note that all lumped and distributed impedance values are real for SD signals.

The experimental setup used to determine Z_{SD} from line measurements is shown in **FIG.**

14. The measured input and termination waveforms of the voltages for a 1002 ft of T1 line, are shown in **FIG. 15A**. These measurements were made with an α of 1.5×10^6 1/sec and termination resistances of 49.5 ohms, 98.6 ohms and an open circuit. The magnitude of the waveform with no reflection, (the incident waveform V_{SD}^+), is determined by dividing the signal measured with an open termination, ($R_t = \infty, \Gamma_\infty = 1$), in half. The ratio of the waveforms $\frac{V_{SD}(R_t)}{V_{SD}^+}$, with known terminations and the incident waveform gives a direct measurement of the reflection coefficient, Γ_t , for each termination resistance, $\Gamma_t = \frac{V_{SD}(R_t)}{V_{SD}^+} - 1$. **FIG. 15B** shows the measured ratios for this cable and α . Using the relationship for the transmission coefficient, $1 + \Gamma_t = \frac{2 \cdot R_t}{R_t + Z_{SD}}$, gives $Z_{SD} = \frac{R_t}{1 + \Gamma_t} (1 - \Gamma_t)$.

Using the two finite values for R_t , 98.6 Ω and 49.5 Ω , yields two estimates for Z_{SD} :

$$Z_{SD} \left(\alpha = 1.5 \times 10^6 \frac{1}{\text{sec}} \right) = \frac{98.6}{1 + (0.919 - 1)} (1 - (0.919 - 1)) = 116 \Omega, \text{ or}$$

$$Z_{SD} \left(\alpha = 1.5 \times 10^6 \frac{1}{\text{sec}} \right) = \frac{49.5}{1 + (0.598 - 1)} (1 - (0.598 - 1)) = 116 \Omega, \text{ for this T1 cable.}$$

This process was repeated for α 's from $5 \times 10^5 \text{ sec}^{-1}$ to $10 \times 10^6 \text{ sec}^{-1}$ and the results are shown in **FIG 16**.

Example 4

5 **Precision measurement of long transmission line length or distance to an impedance change using a reflected wave**

This example describes an alternate precision TOF measurement method of SD test pulses for higher loss, longer lines whose attenuation of the test pulses is so extensive that their amplitudes become too small as they travel down the line for a common threshold crossing measurement to be feasible.

The time of flight measured at a constant voltage threshold becomes more difficult to estimate as the waveform is attenuated. This can be seen by examining the voltage traces along an approximately 6-kft, 24 AWG, fifty twisted wire pair, telephone cable (See **FIG. 17**). Three pairs of the fifty have been connected in series to generate an approximately 18-kft cable with three equal length sections. The applied SD input waveform is $De^{\alpha t}$, with a D of 0.19 volts and an α of $3.5 \times 10^5 \text{ 1/sec}$. This α is approximately one tenth of the α used for the 2-kft T1 cable. The smaller α is required here to reduce the attenuation of the truncated SD leading edge of the propagating pulse over these longer distances as discussed in section two. This allows signal detection over longer distances. The value of A_{SD} and v_{SD} for this cable and α were determined using the procedure of example 1. These measured values are based on assuming a 6000 ft calibration length for each twisted pair length inside the telephone cable. The results are $v_{SD} = 4.348 \times 10^8 \text{ ft/sec}$ and $A_{SD} = 8.279 \times 10^{-4} \text{ 1/ft}$.

Even though the signal was detectable at much greater distances than 6 kft, the constant threshold time of flight could not be measured. **FIG. 17** shows the voltage traces measured at the input, ~6 kft, ~12 kft, and ~18 kft distances, and it also shows the preservation of shape of the SD region of the highly attenuated propagating waveform. The ~18 kft trace is corrected for the reflection from the open termination by dividing the measured waveform in half. The last two traces show that even with this smaller α , beyond 12,000 ft there is little or no SD signal above 0.19 volts. This is the value of the input SD waveform parameter D, thus there is little or no SD voltage threshold common to the input SD waveform and the SD waveforms at these longer

distances.

The first step in precisely measuring highly attenuated SD pulse time of flight is to detect the SD region of the propagating pulse, $V(t)$, measured and recorded at the distance ℓ along the

cable. This is done by computing the ratio $\frac{dV(t)/dt}{V(t)}$ of the measured pulse waveform. In the SD

5 region of the pulse, this ratio is α . The end of the SD region is then found by detecting the time

that this ratio diverges from α . The ratio $\frac{d^2V(t)/dt^2}{dV(t)/dt}$ can also be used to detect the end of the SD

region. In the SD region, this ratio is also α . The later ratio of the second and first derivative responds more quickly, but is also more susceptible to noise than the ratio of the first derivative and the signal. These ratios as a function of time are plotted in **FIG. 18** for the SD waveform

10 measured at 12-kft. In general, the ratio of $\frac{\partial^n V(x,t)}{\partial^{n-1} V(x,t)}$ for any positive n will have this

property. Additionally, the ratio of the signal and its integral can be used to locate the SD region.

In the SD region, we have the relationship: $\frac{De^{\alpha t}}{\int_0^t De^{\alpha \tau} d\tau} = \frac{De^{\alpha t}}{\frac{1}{\alpha}(De^{\alpha t} - D)} = \alpha \frac{1}{1 - e^{-\alpha t}}$. Thus, the

ratio will converge on α as time progresses. This ratio diverges at the end of the SD region. In a
 15 lossy transmission medium with constant transmission line parameters, the end of the estimated
 SD region can provide a good marker to determine the speed of light in the medium. The
 truncated SD leading edge of the pulse does not disperse, even with frequency dependent
 parameters. The high frequency components of the closing pulse, however, do undergo
 dispersion. These fast, high frequency components tend to erode the end of the truncated SD
 region as the pulse propagates. At long distances, these high frequency components are also more
 20 highly attenuated so the amount of SD region erosion is not proportional to the distance traveled.
 In this case, the detected end of the SD region serves as an initial estimate of the pulse time of
 flight or the distance the wave has traveled and serves to define the region examined for more
 precise time of flight measurements in the next step.

The attenuation of the truncated SD signal propagating a distance d was shown to be $e^{-A_{sd} \cdot (1 - v_{sd} \cdot \sqrt{LC})d}$. In the application of time domain reflectometry (TDR) to these lines, the parameter $1/\sqrt{LC}$ is commonly referred to as V_p and is given as a fraction of the speed of light in a vacuum. The standard value of V_p used in TDR for this air core poly 24 AWG cable is 0.67. However, the value of the ratio used in this analysis is 0.59. The reason for this can be seen in **FIG. 19**. The value of 0.67 is appropriate for the high frequencies that are found in a standard TDR pulse used to test these telephone lines, but the α selected here for low truncated SD signal attenuation in these long lines is equivalent to a low frequency and propagates at a slower speed.

Once the initial TOF estimate of end of the SD region in the leading edge of the propagating pulse is obtained at the distance ℓ , the input waveform is attenuated and time shifted assuming the end of the SD region of the traveling wave corresponds to the point of truncation of the initial SD waveform, (**FIG. 20**, top). The input waveform is then incrementally translated in time and attenuated based on each new estimated distance traveled. At each distance estimate, this shifted and attenuated input waveform is subtracted from the measured waveform. If there is no dc voltage offset and no noise, this difference will be zero in the SD region when the estimated distance is equal to the actual distance. Because there is always noise and there is frequently a varying differential, dc offset at different points along the cable, the variance of this difference is evaluated at each estimated distance. This correlation has a minimum when the input waveform has been translated to the actual distance traveled, (**FIG. 20**, middle). If there is a dc differential offset between the lines, this minimum will be non-zero. At this estimated distance, the erosion of the end of the truncated SD region due to the dispersion from the closing pulse is obvious. If the waveform is translated too far, this deviation error measurement increases, (**FIG. 20**, bottom). **FIG. 21** shows the calculated standard deviation of the difference of the shifted and attenuated input waveform and the SD waveform measured at the splice between the second and third twisted wire pairs as a function of estimated distance. The minimum occurs at an estimated distance of 12,216 ft. **FIG. 22A** shows the difference of the two waveforms and the variance of the difference of the waveforms at the minimum. **FIG. 22A** indicates that there is a five to ten millivolt differential dc offset between the two measuring points. **FIG. 22B** is presented to show how the variance has increased when the estimated distance is 38 ft longer than the distance associated with the minimum.

This process was repeated eight times giving average distances of 6,113 ft, 12,216 ft, and 18,327 ft for the best-fit shift locations of the attenuated input waveform to the measured waveforms. The standard deviations of the three lengths were 1 ft, 1ft, and 5 ft respectively. The fits associated with one of these data sets are shown in **FIG. 23**. The actual length of the cable was not exactly known, however, each pair in the cable is parallel and the only differences in length are small and due to different twist rates. This process yields the distance of two lengths within 10 ft of two times the distance of one length. The distance of three lengths is within 11 ft of three times the distance of one length. Using the ~6 kft distance as a standard gives a discrepancy of -0.08 % for the ~12 kft length and -0.06 % for the ~18 kft length.

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Example 5

Time Domain Reflectometer (TDR): Precision measurement of transmission line distance to impedance change using a reflected wave

A Time Domain Reflectometer (TDR) is a test instrument used to find faults in transmission lines and to empirically estimate transmission line lengths and other parameters characterizing the line such as inductance per unit length, capacitance per unit length, resistance per unit length and conductance per unit length. A fundamental measurement in TDR test technology is the time of flight (TOF) of a test pulse generated by the instrument and applied to the line. This time of flight may be measured by timing the passage of the pulse detected at two locations along the line, referred to as Time Domain Transmission measurements (TDT). Or by Time Domain Reflection measurements which estimate the launch time of a pulse at one position and the subsequent return time of the pulse back to the same location after the pulse has been reflected from a fault or other impedance change some distance along the line. These measured TOF values, along with a value of the propagation speed of the pulse, allows one to obtain the distance between measurement points or in the case of the reflected wave, the distance from the pulse launch point to the location of the impedance change causing the pulse to be reflected and returned to the launch point.

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A fundamental limitation in TDR technology is the poor accuracy of these TOF measurements in lossy, dispersive transmission lines. The relatively high TDR accuracy of TOF values obtainable in short low loss, low dispersion transmissions lines is possible only because

the propagating test pulses keep their shape and amplitude in tact over the distances they travel during TOF measurements. By contrast, in dispersive, lossy long transmission lines the test pulses used in the art change shape, change amplitude and speed as they travel. The TOF measurements used in the art under these circumstances focus on estimating the emergence time of the leading profile of the test pulses. This part of the signature of test pulses used in the art has characteristically a low signal level and low signal slope making an accurate pulse emergence time measurement difficult to obtain.

Several advantages can be obtained in TDR technology for lossy, dispersive transmission lines by using a test pulse that contains a truncated SD signal in its leading edge. This truncated SD leading edge travels at a constant speed along these transmission lines without changing shape. The speed of propagation of this SD edge is a function of the line parameters and the SD signal parameter alpha and is controllable by changing alpha. The truncated SD leading edge of the test pulse will be attenuated as it travels along a lossy line. However, this rate of attenuation is also a function of the line parameters and the SD signal parameter alpha and is controllable by changing alpha.

The same principles used for TDR can be applied to waveforms other than electromagnetic waves in a transmission line. The reflections of acoustic waves in SONAR or geophysical applications can be analyzed using the techniques discussed in this section to provide accurate time of flight and distance estimates as well as be used to characterize the transmission media.

The accuracy of the SD test waveform in a TDR application can be demonstrated by an example using the T1 cable, as discussed in examples 1 and 2. The two twisted wire pairs inside the T1 cable are connected in series to form a 2004 ft cable and the truncated SD signal is applied to the input. The voltage trace is measured at the splice of the two lines at 1002 ft. This can eliminate the need to correct for any line impedance mismatch at the point of measurement. The exponential coefficient α of the applied wave is $6.7 \times 10^6 \text{ sec}^{-1}$. The SD parameters, A_{SD} and v_{SD} , are obtained from the empirical transfer function analysis presented in example 2. Interpolating from the data of table I gives an A_{SD} of $10.703 \times 10^{-3} \text{ ft}^{-1}$ and a v_{SD} of $6.276 \times 10^8 \text{ ft/sec}$ for this α .

The first demonstration terminates the cable with an open circuit. The reflected coefficient of an open circuit is +1, which results in a positive reflection. **FIG. 24** shows the

voltage waveform measured at the 1002 ft point showing the forward traveling applied wave and the wave reflected from an open termination at the end of the cable. **FIG. 25** shows the result of the measured reflected wave together with the applied wave time shifted and attenuated according to the procedure of example 4. The total distance traveled by the reflected wave is measured to be 2004 ft by this process, or exactly two times the 1002 ft length of this twisted wire pair section. **FIGS. 26A-B** show the constant threshold timing method result for TOF discussed in example 1. This method results in a time of flight of 3,203 nsec, which gives an estimated total distance traveled of 2010 ft. This results in an estimated twisted wire pair section length of 1005 ft, or 3 ft in error (+0.3%). This is an illustration of reduced accuracy using the constant threshold TOF measurements when the pulse travels far enough to experience significant attenuation. The accuracy of this method may decline more in cases where the test pulse experiences even more attenuation. This level of attenuation may not generally occur in digital circuits. The maximum attenuation of digital signals in interconnects is generally less than one-half of the driving signal. In such cases, the constant threshold TOF measurements with the SD wave should be preferably more accurate than current TDR methods using estimated time of emergence of the reflected pulses.

The next demonstration terminates the second twisted wire pair with a short circuit. The reflected coefficient at a short circuit is -1, which results in a negative reflection. **FIG. 27** shows the voltage waveform measured at the 1002 ft point showing the forward traveling applied wave and the wave reflected from a short circuit termination at the end of the second twisted wire pair. **FIG. 28** shows the result of the inverted reflected wave and the applied wave time shifted and attenuated according to the procedure of example 4. The total distance traveled is measured to be 2004 ft by this process, or exactly two times the 1002 ft twisted wire pair length. **FIGS. 29A-B** show the constant threshold timing method of estimating TOF discussed in example 1. This method results in a time of flight of 3,200 nsec, which gives an estimated total distance traveled of 2008 ft. This results in an estimated twisted wire pair section length of 1004 ft, or 2 ft in error (+0.2%).

The utility of the SD waveform to the TDR process can be increased by using an empirical transfer function obtained with the use of a non-SD pulse applied to the line in a way similar to the process discussed in example 2. As an example with this cable, an experimental voltage pulse is generated by an arbitrary function generator connected to a single unterminated

twisted wire pair from the T1 cable. This waveform is shown in **FIGS. 30-31** (dotted). The same waveform can be applied by the signal generator to a variable resistor load substituted for the 1002 ft twisted wire pair. The resistor is adjusted until the waveform approximates the shape and magnitude of the waveform applied by the arbitrary waveform generator to the twisted wire pair, see **FIGS. 30-31** (solid). The waveform match is not exact because the resistor does not have the exact same frequency response as that of the twisted wire pair. A load resistance equal to the characteristic impedance, Z_o , of the twisted wire pair results in a waveform that is closely matched to the waveform applied to the cable and has similar frequency content. The difference between the signal measured with the resistive load and the signal measured with the transmission line consists of the response of the transmission line to the applied signal and contains any reflected waves from impedance mismatches on the transmission line. The transfer function is then calculated from the ratio of the fast Fourier transforms (FFT) of the difference of the two signals, and the signal of the resistor alone:

$$H_{est}(j\omega) = \frac{\text{FFT}\left(V_{\text{Measured with Cable}} - V_{\text{Measured with Load Resistance } Z_o}\right)}{\text{FFT}\left(V_{\text{Measured with Load Resistance } Z_o}\right)}. \quad \text{This}$$

function, shown in **FIG. 32**, can be used to evaluate the reflected transmission line response to a SD pulse.

As a test of this process, a SD pulse with an α equal to the α used in the previous two experiments was simulated. The transfer function determined for the open terminated T1 cable, (**FIG. 32**), was used to calculate the expected response, see **FIG. 33**. There is a small pulse at the time of the applied pulse due to the resistor and cable having different high frequency impedance values. The influence of this pulse was eliminated with appropriate time windowing. The simulated response was then used to estimate the distance to the open termination using the methods of example 4 (**FIG. 34**), and example 1 (**FIGS. 35A-B**). The result of the slide and attenuate method resulted in an estimated total round trip distance the pulse traveled of 2004, or double the exact distance of 1002 ft to the open. The constant threshold method resulted in time of flight of 3,196 nsec for a estimated distance traveled of 2006 ft, or two times an estimated distance of 1003 ft to the open (0.1% Error).

No actual SD signal was experimentally applied to the twisted wire pair to obtain these results. The test signal applied to the line (**FIG. 30**) was similar to the type obtained from a commercial TDR instrument for testing communication cables with lengths ranging up to 5-kft. The measurements resulting from this test line voltage were used to obtain an empirical line transfer function. The SD TOF and the line distance estimates were all obtained computationally by simulating the SD propagation along the line utilizing this empirical line transfer function estimated with non-SD signals.

Example 6

Communication by high speed data transmission using speedy delivery waveforms

The SD waveform has several properties that make it suitable as the basis for a data transmission scheme. For example: the exponential shape is maintained as it propagates at constant speed in a uniform cable; the attenuation of the SD waveform is adjustable by changing the exponential waveform parameter α ; the propagating speed of the SD waveform is adjustable by changing the exponential waveform parameter α ; and the SD waveforms with different α 's are linearly independent.

This example teaches using the first property as the basis for a method of transmitting data using SD waveforms. The signals are transmitted 2004 ft along a transmission line consisting of the two twisted wire pairs in a 1002 ft T1 cable connected in series. The second twisted wire pair is terminated with a 100-ohm resistor. The data is encoded on four SD waveforms, each having a distinct α . Each waveform can be transmitted with a positive or a negative D, resulting in eight total states for three bits per symbol. One symbol is transmitted every three microseconds, giving a data rate of 1 Mbps. As one of ordinary skill in the art will recognize with the benefit of this disclosure, several other communication schemes can be readily derived from the teachings contained herein.

The symbols can consist of a two-microsecond truncated SD leading edge period followed by a one-microsecond recovery period. This recovery period includes a variable width half-sine compensating pulse adjusted to shorten the time required for the transmission line to return to a zero voltage. The need for this recovery period and compensating pulse is illustrated in **FIGS. 36A-B** and **FIGS. 37A-B**. **FIG. 36A** shows the transmitted pulse without the

compensating pulse, and **FIG. 36B** shows the received pulse without the compensating pulse. The truncated SD waveform in the leading edge of the input pulse is closed with a simple ramp signal, which is rounded due to the line response. Although the SD portion of the propagating pulse does not undergo deformation, the closing ramp does disperse into the region beyond the
 5 three-microsecond symbol period. **FIGS. 37A-B** show the effect of the compensating pulse in eliminating this inter-symbol interference. Each truncated SD waveform may require a different amount of compensation so the width of the half-sine pulse can be varied based on observing the pulse at the terminated end of the second twisted wire pair.

FIG. 38 shows a series of four symbols applied to the twisted wire pair input. Each
 10 symbol has a unique α , and a unique D. The α 's are equally spaced $1.1 \times 10^6 \text{ sec}^{-1}$ apart at $2.2 \times 10^6 \text{ sec}^{-1}$, $3.3 \times 10^6 \text{ sec}^{-1}$, $4.4 \times 10^6 \text{ sec}^{-1}$, and $5.5 \times 10^6 \text{ sec}^{-1}$. The order α 's used for this illustration of four successive symbols in **FIG. 9.3** is $4.4 \times 10^6 \text{ sec}^{-1}$, $2.2 \times 10^6 \text{ sec}^{-1}$, $3.3 \times 10^6 \text{ sec}^{-1}$, and $5.5 \times 10^6 \text{ sec}^{-1}$. The D's were chosen to maintain an approximately equal maximum and minimum symbol voltage. For this demonstration, the fourth waveform is inverted. The signal measured at the 100-
 15 ohm termination at a distance of 2004-ft is shown in **FIG. 39**. **FIG. 40** shows the natural log of the signal within each symbol period for the signal of **FIG. 39**. A set of matched filters could be used to detect the individual α 's. The slopes of the least square linear fits of the linear regions of these waveforms are shown in Table II (set forth below). The decision level for this demonstration would be one half of the step size in α ($1.1 \times 10^6 \text{ sec}^{-1}$) of $0.55 \times 10^6 \text{ sec}^{-1}$. All errors
 20 are well below this threshold.

Table II

Encoded and decoded α 's from data transmission demonstration

| Symbol | Applied α ($\times 10^6 \text{ sec}^{-1}$) | Detected α ($\times 10^6 \text{ sec}^{-1}$) | Error ($\times 10^6 \text{ sec}^{-1}$) |
|--------|--|---|---|
| 1 | 4.4 | 4.38 | -0.02 |
| 2 | 2.2 | 2.23 | +0.03 |
| 3 | 3.3 | 3.28 | -0.02 |
| 4 | 5.5 | 5.53 | +0.03 |

Example 7

Communication By High Speed Data Transmission Using Orthogonal Linear Combinations of Speedy Delivery Waveforms

This example discusses using the fourth SD waveform property, that SD waveforms with different α 's are linearly independent. This property can be utilized to generate an orthogonal set of pulses. Data can be encoded on these pulses by varying their amplitude. These amplitude-modulated orthogonal pulses are transmitted simultaneously on the transmission line. At the receiver, the individual pulse amplitudes are computed by taking advantage of their orthogonality. The basic pulses derived from the SD waveforms must be orthogonal at the receiver for this scheme to work.

The repetitive transmission of data may require closed pulses. As shown in example 6, that the SD portion of the leading edge of a closed propagating pulse maintains its shape, while the shape used to close the pulse will disperse as it propagates. The attenuation and propagation speed of the truncated SD leading region of the pulse is a function of the SD parameter α . Therefore, the orthogonality of a set of these pulses with a variety of α values will be reduced as the pulses propagate. Fortunately, the pulses only have to be orthogonal at the receiver. Techniques can be used to generate a set of orthogonal pulses from linearly independent component pulses. The effect of the transmission line on these component pulses can be determined empirically. These transmitted component waveforms, measured at the receiver, will be linearly independent. Linear combinations of these measured, linearly independent component pulses are used to generate a set of pulses that are orthogonal at the receiver. The constants determined by the orthogonalization process are supplied to the transmitter and used to generate transmitted pulses that will be orthogonal at the receiver. An example of this procedure follows. The process of generating and transmitting the orthogonal pulses is simulated for an 8 kft, 26 AWG twisted wire pair transmission line terminated with 100 Ohms resistance.

The example orthogonalization process can begin by using a set of SD waveforms $X_m(t) = e^{(\alpha_o + (m-1)\Delta\alpha)t}$ for $m = 1 \dots 5$, $\alpha_o = 1 \times 10^4 \frac{1}{\text{sec}}$, $\Delta\alpha = 1 \times 10^5 \frac{1}{\text{sec}}$, and $t = 0 \dots 7 \times 10^{-6}$ sec. The five waveforms are shown for a seven-microsecond interval in **FIG. 41**. The waveforms are linearly independent on this time interval. Additionally, these SD

waveforms have a polynomial structure, i.e. each waveform is equal to $e^{(\Delta\alpha)t}$ times the previous waveform. This allows the use of a simple recursive algorithm for polynomials [4] to create orthonormal linear combinations of these linearly independent waveforms. The set of five orthonormal Y's resulting from this procedure is shown in **FIG. 42**. In this case, X and Y are related by:

$$Y = \bar{A} \cdot X$$

$$\begin{bmatrix} Y_1 \\ Y_2 \\ Y_3 \\ Y_4 \\ Y_5 \end{bmatrix} = \begin{bmatrix} 0.026 & 0 & 0 & 0 & 0 \\ -0.129 & 0.088 & 0 & 0 & 0 \\ 0.717 & -1.003 & 0.337 & 0 & 0 \\ -4.074 & 8.578 & -5.874 & 1.310 & 0 \\ 23.304 & -65.519 & 67.862 & -30.704 & 5.124 \end{bmatrix} \cdot \begin{bmatrix} X_1 \\ X_2 \\ X_3 \\ X_4 \\ X_5 \end{bmatrix}.$$

The set, Y, is orthonormal in that the inner product $\langle Y_i, Y_j \rangle \equiv \int_0^{\Gamma=12\mu\text{sec}} Y_i(t) \cdot Y_j(t) dt = \begin{cases} 0 & i \neq j \\ 1 & i = j \end{cases}$.

The sharp waveform transitions at zero and seven microseconds have high frequency content that would disperse during transmission and contribute to inter-symbol interference.

This effect is reduced by multiplying this orthonormal set by $\sin(\omega \cdot t)$, with $\omega = \frac{\pi}{7 \times 10^{-6}}$. This

has the effect of changing the SD waveform to X', where $X'_m(t) = e^{(\alpha_o + (m-1)\Delta\alpha) \cdot t} \cdot \sin(\omega t) = \frac{1}{2j} \left[e^{[(\alpha_o + (m-1)\Delta\alpha) + j\omega]t} - e^{[(\alpha_o + (m-1)\Delta\alpha) - j\omega]t} \right]$ for $m = 1 \dots 5$.

Although the SD waveforms, X', are still linearly independent, the previous linear combination of these waveforms, $Y' = \bar{A} \cdot X'$, is no longer orthogonal. The waveforms, Y', are shown in **FIG. 43**. These waveforms may be transmitted through the channel to the receiver.

At the receiver, the waveforms show the trailing effects of dispersion beyond the 12 μ sec symbol period, see **FIG. 44**. If the spread of the pulses into the following symbol periods is not removed, this inter-symbol interference at the receiver will render data encoded at the transmitter unreadable. A simple method of eliminating this interference is by adding a pulse with a

compensating tail, $\mathbf{Z} = \mathbf{Y}' + \mathbf{b} \cdot \mathbf{C}$. In this example, the compensation pulse, $C(t)$, is one cycle of

$1 - \cos\left(\frac{\pi}{2.9 \times 10^{-6}} \cdot t\right)$ delayed one microsecond after the start of the symbol period, see **FIG. 45**.

FIGS. 46, 47 show the waveforms at the input and at 8 kft with appropriately sized compensating pulses added to each signal. The duration of these compensated component pulses at the receiver are now shorter than the desired 12 μ sec symbol period (see **FIG. 47**), so any linear combination of these pulses will not interfere with the next symbol.

The multiplication by the sine pulse, the adding of the compensating pulse, and the transmission in the channel, have all reduced the orthogonality of the initial five pulses. Initially the angles between the pulses in **FIG. 42** were all 90° . Now the angles between the pulses in **FIG. 47** range from 11° to 74° . These five linearly independent waveforms are now combined into set, S , of four orthonormal waveforms using principle component decomposition (PCD) [5]. The Gram-Schmidt (G-S) orthogonalization method [6] may also be used for this step in the procedure, however the PCD method is preferable. The G-S method expands an orthonormal vector set one vector at a time. This can lead to a poorly conditioned transformation matrix if the components are close to being linearly dependent. The PCD method is preferable because it takes into account the entire inner-product space spanned by all of the components. The process makes use of the eigenvectors of the entire inner-product space to generate the transformation matrix. Additionally, the transformation matrix can be easily manipulated to order the eigenvalues. In this example the four largest eigenvalues were selected by truncating the transformation matrix. This process yields a transformation matrix, $\bar{\mathbf{B}}$, that generates four waveforms that are orthonormal at 8-kft by linearly combining the five component signals. These waveforms are shown in **FIG. 48** (the transmitted signals) and **FIG. 49** (the received orthogonal signals).

$$\mathbf{S} = \overline{\mathbf{B}} \cdot \mathbf{Z}$$

$$\mathbf{S} = \overline{\mathbf{B}} \cdot (\mathbf{Y}' + \mathbf{b} \cdot \mathbf{C})$$

$$\begin{bmatrix} \text{S1} \\ \text{S2} \\ \text{S3} \\ \text{S4} \end{bmatrix} = \begin{bmatrix} 4.69 & 5.75 & -21.05 & -41.44 & 46.26 \\ 3.55 & 0.41 & 32.53 & -33.72 & 6.00 \\ -2.74 & 9.11 & 0.48 & -63.36 & -91.60 \\ -3.35 & -3.35 & 2.06 & -56.60 & 123.38 \end{bmatrix} \cdot \left(\begin{bmatrix} \text{Y}'_1 \\ \text{Y}'_2 \\ \text{Y}'_3 \\ \text{Y}'_4 \\ \text{Y}'_5 \end{bmatrix} + \begin{bmatrix} -0.04034 \\ -0.00283 \\ 0.01971 \\ -0.00007 \\ -0.00110 \end{bmatrix} \cdot \mathbf{C} \right)$$

$$\begin{bmatrix} \text{S1} \\ \text{S2} \\ \text{S3} \\ \text{S4} \end{bmatrix} = \begin{bmatrix} 1,231.1 & -3,365.0 & 3,375.8 & -1,474.7 & 237.0 \\ 300.5 & -714.9 & 616.1 & -228.4 & 30.7 \\ -1,877.3 & 5,458.0 & -5,843.5 & 2,729.4 & -469.3 \\ 3,107.7 & -8,571.5 & 8,705.9 & -3,862.4 & 632.2 \end{bmatrix} \cdot \begin{bmatrix} e^{1 \times 10^4 t} \\ e^{1.1 \times 10^5 t} \\ e^{2.1 \times 10^5 t} \\ e^{3.1 \times 10^5 t} \\ e^{4.1 \times 10^5 t} \end{bmatrix} \cdot \sin(\omega t) + \begin{bmatrix} -0.668 \\ 0.493 \\ 0.199 \\ 0.054 \end{bmatrix} \cdot \mathbf{C}$$

The orthonormal set S is orthogonal and normalized at the receiver. The process accounts for the effects of the channel and allows the transmitter input waveforms to be easily constructed as a linear combination of defined components prior to transmission. This set S can be used as the basis functions for an orthogonal pulse amplitude modulation data transmission scheme. This process is depicted in FIG. 50.

The data can be encoded by amplitude modulation with five bits on each of the four orthonormal pulses, S. Five bits requires thirty-two states. Each state corresponds to one amplitude level, a_i , on each orthogonal pulse, S_i . For this example the amplitude levels were $\pm 0.5, \pm 1.5, \dots, \pm 15.5$. These four modulated orthogonal pulses are summed to generate a symbol, $Q = a_1 S_1 + a_2 S_2 + a_3 S_3 + a_4 S_4$.

The symbol is transmitted and decoded at the receiver by using the orthogonality of the pulses. For example, the applied signal on orthogonal pulse one can be found by taking the inner product of the symbol, Q, with the pulse one signal, S_1 . This is done by multiplying the received symbol by the known orthogonal pulse, integrating over the symbol period, and normalizing the result.

$$\begin{aligned}
 \frac{\langle Q, S_1 \rangle}{\langle S_1, S_1 \rangle} &= \frac{\langle a_1 S_1 + a_2 S_2 + a_3 S_3 + a_4 S_4, S_1 \rangle}{\langle S_1, S_1 \rangle} \\
 &= \frac{\langle a_1 S_1, S_1 \rangle + \langle a_2 S_2, S_1 \rangle + \langle a_3 S_3, S_1 \rangle + \langle a_4 S_4, S_1 \rangle}{\langle S_1, S_1 \rangle} \\
 &= \frac{a_1 \langle S_1, S_1 \rangle + 0 + 0 + 0}{\langle S_1, S_1 \rangle} \\
 &= a_1
 \end{aligned}$$

FIG. 51 shows three successive symbols at the output of the transmitter and FIG. 52 shows the same three symbols at the input of the receiver. The symbols were encoded with a random selection of five bits per orthogonal pulse for a data rate of

$$5 \quad \frac{4 \frac{\text{Pulses}}{\text{Symbol}} \times 5 \frac{\text{Bits}}{\text{Pulse}}}{12 \times 10^{-6} \frac{\text{sec}}{\text{Symbol}}} = 1.67 \text{ Mbps} .$$

FIG. 53 is a histogram of the error in detected level, $a_{\text{expected}} - a_{\text{detected}}$, for one second, or 1.67×10^6 bits, of data transmitted with -140 dBm additive white gaussian noise (AWGN) present [7]. The standard deviations of the error distributions are 0.016, 0.018, 0.020, and 0.018 with maximum detected errors of 0.074, 0.071, 0.083, and 0.078. The decoding amplitude decision level is ± 0.5 .

Example 8 On-Chip Clock Circuits

This example teaches the use of SD waveforms in digital circuits, specifically, in on-chip clock circuits. On-chip clock circuits may behave like transmission lines with the increased clock rates.

A major design advantage of the SD waveforms when used as clock signals is that the delay in an RC line ($\sqrt{RC/\alpha} \cdot \ell$) is linear with length, ℓ , instead of quadratic with length as it is with the commonly assumed step signal. Linear delay with length ($\sqrt{LC} \sqrt{1 + \frac{R}{L\alpha}} \cdot \ell$) is also true for the SD signal in an RLC line. In contrast, the delay in RLC transmission lines excited by a

conventional signal exhibit an exponentially growing increase in delay with length [8].

The fact that the SD signal delay increases as a linear function of length for all types of clock lines can simplify the clock line layout design process. This linear relationship holds for the entire range of clock line behaviors, from the very lossy clock lines that behave like RC
5 transmission lines to low loss clock lines that behave like RLC transmission lines. This simple linear relationship between line length and SD clock signal delay provides a basis for implementing software CAD tools that may enhance accuracy while reducing design computational requirements.

Time Skew Clock control is a major issue limiting system performance. A solution to
10 this problem includes equalizing conductor lengths to the various locations on the chip where the clock signal is needed. Several geometrical patterns are commonly used (H, etc.) to equalize the signal path lengths, thereby equalizing signal delays. This requires equalizing the path lengths along the traces delivering the clock signal to pin locations that are physically close to the master chip clock driver with those clock signal path lengths to pin locations at the longest distances
15 from the master driver. Small adjustments to equalize the delays may be accomplished by adjusting the line widths to modify the line parameters and thereby the delay per unit length and alternatively by active means such as varying the delay of delayed-locked loops embedded in the clock lines.

The invention provides a method and/or apparatus for utilizing truncated SD clock
20 waveforms to reduce the clock skew. An advantage of the invention is that the delay per unit length can be adjusted with the exponential coefficient, α , chosen in the clock line driver design.

This controllable delay per unit length has been demonstrated for a 100-ft coaxial cable in the initial patent application [9]. The minimum coaxial cable delay measured was increased by a factor of 1.8.

25 The large increases in the clock signal delay per unit length achieved by varying the exponential coefficient α allows lines to the pins close to the clock driver to be shortened to more closely approximate the minimum physical distance between the driver and pin while maintaining a path delay equal to the delay of the longest clock lines on the chip. The shorter clock distribution lines may utilize less physical space on the chip and reduce the total clock line capacitance. This reduction in clock line capacitance reduces the total power consumption
30 required by the chip for a given clock frequency. In one embodiment, an SD waveform driver

may also be made to create an SD clock signal with adjustable delay by including a mechanism for adjusting the exponential coefficient α in the clock generator. This adjustable delay of the SD clock signal may provide a larger delay adjustment range than the delayed-locked loops utilized in the current technology.

5 The invention can include transmission media. The invention can include a method and/or apparatus for a time domain reflectometer. Further, the invention can include a method and/or apparatus for high-speed communication over a transmission line. The invention can also include a method and/or apparatus for high-speed clock signals in an integrated circuit.

10 The terms a or an, as used herein, are defined as one or more than one. The term plurality, as used herein, is defined as two or more than two. The term another, as used herein, is defined as at least a second or more. The terms including and/or having, as used herein, are defined as comprising (i.e., open language). The term coupled, as used herein, is defined as connected, although not necessarily directly, and not necessarily mechanically. The term approximately, as used herein, is defined as at least close to a given value (e.g., preferably within 10% of, more
15 preferably within 1% of, and most preferably within 0.1% of). The term substantially, as used herein, is defined as at least approaching a given state (e.g., preferably within 10% of, more preferably within 1% of, and most preferably within 0.1% of). The term means, as used herein, is defined as hardware, firmware and/or software for achieving a result. The term program or
20 phrase computer program, as used herein, is defined as a sequence of instructions designed for execution on a computer system. A program, or computer program, may include a subroutine, a function, a procedure, an object method, an object implementation, an executable application, an applet, a servlet, a source code, an object code, a shared library/dynamic load library and/or other sequence of instructions designed for execution on a computer system.

25 All the disclosed embodiments of the invention disclosed herein can be made and used without undue experimentation in light of the disclosure. The invention is not limited by theoretical statements recited herein. Although the best mode of carrying out the invention contemplated by the inventors is disclosed, practice of the invention is not limited thereto. Accordingly, it will be appreciated by those skilled in the art that the invention may be practiced otherwise than as specifically described herein.

30 Further, variation may be made in the steps or in the sequence of steps composing methods described herein.

The appended claims are not to be interpreted as including means-plus-function limitations, unless such a limitation is explicitly recited in a given claim using the phrase(s) "means for" and/or "step for." Subgeneric embodiments of the invention are delineated by the appended independent claims and their equivalents. Specific embodiments of the invention are
5 differentiated by the appended dependent claims and their equivalents.

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20

CLAIMS

1. A method for transmitting a waveform having an essentially constant propagation velocity along a transmission line, comprising:

generating an exponential waveform, the exponential waveform (a) being characterized
5 by the equation $V_{in} = De^{-A_{SD}[x-v_{SD}t]}$, where D is a magnitude, V_{in} is a voltage, t is time, A_{SD} is an attenuation coefficient, and v_{SD} is a propagation velocity; and (b) being truncated at a maximum value.

2. The method of claim 1, further comprising determining the attenuation coefficient and the
10 propagation velocity, comprising:

applying the exponential waveform to a test length of a transmission line;

measuring a time-of-flight;

utilizing the time-of-flight measurement and the test length of the transmission line to
determine the propagation velocity; and

15 utilizing an exponential coefficient α and the propagation velocity to determine the
attenuation coefficient.

3. The method of claim 2, further comprising utilizing a constant threshold time of flight
measurement to measure a distance to a discontinuity in another transmission line.

4. The method of claim 1, further comprising determining an attenuation coefficient and a
propagation velocity for an exponential coefficient α , including:

determining a transmission line transfer function;

simulating an input waveform characterized by an exponential coefficient α ;

25 utilizing the transmission line transfer function to calculate an output waveform;

calculating a simulated time-of-flight duration;

utilizing the time-of-flight duration and a length of the transmission line to determine the
propagation velocity; and

30 utilizing the exponential coefficient α and the propagation velocity to determine the
attenuation coefficient.

5. The method of claim 4, further comprising simulating a set of input waveforms characterized by a range of exponential coefficients α .

5 6. The method of claim 5, further comprising determining a range of propagation velocities and attenuation coefficients corresponding to the range of exponential coefficients α .

7. The method of claim 4, wherein determining a transmission line transfer function includes applying an input signal and measuring an output signal at a distance along the
10 transmission line.

8. The method of claim 1, further comprising determining an impedance of the transmission line, including:

15 applying the exponential waveform to the transmission line;
measuring a signal at a termination;
calculating a ratio between a magnitude of a waveform at the termination and a
magnitude of an incident waveform;
utilizing the ratio to determine a reflection coefficient; and
utilizing the reflection coefficient to determine the impedance of the transmission line.

20

9. The method of claim 8, further comprising determining the magnitude of the waveform at the termination by applying the exponential waveform to the transmission line of open circuit termination.

25 10. The method of claim 8, further comprising controlling the impedance of the transmission line by selecting an exponential coefficient α .

11. The method of claim 1, further comprising determining a transmission line length, including:

30 applying the exponential waveform to the transmission line;
estimating a time-of-flight of a leading edge of the exponential waveform at a length of

the transmission line;
attenuating and time-shifting the exponential waveform based on the estimated time-of-flight; and
determining a difference between an attenuated and time-shifted waveform and a
5 measured waveform at the length of the transmission line.

12. The method of claim 1, further comprising measuring a distance to a transmission line impedance change, including:

10 applying an arbitrary pulse to the transmission line and measuring a first reflected pulse;
applying the arbitrary pulse to a variable resistor load and measuring a resulting pulse;
varying the resistance of the variable resistor until the resulting pulse approximates the
first reflected pulse;

calculating a reflected transmission line transfer function utilizing a relation between the
first reflected pulse and the resulting pulse from the variable resistor load; and

15 utilizing the transfer function to determine the distance to the transmission line
impedance change.

13. The method of claim 12, wherein the impedance change corresponds to a fault.

20 14. The method of claim 12, further comprising determining at least one transmission line
parameter from the group consisting of: an inductance per unit length, a capacitance per unit
length, a resistance per unit length, and a conductance per unit length.

25 15. The method of claim 1, further comprising utilizing the exponential waveform to transmit
data via the transmission line.

16. The method of claim 15, further comprising increasing a number of bits transmitted per
pulse by varying an exponential coefficient α to encode the data.

30 17. The method of claim 15, further comprising utilizing a set of exponential waveforms,
each with a distinct exponential coefficient α .

18. The method of claim 15, further comprising utilizing a set of exponential waveforms, each with a distinct magnitude.

5 19. The method of claim 15, further comprising generating a set of pulses consisting of linear combinations of exponential waveforms.

20. The method of claim 19, wherein the set of pulses is orthogonal at a receiver.

10 21. The method of claim 20, wherein data is encoded in each of the set of pulses by varying each pulse amplitude.

22. The method of claim 20, further comprising simultaneously transmitting the set of pulses to a receivers via the transmission line.

15

23. The method of claim 1, further comprising equalizing a clock propagation delay.

24. The method of claim 23, further comprising reducing a clock skew.

20 25. The method of claim 1, wherein an attenuation of a signal at a point, t_j , on the exponential waveform truncated at time t_i such that $0 \leq t_j \leq t_i$ propagating in the transmission line at a distance ℓ when measured in a reference frame t' traveling with a velocity $1/\sqrt{LC}$ is related to α

in accordance with the equation:
$$\frac{V(x = \ell, t' = t_j)}{V(x = 0, t = t_j)} = e^{-\bar{A}_{sd}(\alpha) \cdot [1 - \bar{v}_{sd}(\alpha) \cdot \sqrt{\bar{L}\bar{C}}] \cdot \ell};$$

wherein \bar{L} represents an inductance per unit length, \bar{C} represents a capacitance per unit

25 length of the transmission line, $\bar{A}_{sd} = \sqrt{\bar{L}\bar{C} \cdot \alpha^2 + (\bar{R}\bar{C} + \bar{L}\bar{G}) \cdot \alpha + \bar{R}\bar{G}}$, and

$$\bar{v}_{sd} = \frac{\alpha}{\sqrt{\bar{L}\bar{C} \cdot \alpha^2 + (\bar{R}\bar{C} + \bar{L}\bar{G}) \cdot \alpha + \bar{R}\bar{G}}}.$$

26. The method of claim 1, wherein an attenuation of a signal at a point, t_j , on the exponential waveform truncated at time t_i such that $0 \leq t_j \leq t_i$ propagating in a pair of coupled transmission lines at a distance ℓ when measured in a reference frame t' traveling with a velocity $1/\sqrt{LC}$ is related to α in accordance with the equation:

$$\begin{aligned}
 \frac{V_1(x = \ell, t' = t_j)}{V_1(x = 0, t = t_j)} = & \frac{\left(A_{11} \cdot e^{-A_{SD11} \left(1 - v_{SD11} \sqrt{LC}\right) \ell} + A_{21} \cdot e^{-A_{SD21} \left(1 - v_{SD21} \sqrt{LC}\right) \ell} \right)}{\left(A_{11} + A_{21} \right) + \left(A_{12} + A_{22} \right) \cdot e^{\left(\alpha_2 - \alpha_1 \right) t_i}} \\
 & + \frac{\left(A_{12} \cdot e^{-A_{SD12} \left(1 - v_{SD12} \sqrt{LC}\right) \ell} + A_{22} \cdot e^{-A_{SD22} \left(1 - v_{SD22} \sqrt{LC}\right) \ell} \right)}{\left(A_{11} + A_{21} \right) \cdot e^{\left(\alpha_1 - \alpha_2 \right) t_i} + \left(A_{12} + A_{22} \right)} ;
 \end{aligned}$$

wherein \bar{L} represents an inductance per unit length and \bar{C} represents a capacitance per unit length.

27. The method of claim 1, wherein an attenuation of a signal at a point, t_j , on a varying exponential waveform envelope $E(x,t)$ of an optical pulse truncated at time t_i such that $0 \leq t_j \leq t_i$ propagating in a single mode communication fiber at a distance ℓ when measured in a reference frame t' traveling with a group velocity $1/\beta_1$ is related to α in accordance with the

$$\text{equation: } \frac{|E(x = \ell, t' = t_j)|^2}{|E(x = 0, t = t_j)|^2} = e^{-\left(\frac{\gamma}{2} - \frac{\beta_3}{6T_o^3} \right) \ell} ;$$

wherein $T_o = \frac{1}{\alpha}$ and β_3 is a third-order dispersion parameter of the single mode optical fiber, and γ is a fiber attenuation parameter.

28. The method of claim 1, wherein the velocity of propagation of a leading edge of a varying exponential waveform envelope $E(x,t)$ of an optical pulse propagating in a single mode communication fiber at a distance ℓ is related to α in accordance with the equations:

$$v_{sd}(T_o) = \frac{1}{\left(\beta_1 - \frac{\beta_3}{6T_o^2} + \frac{T_o\gamma}{2} \right)};$$

wherein $T_o = \frac{1}{\alpha}$, $\frac{1}{\beta_1}$ is a group velocity, β_3 is a third-order dispersion parameter, and γ is a fiber attenuation parameter of the single mode optical fiber.

- 5 29. The method of claim 1, wherein an attenuation parameter of a varying exponential waveform envelope $E(x,t)$ of an optical pulse propagating in a single mode communication fiber at a distance l is related to α in accordance with the equations:

$$A_{sd}(T_o) = \left(\frac{\beta_1}{T_o} - \frac{\beta_3}{6T_o^3} + \frac{\gamma}{2} \right);$$

10

wherein $T_o = \frac{1}{\alpha}$, $\frac{1}{\beta_1}$ is a group velocity, β_3 is a third-order dispersion parameter, and γ is a fiber attenuation parameter of the single mode optical fiber.

- 15 30. The method of claim 1, wherein the impedance of a transmission line is a real number and is related to α in accordance with the equation:

$$Z_{SD}(\alpha) = \sqrt{\frac{R + \alpha \cdot L}{G + \alpha \cdot C}};$$

wherein L represents an inductance per unit length, C represents a capacitance per unit length, R represents a resistance per unit length, and G represents a conductance per unit length of the transmission line.

20

31. The method of claim 1, wherein the impedance of a transmission line is a real number and is related to α in accordance with the equation:

$$Z_{SD}(\alpha) = \sqrt{\frac{\overline{R}(\alpha) + \alpha \cdot \overline{L}(\alpha)}{\overline{G}(\alpha) + \alpha \cdot \overline{C}(\alpha)}};$$

wherein \bar{L} represents an inductance per unit length, \bar{C} represents a capacitance per unit length, \bar{R} represents a resistance per unit length, and \bar{G} represents a conductance per unit length of the transmission line, each varying with frequency.

- 5 32. A method for increasing the accuracy of a time domain reflectometer unit, comprising:
utilizing a standard time domain reflectometer waveform to determine a transfer function
for a transmission line under test; and
utilizing the transfer function to simulate a propagation of an exponential waveform.
- 10 33. A method for communicating data, comprising:
generating an exponential waveform at a first location;
varying an exponential coefficient α to encode a message;
applying the exponential waveform to a transmission line;
receiving the exponential waveform at a second location; and
15 decoding a received exponential waveform to obtain the message.
34. An apparatus for determining an attenuation coefficient and a propagation velocity,
comprising:
an exponential waveform generator;
20 an input recorder coupled to an output of the exponential waveform generator;
a transmission line under test coupled to the output of the exponential waveform
generator;
an output recorder coupled to the transmission line under test;
an additional transmission line coupled to the transmission line under test; and
25 a termination impedance coupled to the additional transmission line and to a ground.
35. The apparatus of claim 34, wherein the transmission line electrical parameters of the
additional transmission line is substantially identical to the transmission line electrical
parameters of the transmission line under test and long enough to delay reflected waves to a time
30 later than the end of the incident wave exponential region.

36. An apparatus for determining an impedance of a transmission line, comprising:
an exponential waveform generator;
an input recorder coupled to an output of the exponential waveform generator;
a transmission line coupled to the output of the exponential waveform generator;
5 an output recorder coupled to the transmission line; and
a termination impedance, coupled to the transmission line and to a ground.
37. An apparatus for performing a time domain reflectometry measurement of a transmission
line, comprising:
10 an exponential waveform generator;
a transmission media connector coupled to the exponential waveform generator;
an applied signal recorder coupled to the transmission media connector; and
a reflected signal recorder coupled to the transmission media connector.
- 15 38. The apparatus of claim 37, further comprising a computer coupled to the exponential
waveform generator.
39. The apparatus of claim 38, wherein the computer is coupled to the applied signal recorder
and to the reflected signal recorder.
20
40. A method of communicating data, comprising:
generating a set of truncated exponential waveforms;
combining the set of truncated exponential waveforms to form an orthogonal set at a first
location;
25 shaping a power spectral density of the orthogonal set to form a shaped orthogonal set;
and
sequentially transmitting the shaped orthogonal set to a second location.
41. The method of claim 40, further comprising:
30 transmitting a compensation pulse to the second location;
creating an orthogonal linear combination of the shaped orthogonal set and the

compensation pulse; and
transmitting a set of constants from the second location to the first location.

42. The method of claim 41, further comprising:

5 generating a set of pulses utilizing the set of constants by linearly combining the shaped
orthogonal set and the compensating pulse at the first location;
encoding data on the set of generated pulses by varying each pulse amplitude;
simultaneously transmitting the set of generated pulses to the second location; and
decoding the data by correlating the received simultaneous combination of transmitted
10 signals with an expected orthogonal set.

43. A method of detecting an exponential region and value of an exponential constant of an
exponential waveform comprising:

filtering a waveform to obtain $\frac{\partial^n V(x,t)}{\partial^n t}$ and $\frac{\partial^{n-1} V(x,t)}{\partial^{n-1} t}$ for $n = 0, 1, 2, \dots$;

15 taking the ratio of the filter outputs $\frac{\frac{\partial^n V(x,t)}{\partial^n t}}{\frac{\partial^{n-1} V(x,t)}{\partial^{n-1} t}}$;

locating an exponential region by determining a period when the ratio is constant; and
determining the exponential constant as the constant value of the ratio.

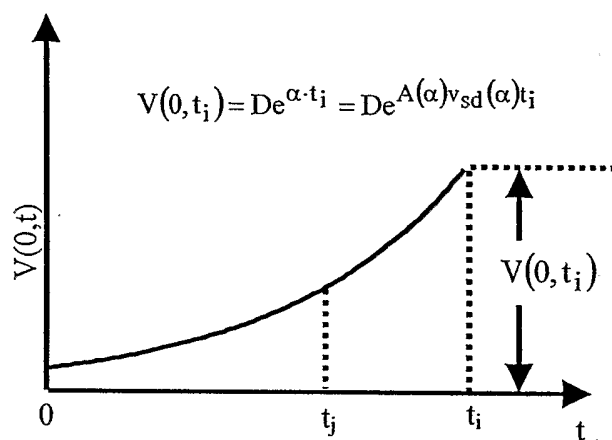


FIG. 1A

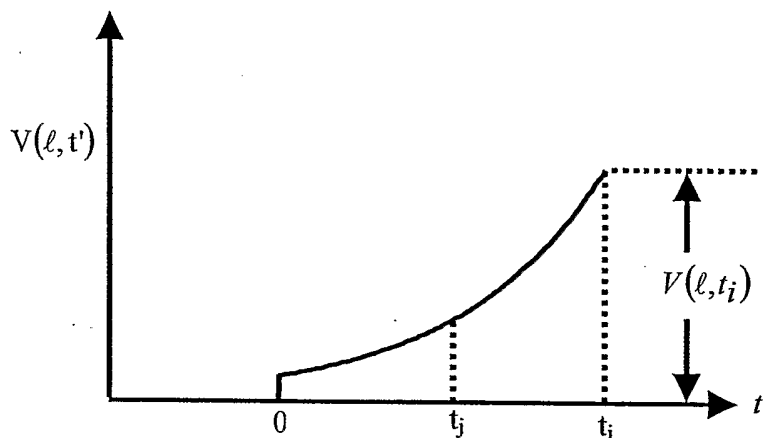


FIG. 1B

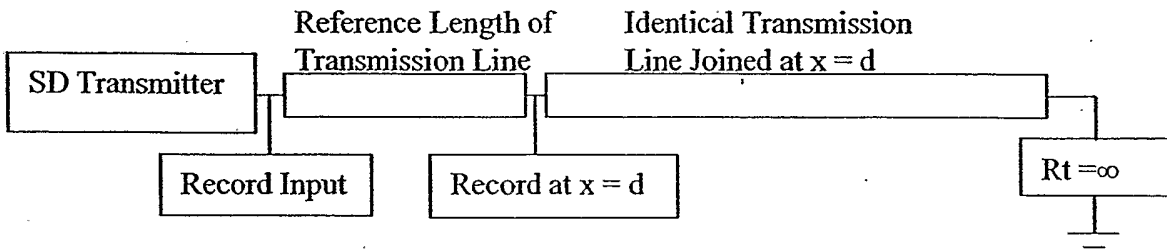


FIG. 2

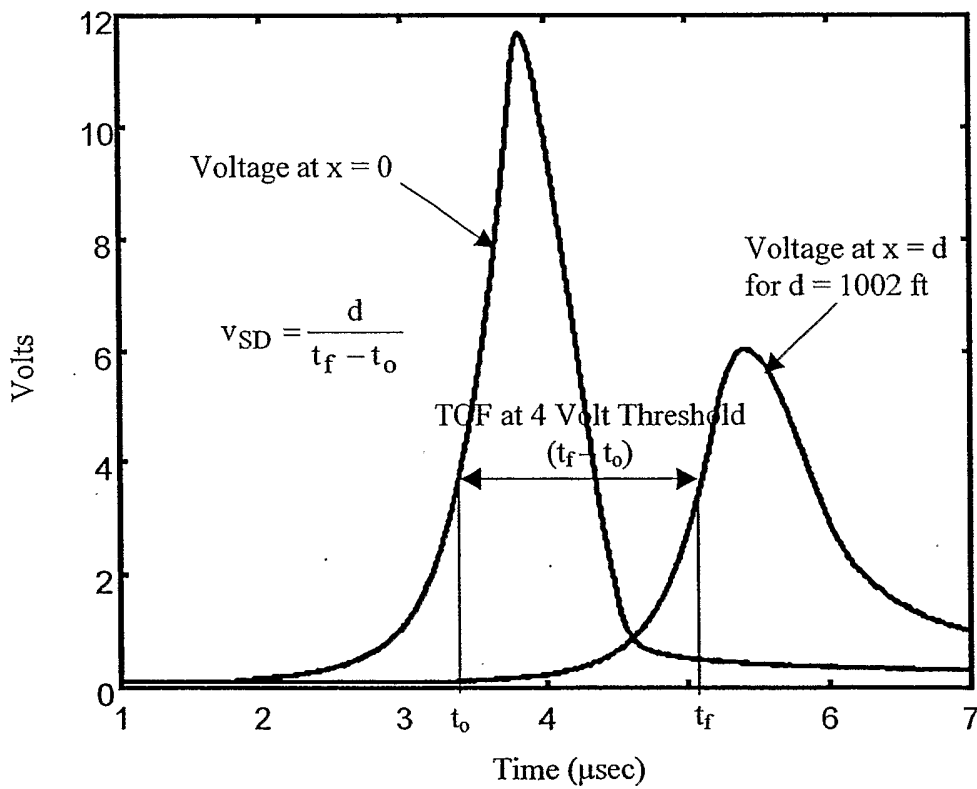


FIG. 3

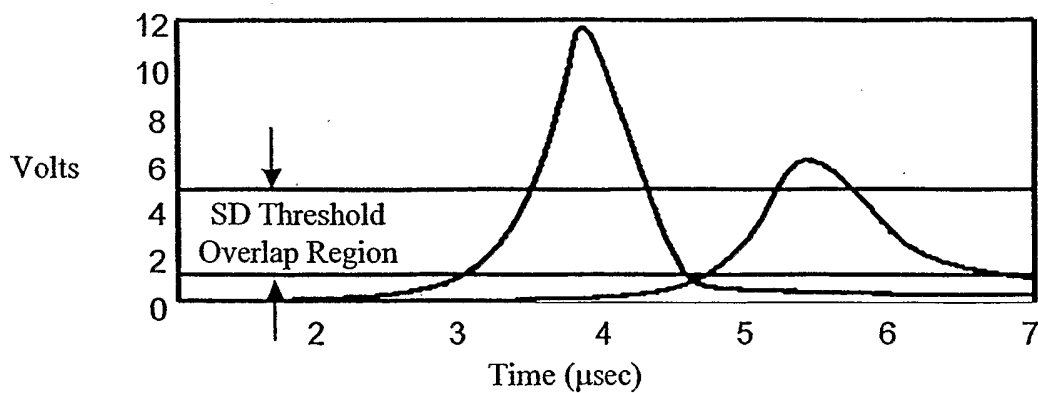


FIG. 4A

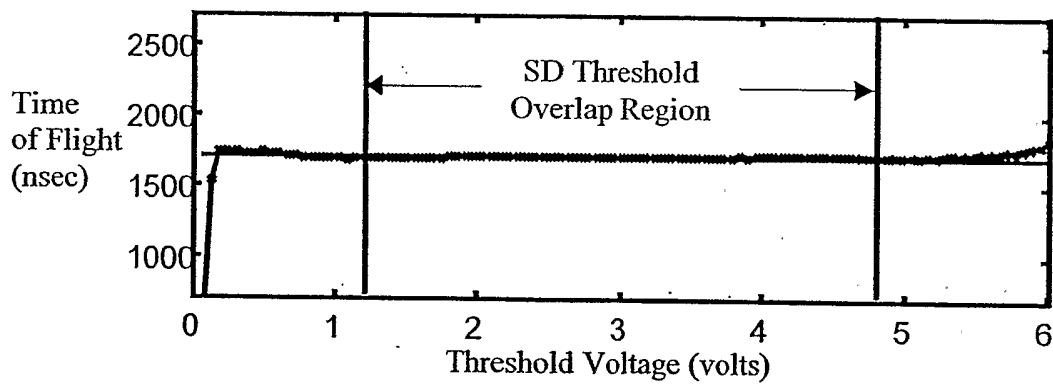


FIG. 4B

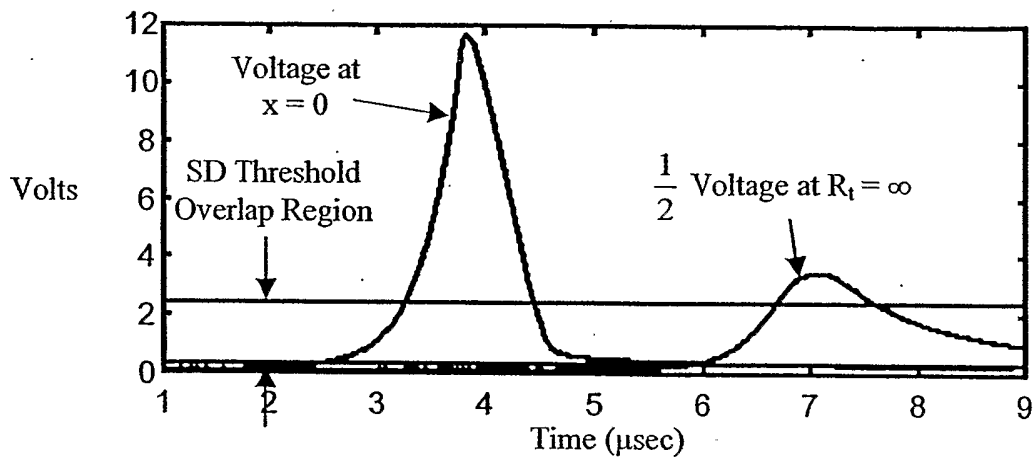


FIG. 5A

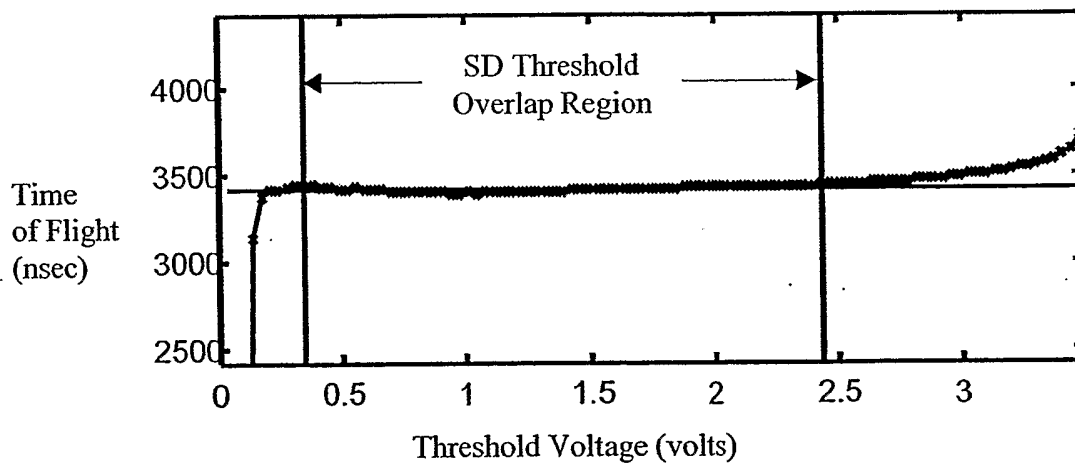


FIG. 5B

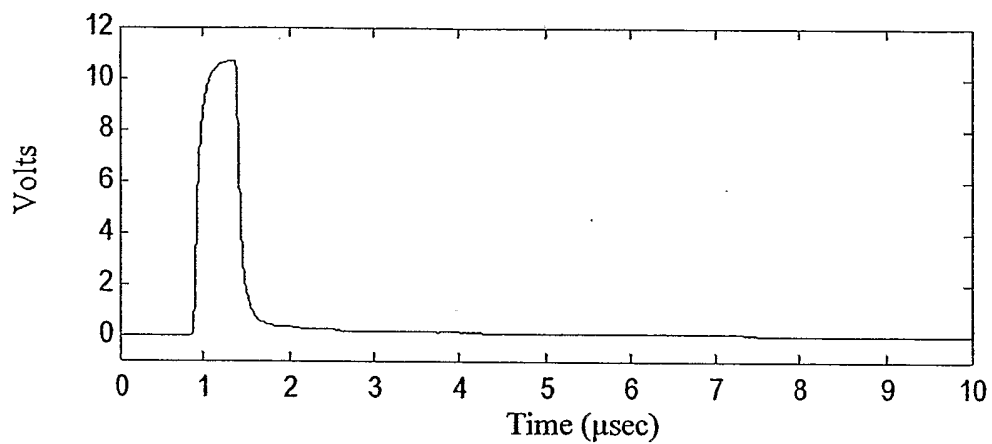


FIG. 6A

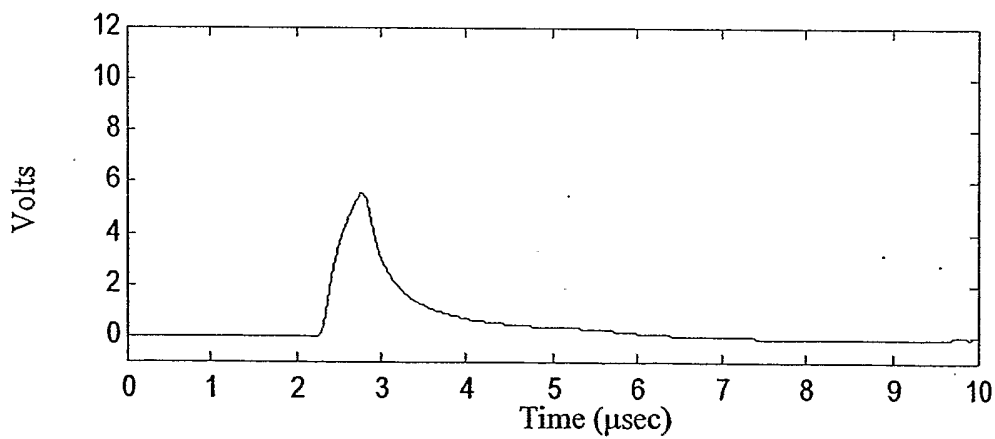


FIG. 6B

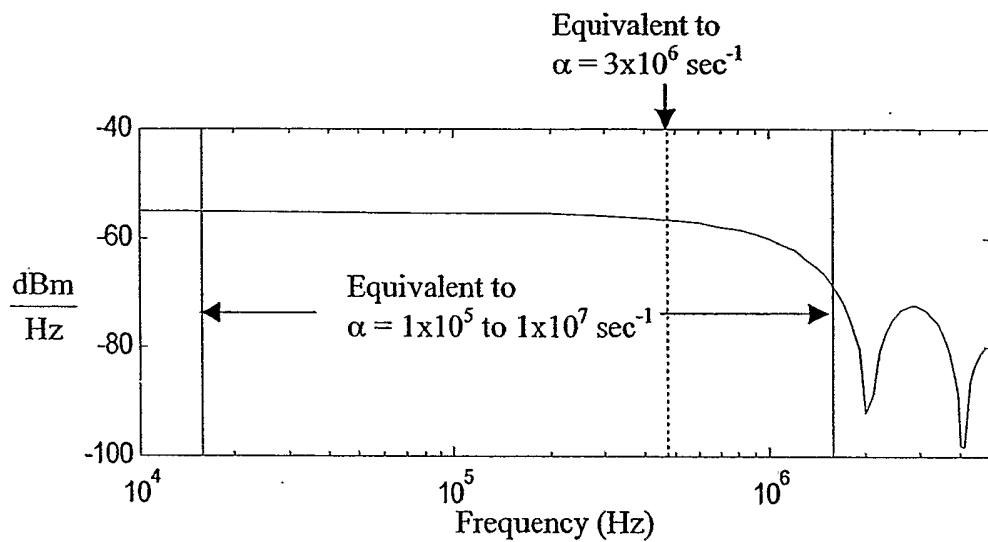


FIG. 7A

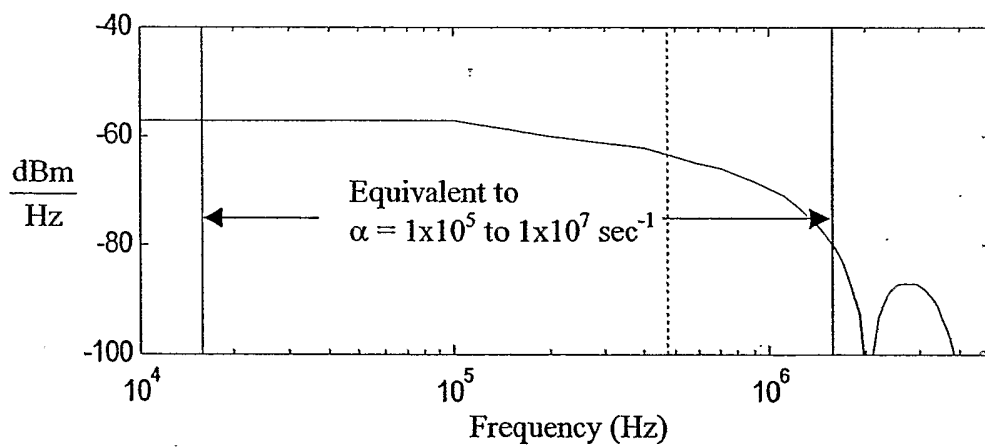


FIG. 7B

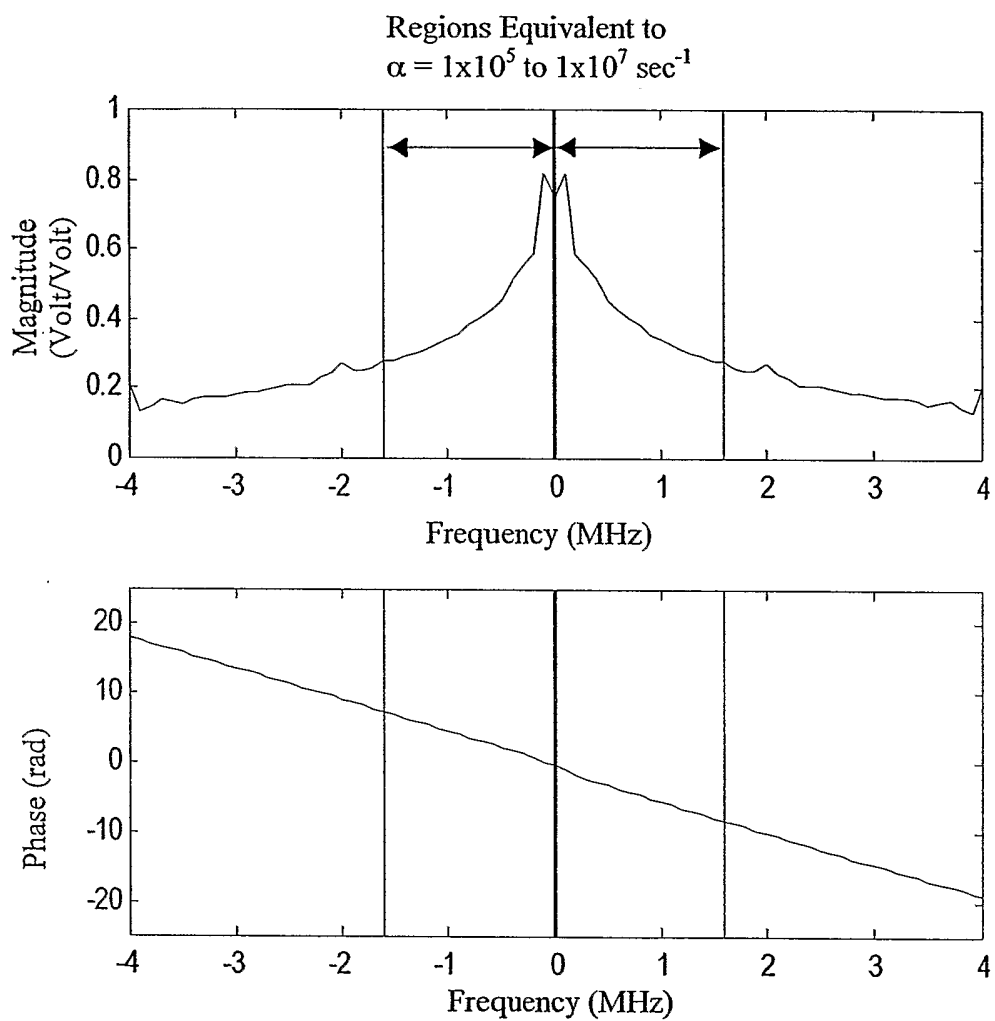


FIG. 8

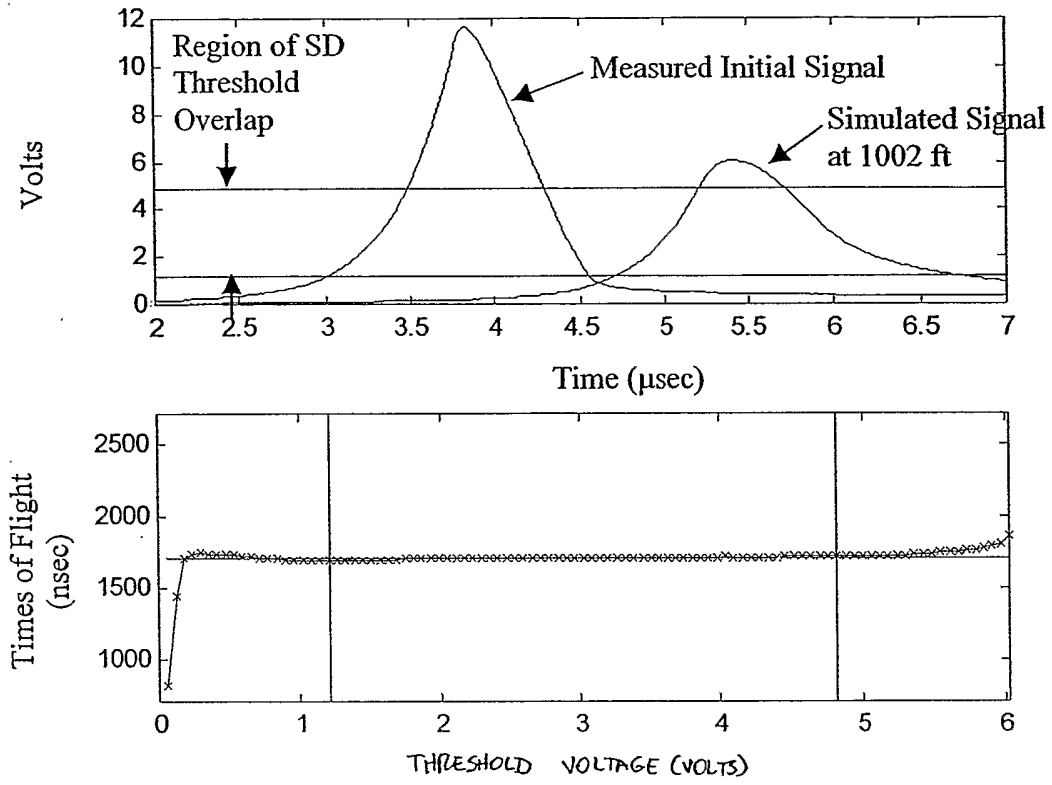


FIG. 9

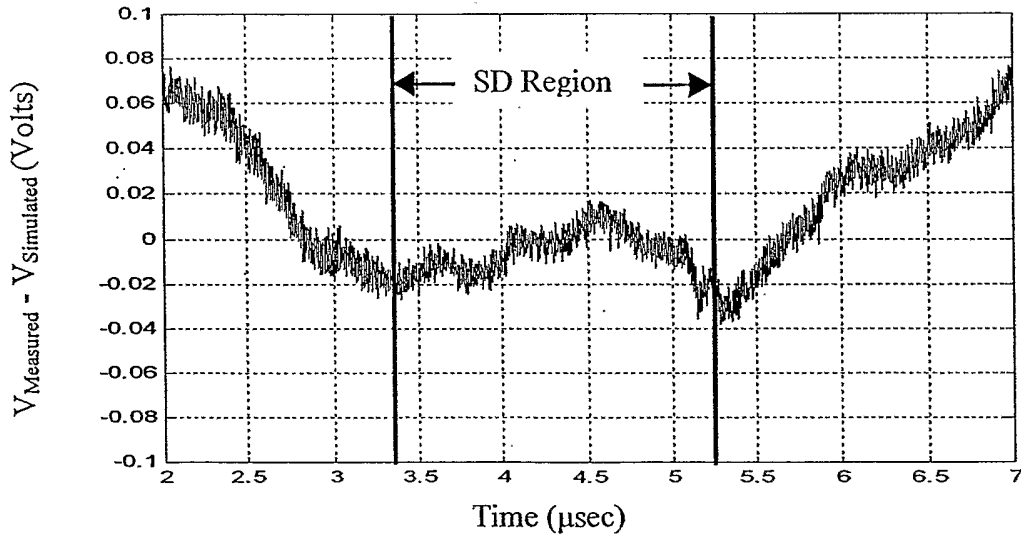


FIG. 10

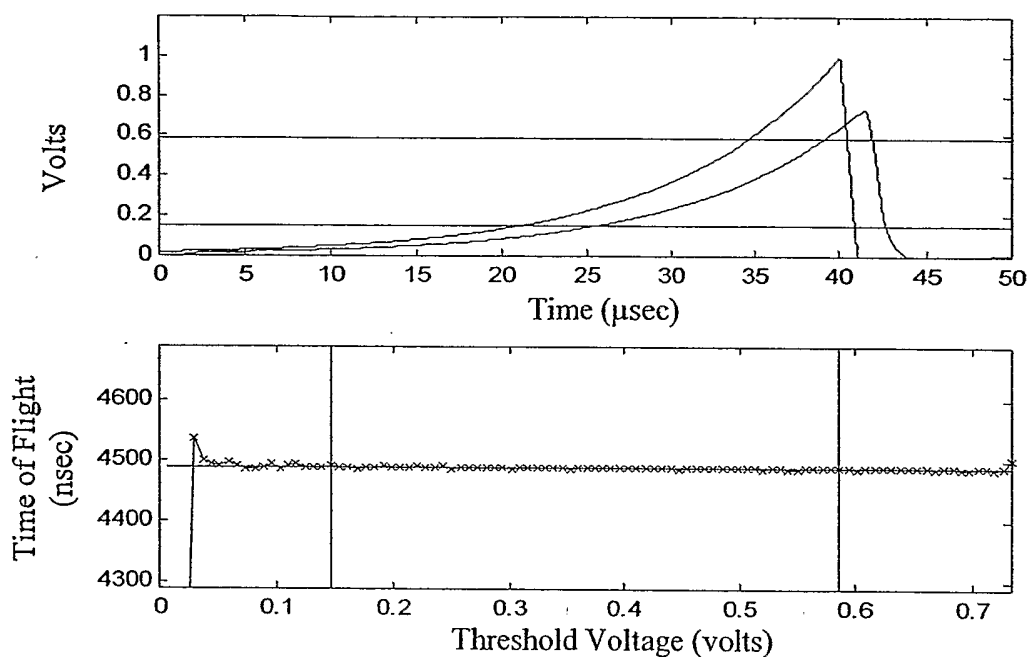


FIG. 11

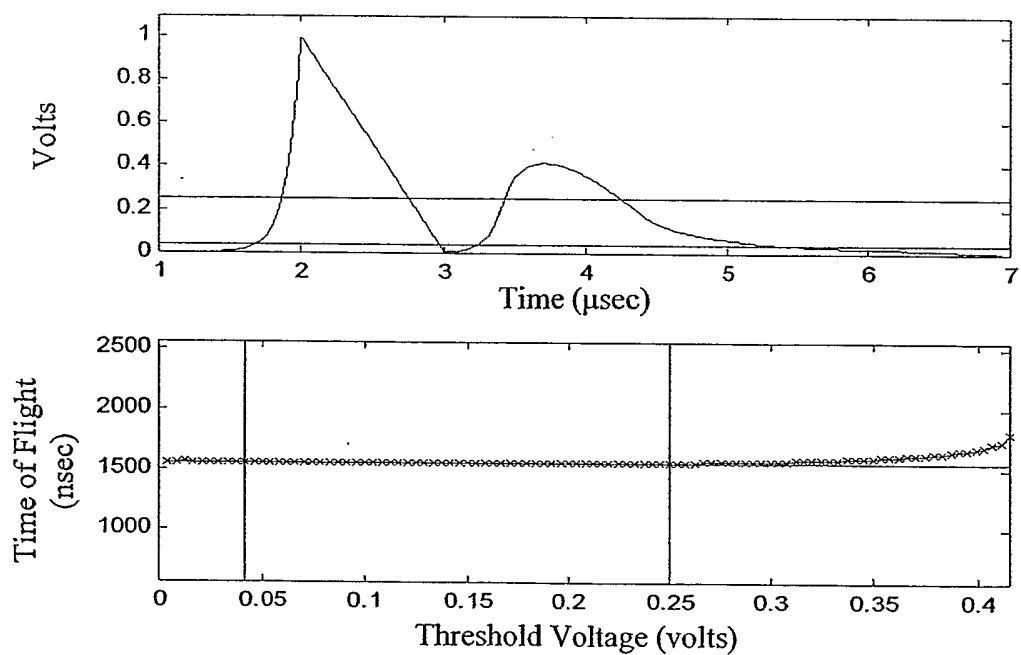


FIG. 12

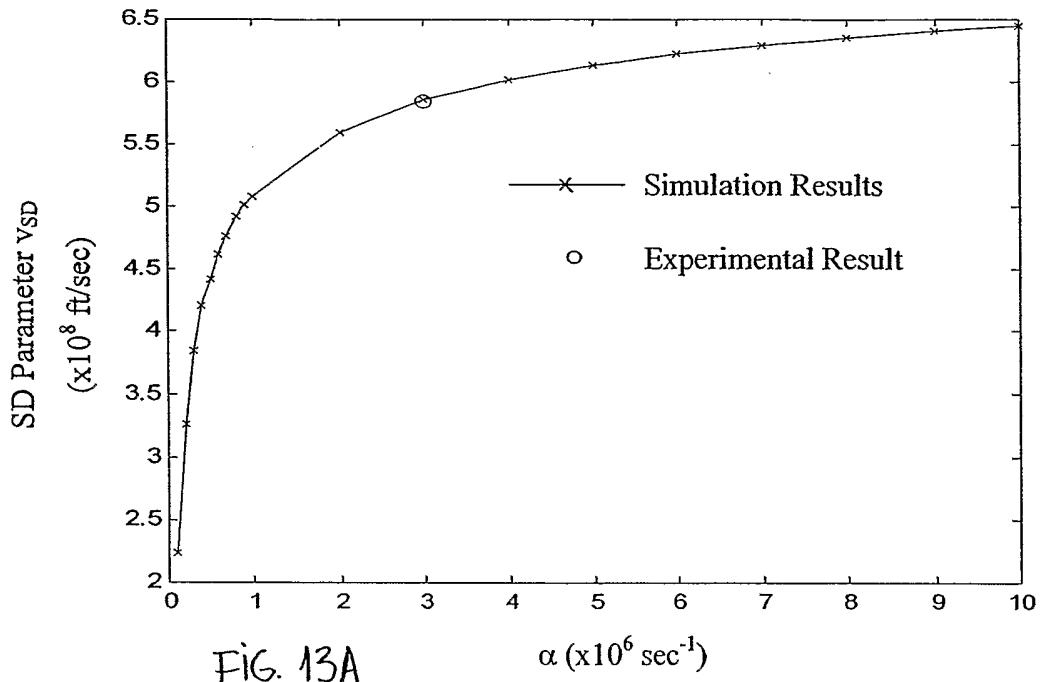


FIG. 13A

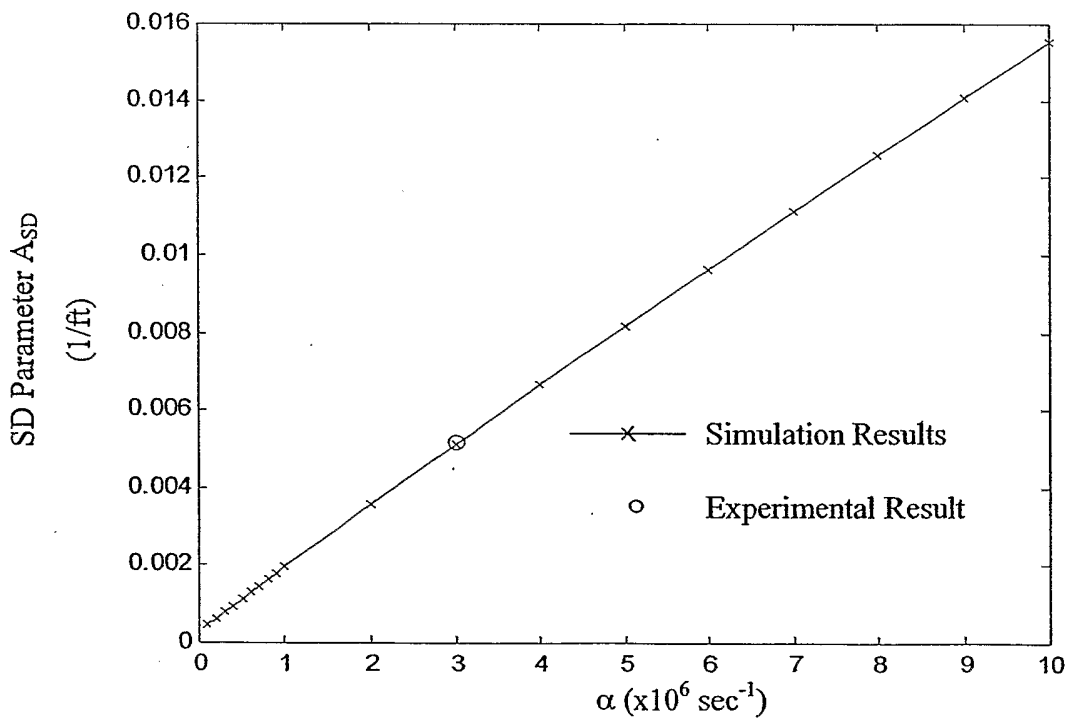


FIG. 13B

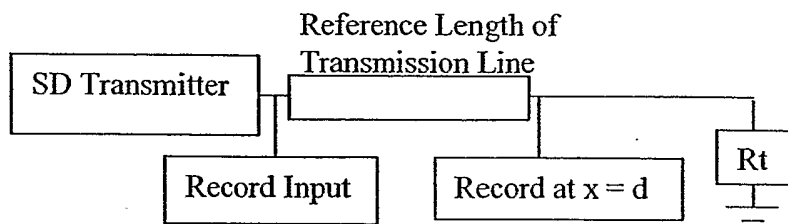


FIG. 14

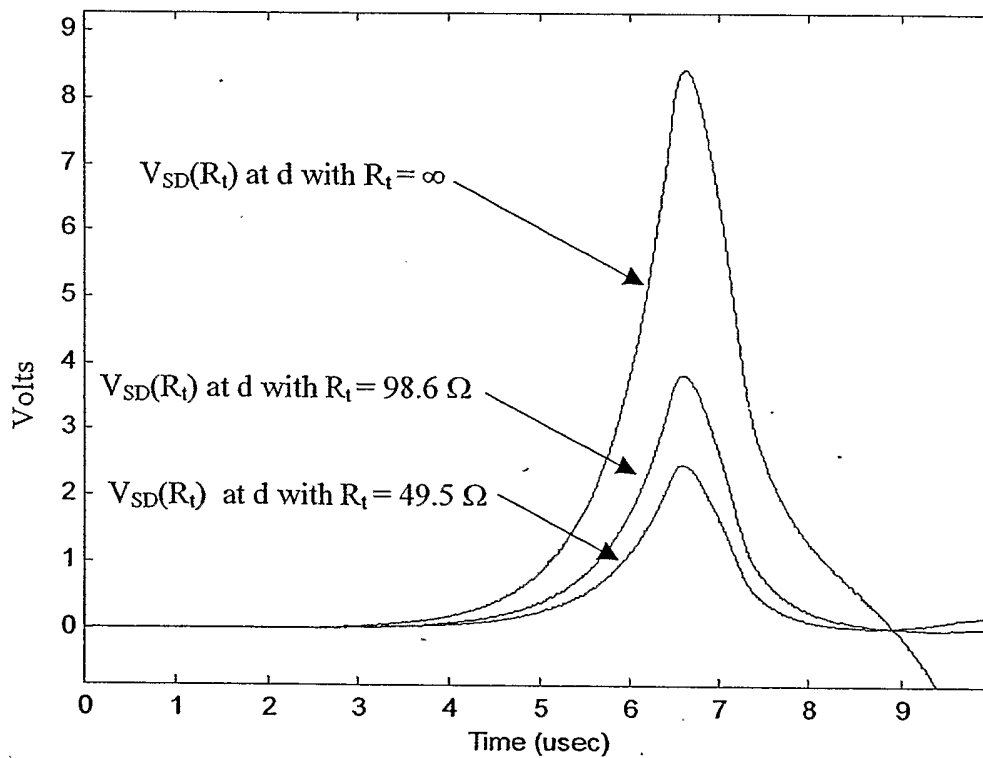


FIG. 15A

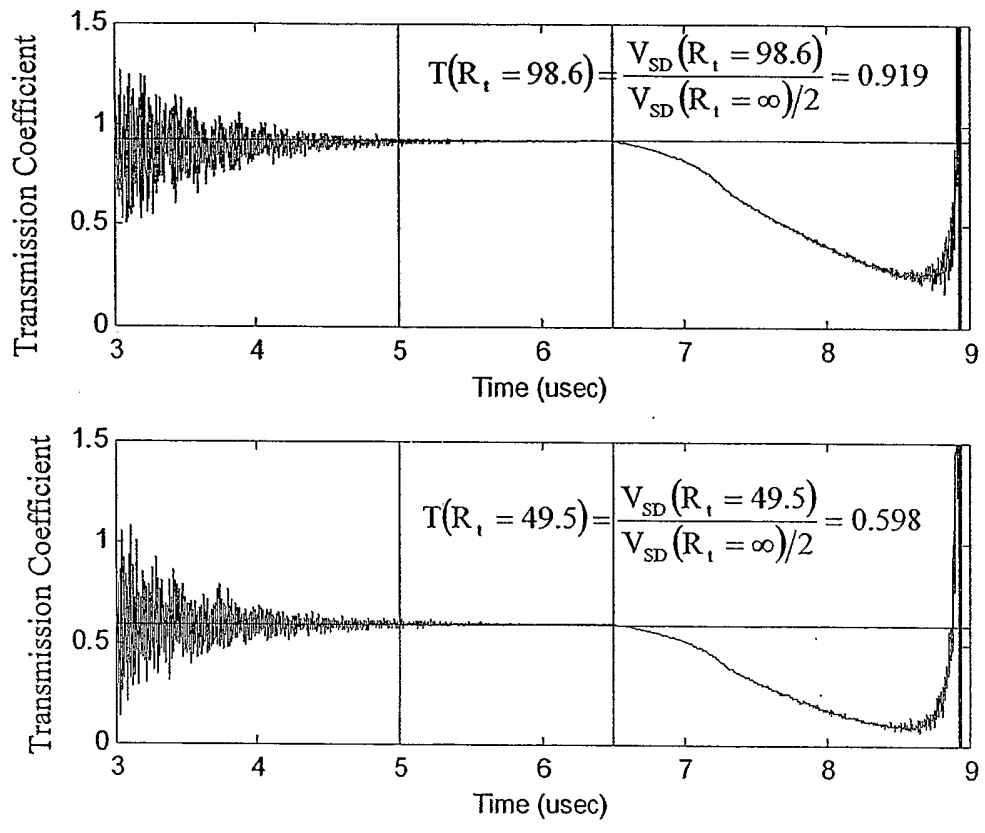


FIG. 15B

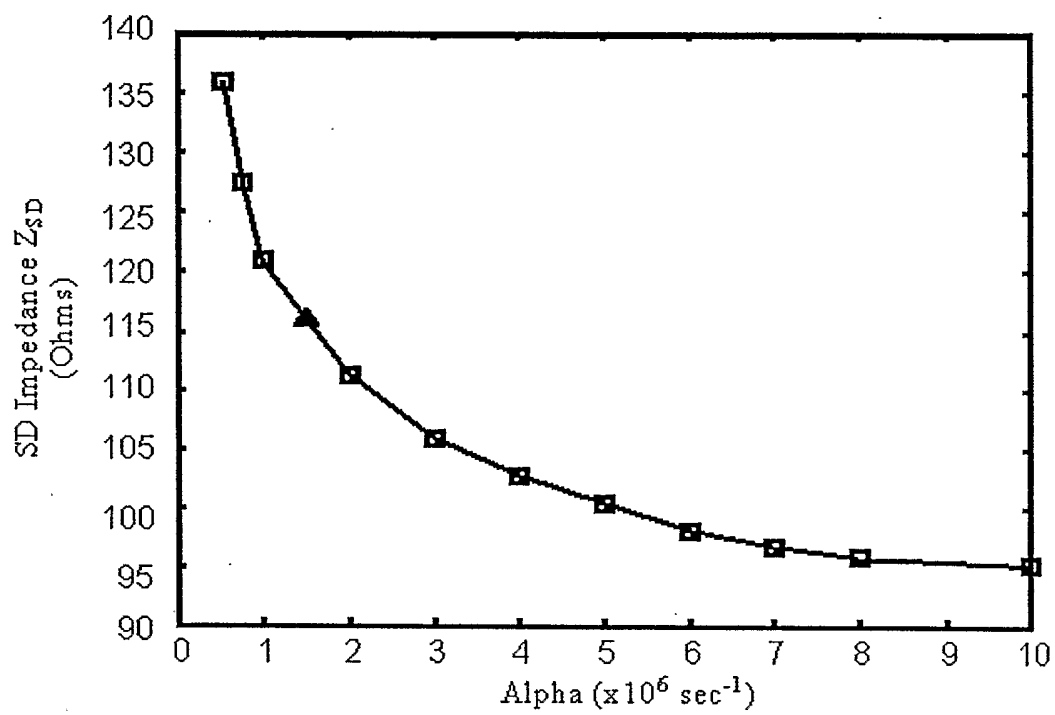


FIG. 16

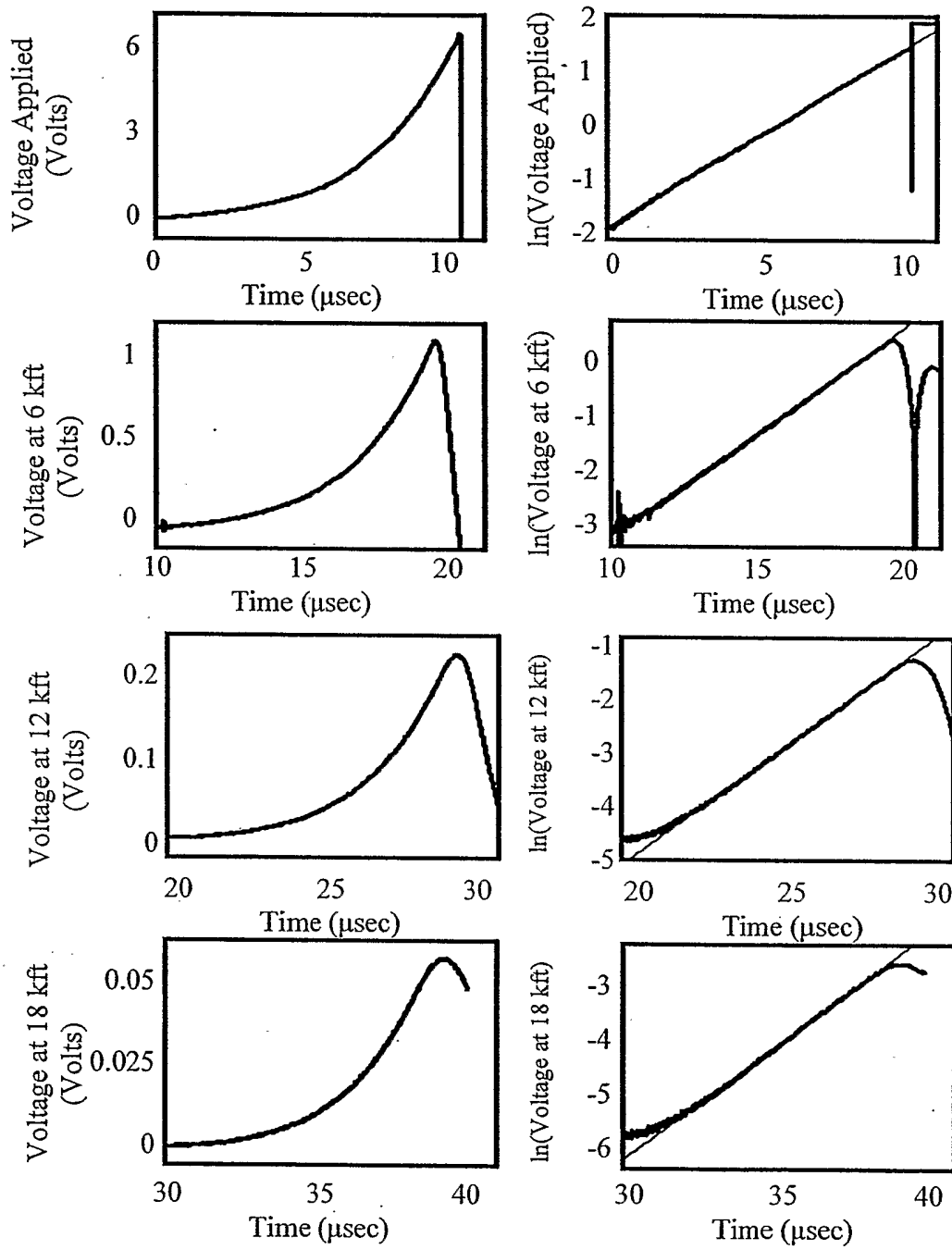


FIG. 17

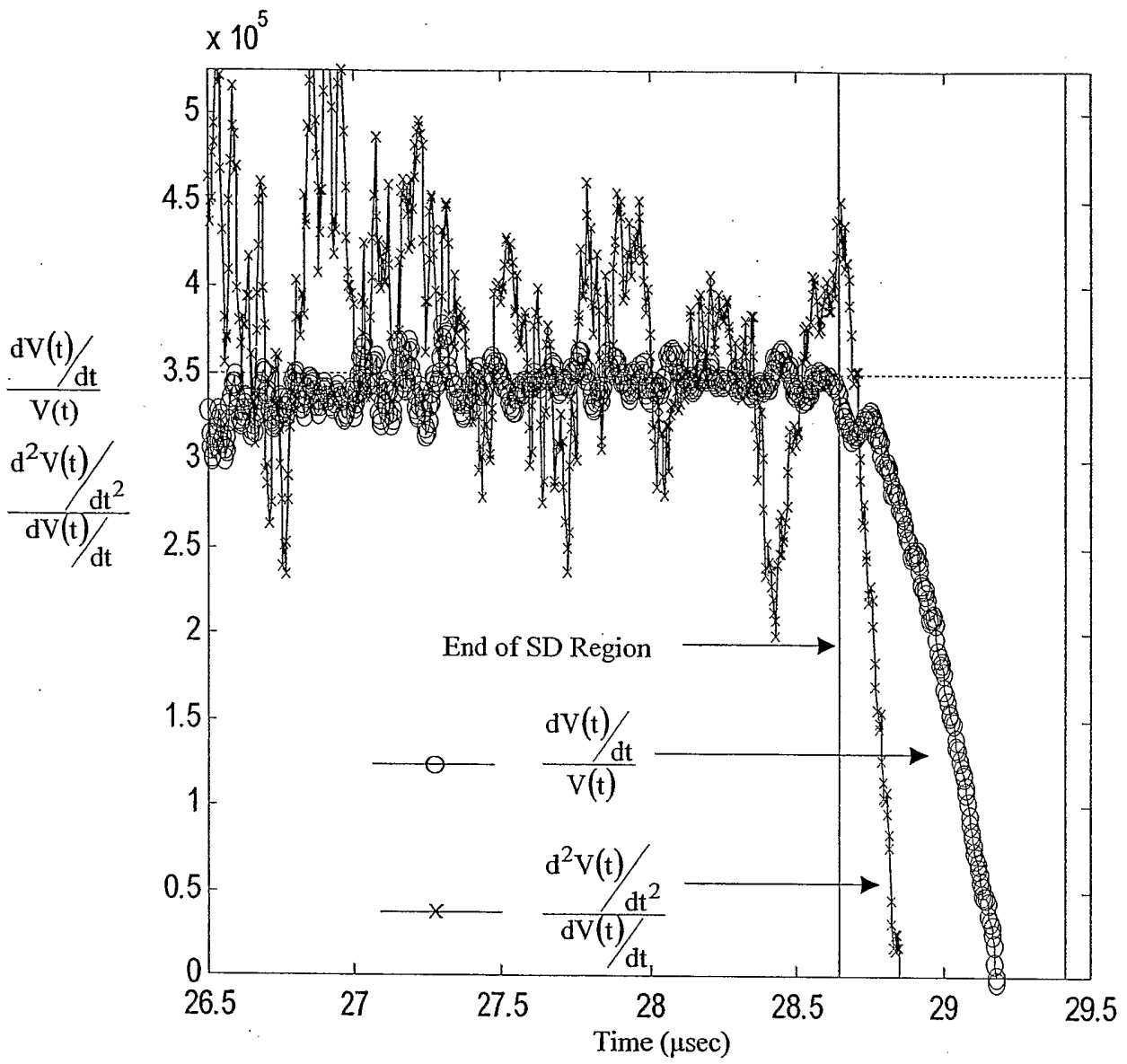


FIG. 18

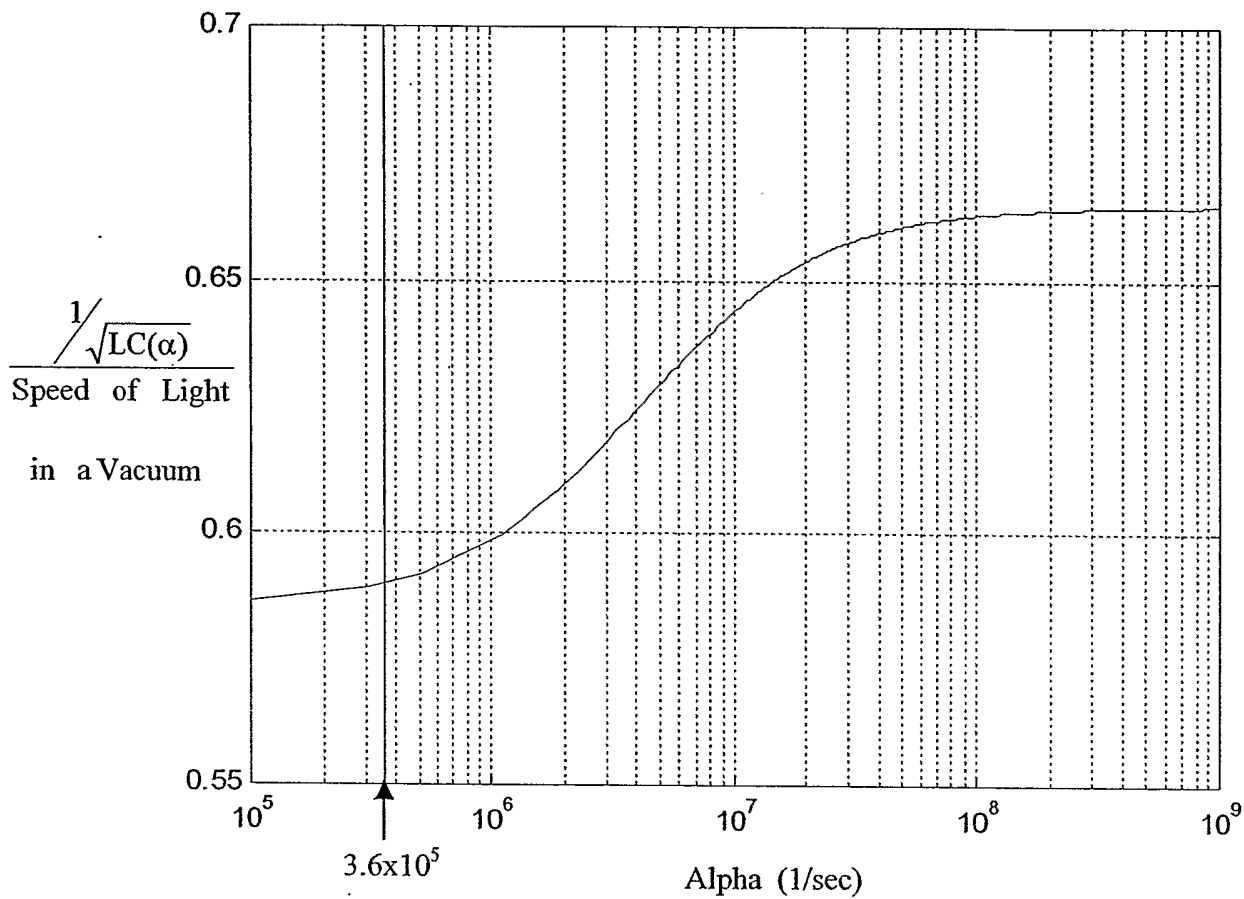


FIG. 19.

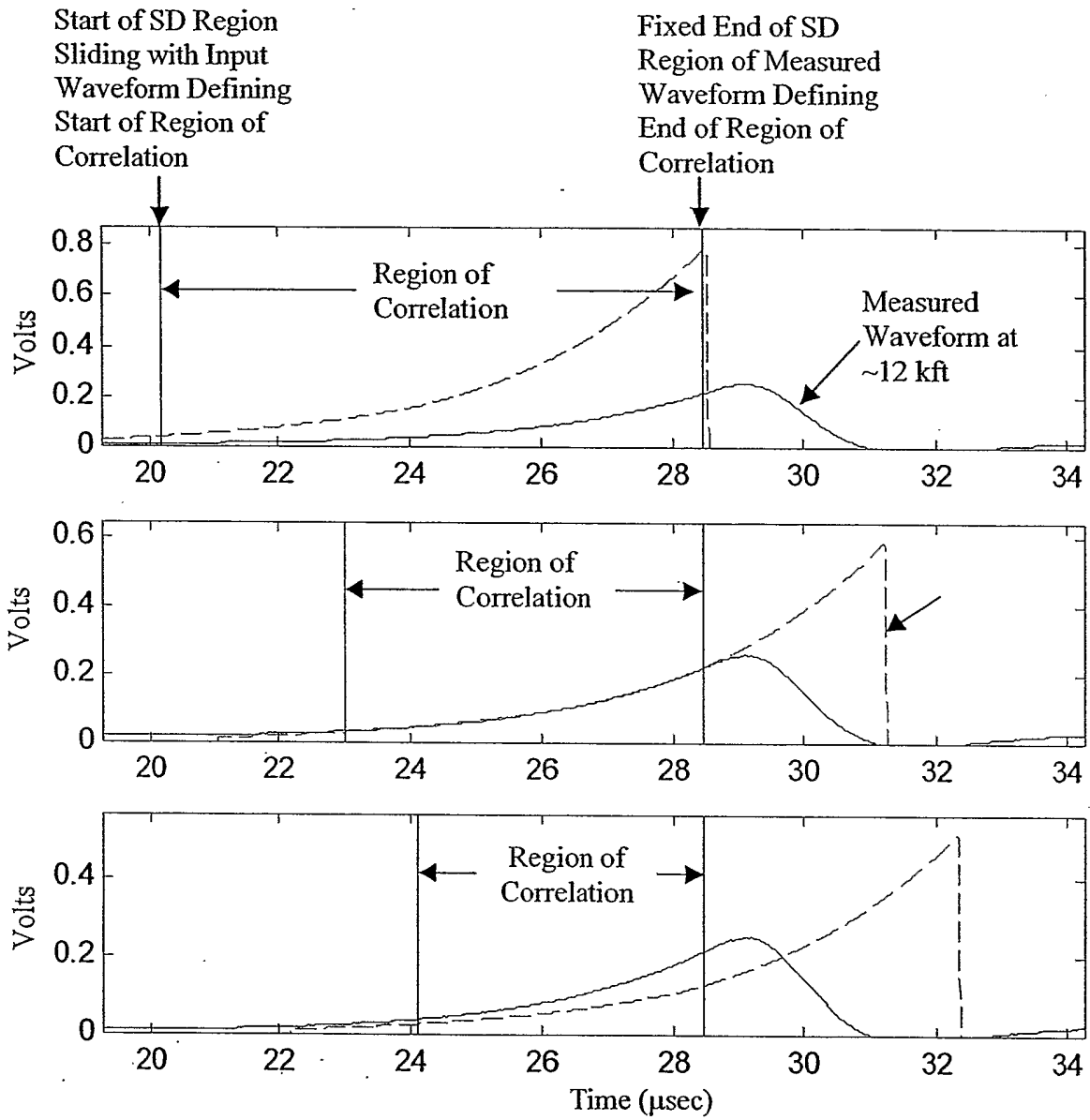


FIG. 20

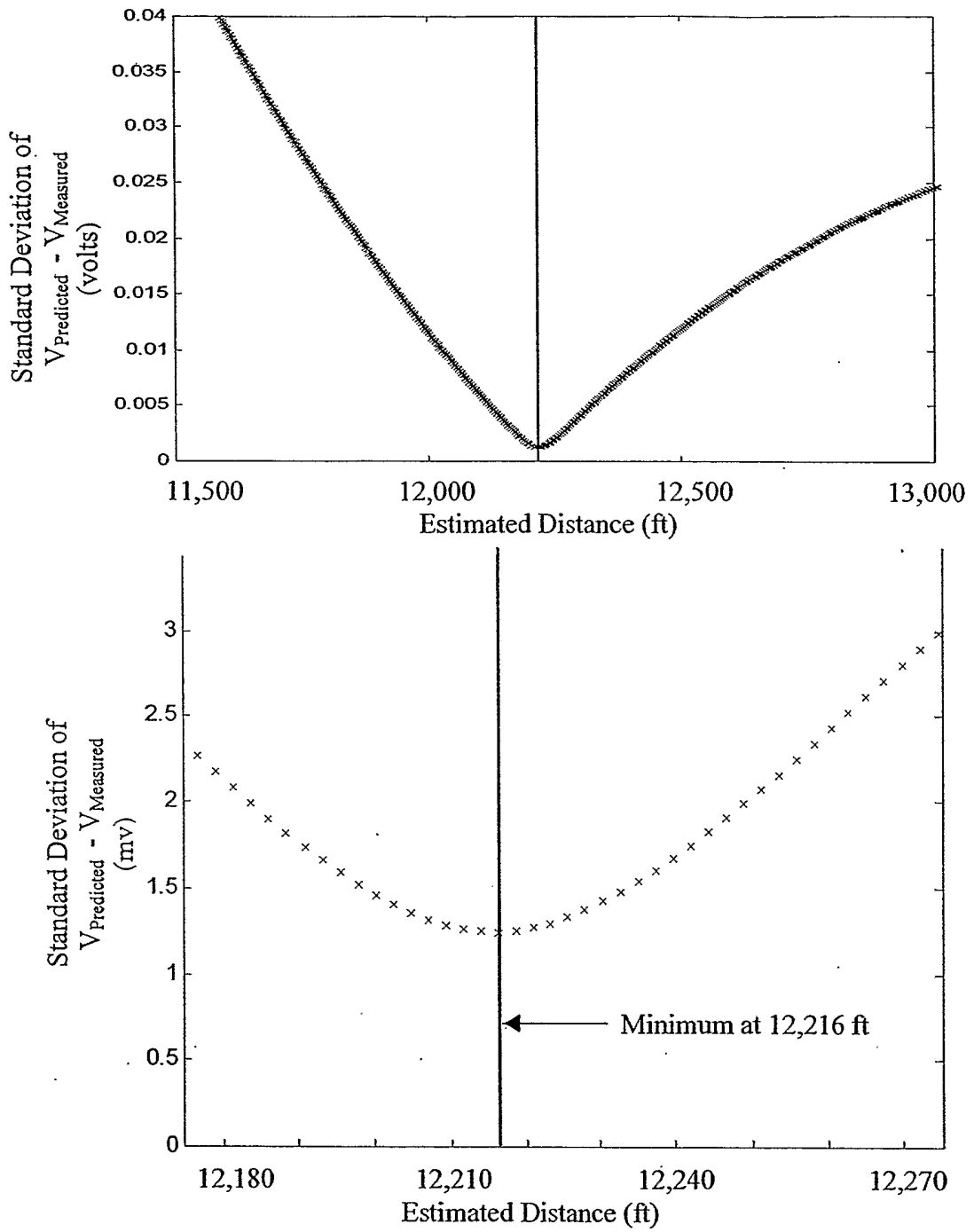


FIG. 21

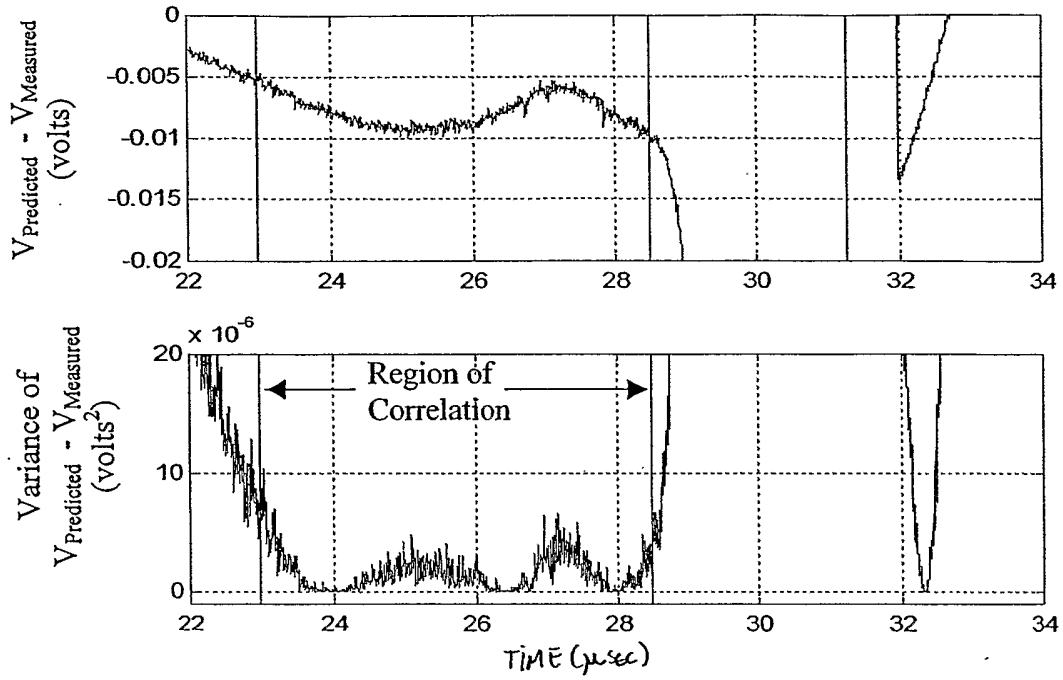


FIG. 22A

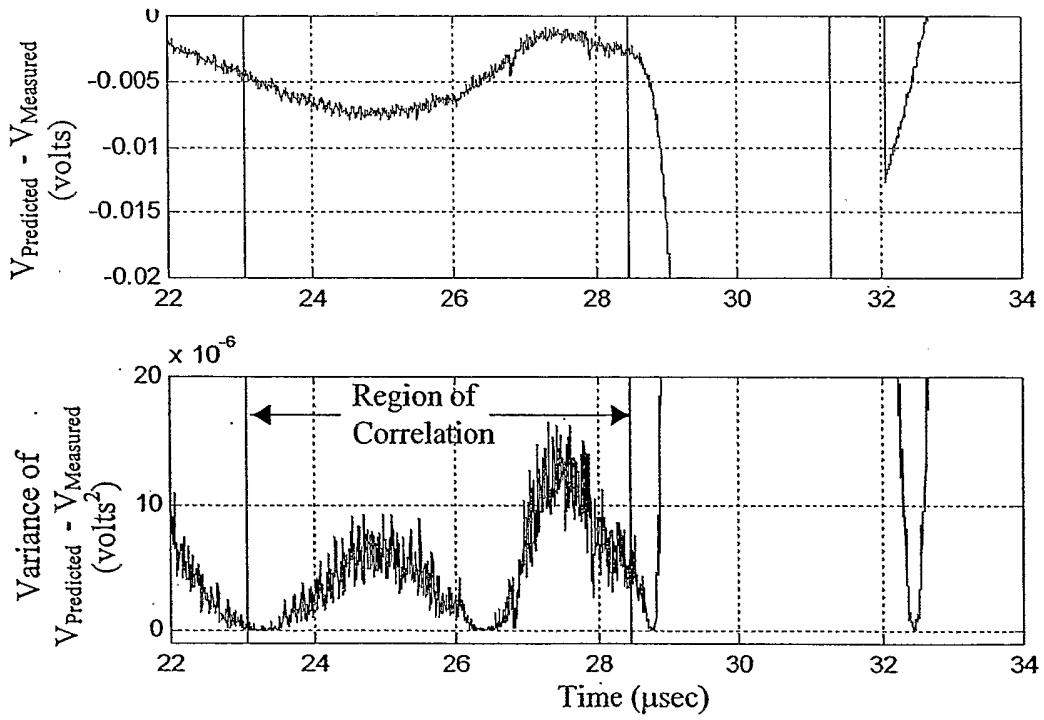


FIG. 22B

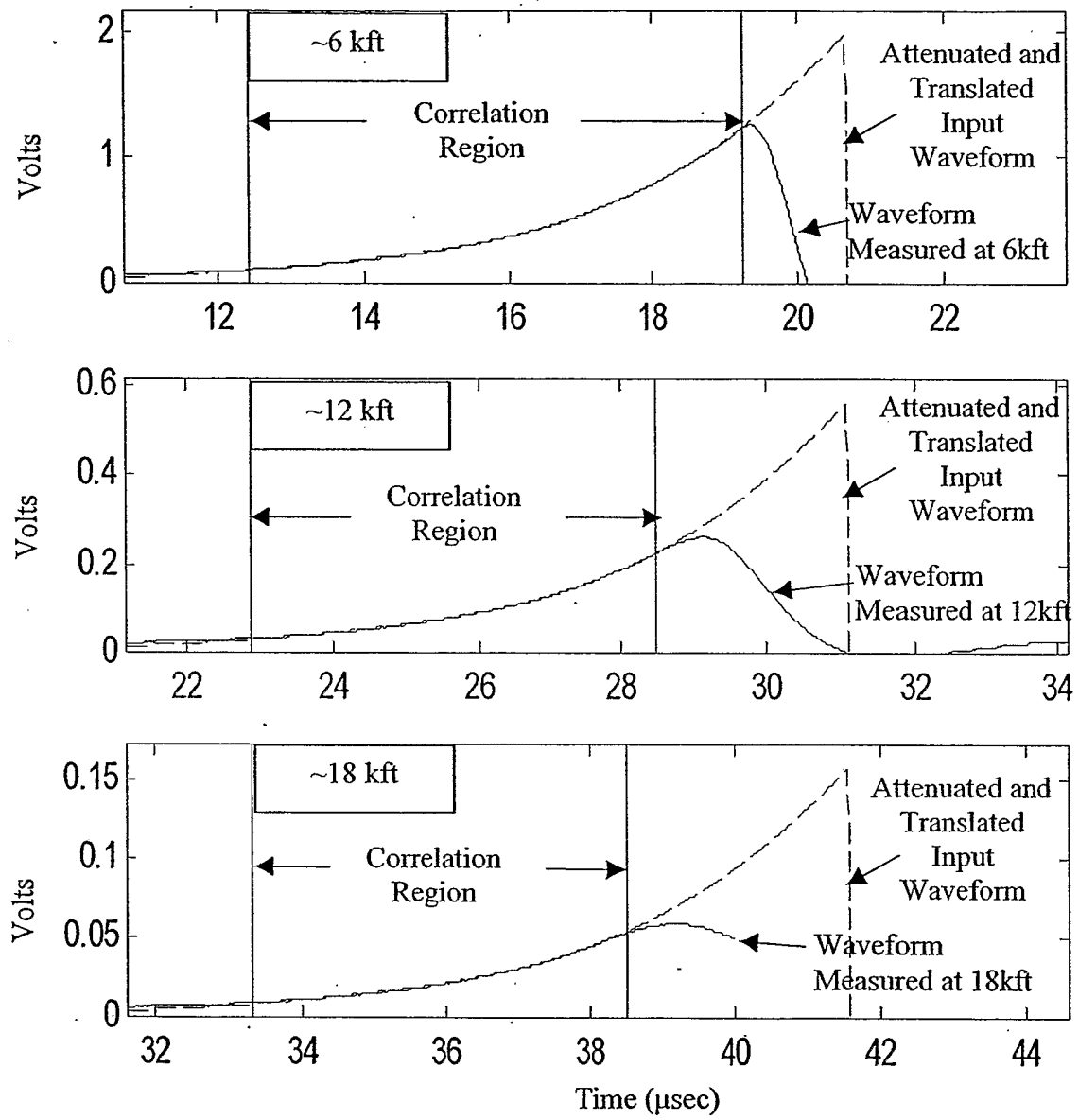


FIG. 23

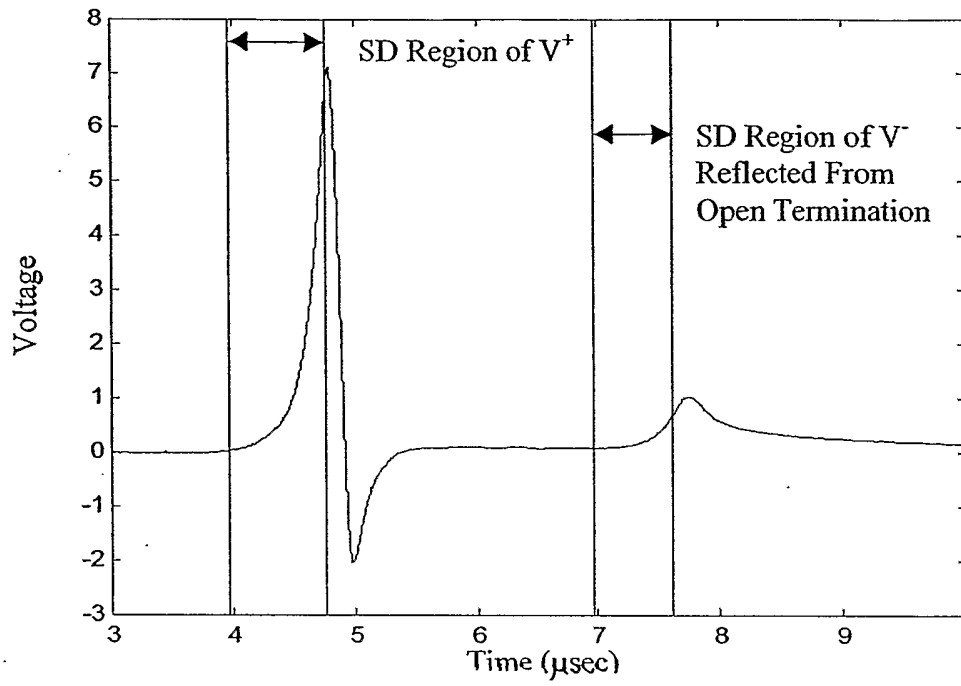


FIG. 24

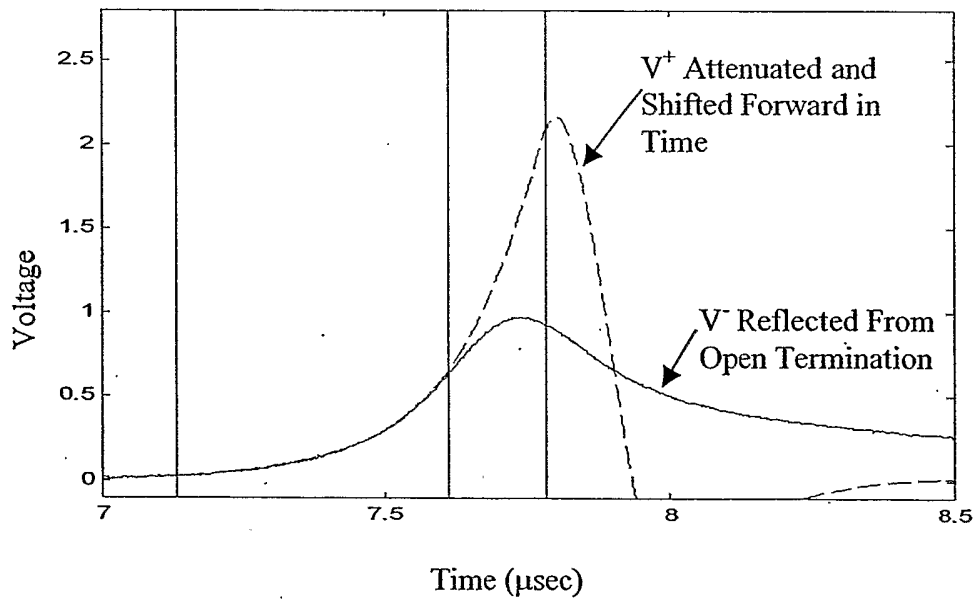


FIG. 25

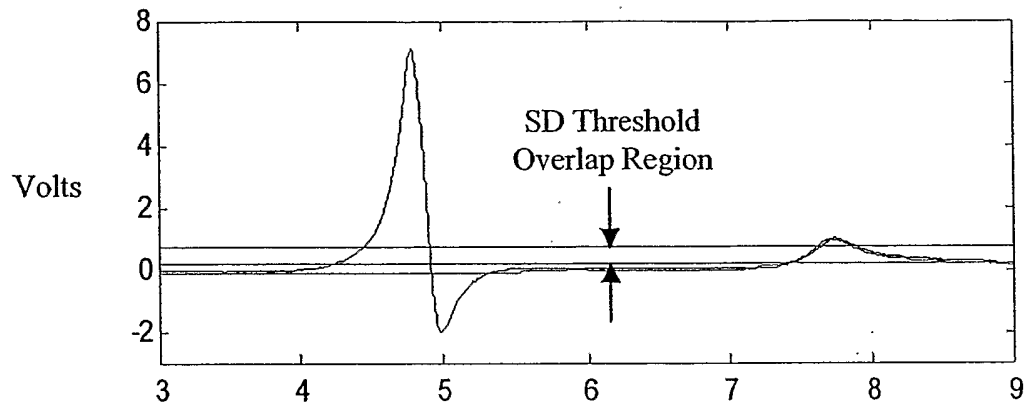


FIG. 26A

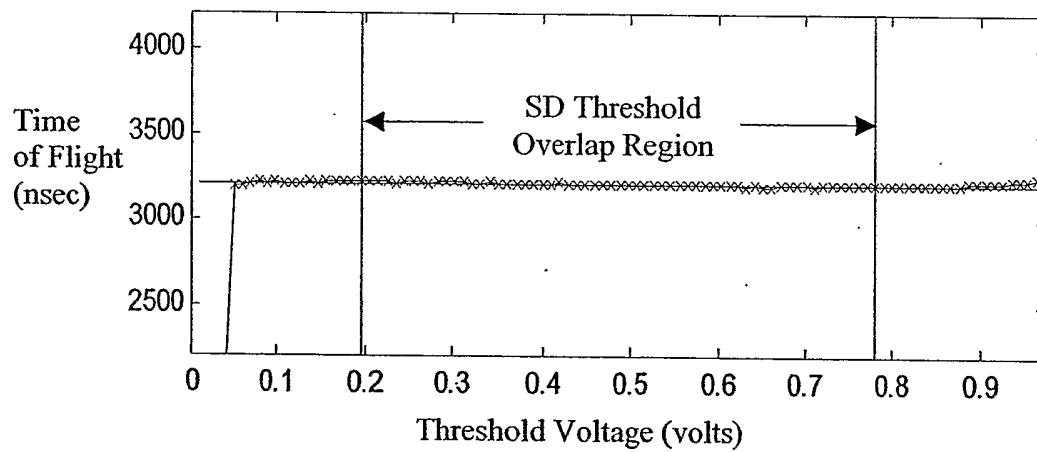


FIG. 26B

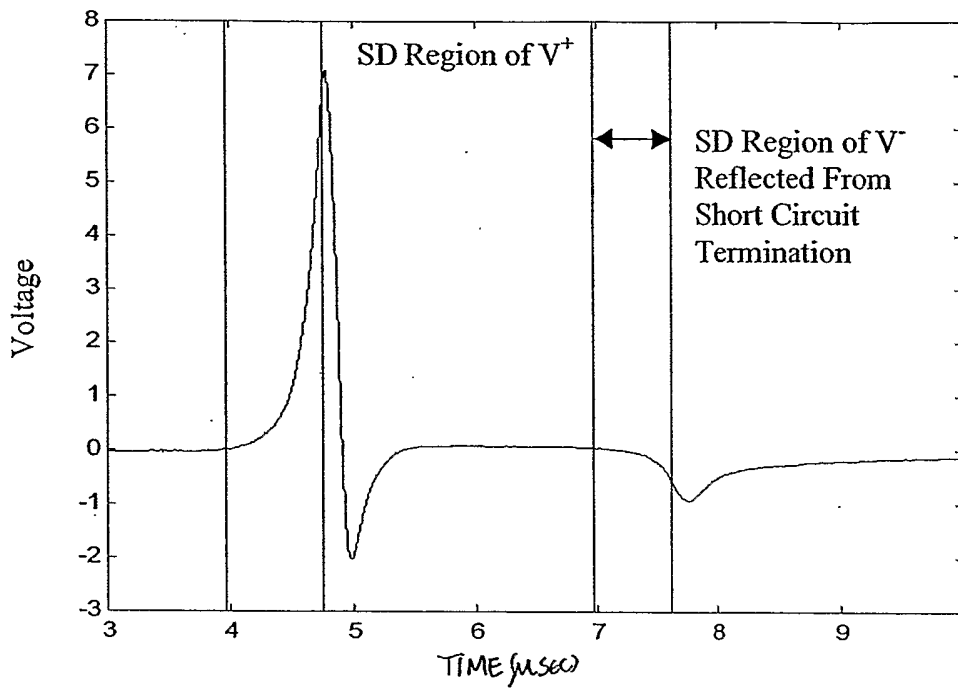


FIG. 27



FIG. 28

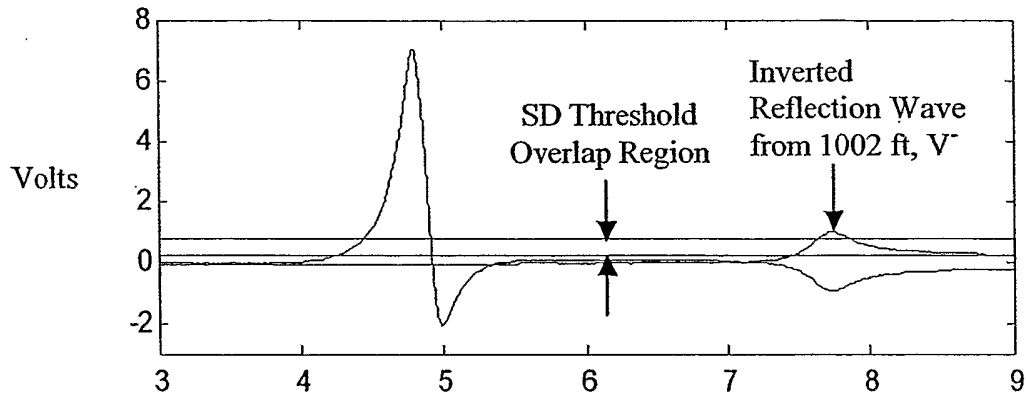


FIG. 29A

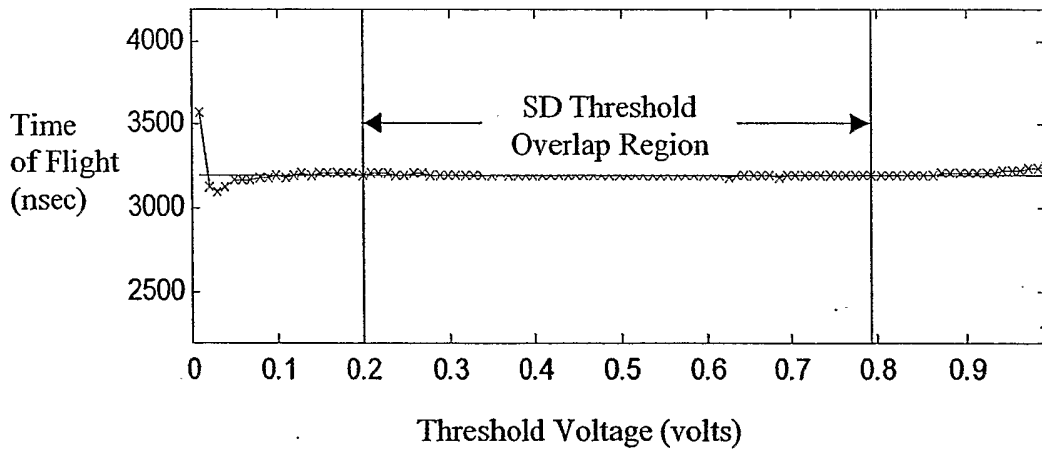


FIG. 29B

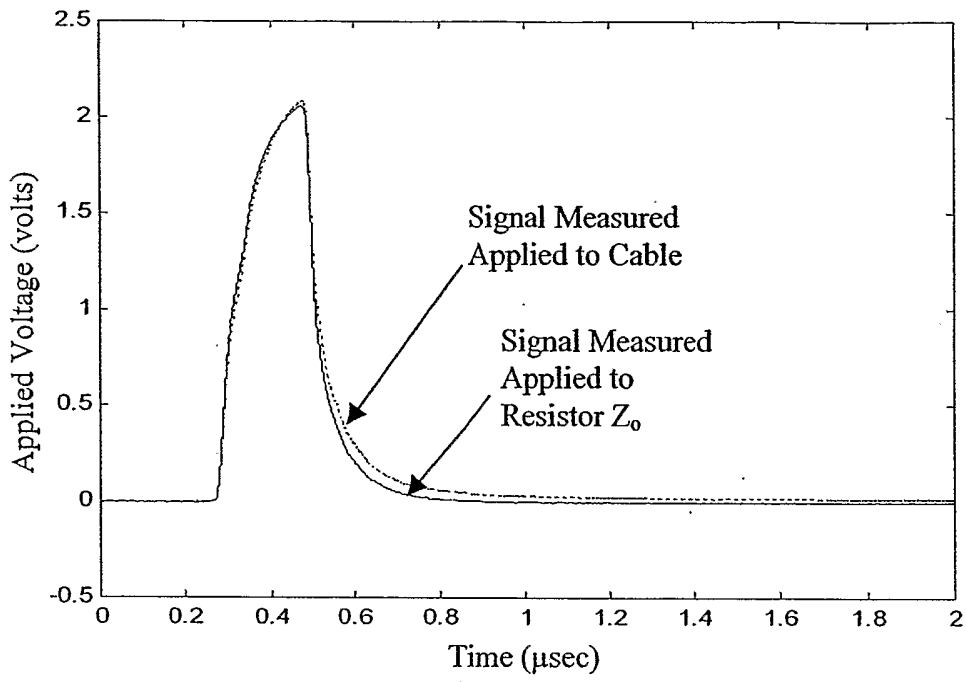


FIG. 30

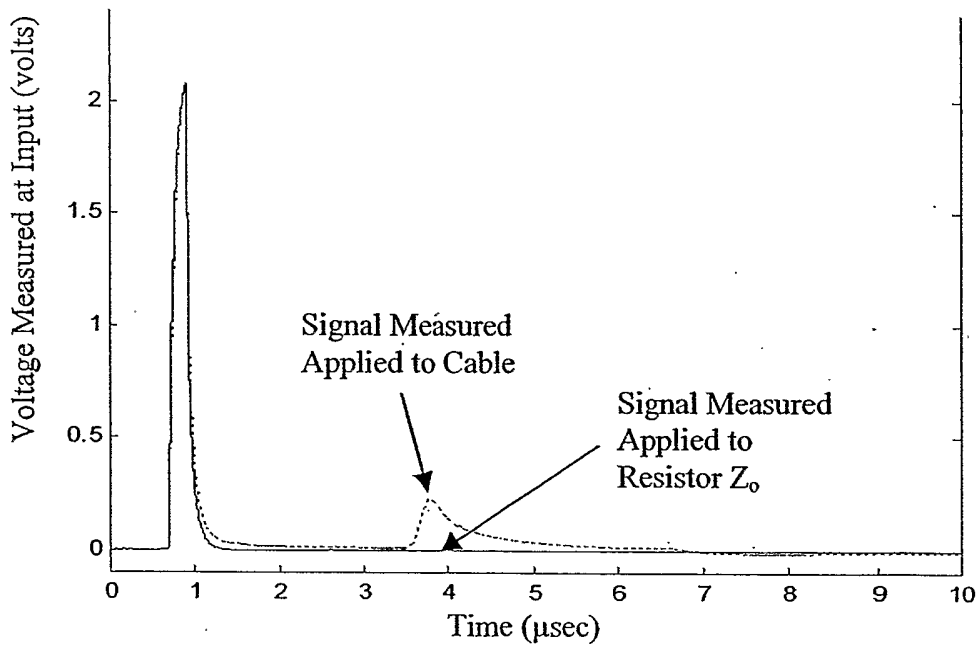


FIG. 31

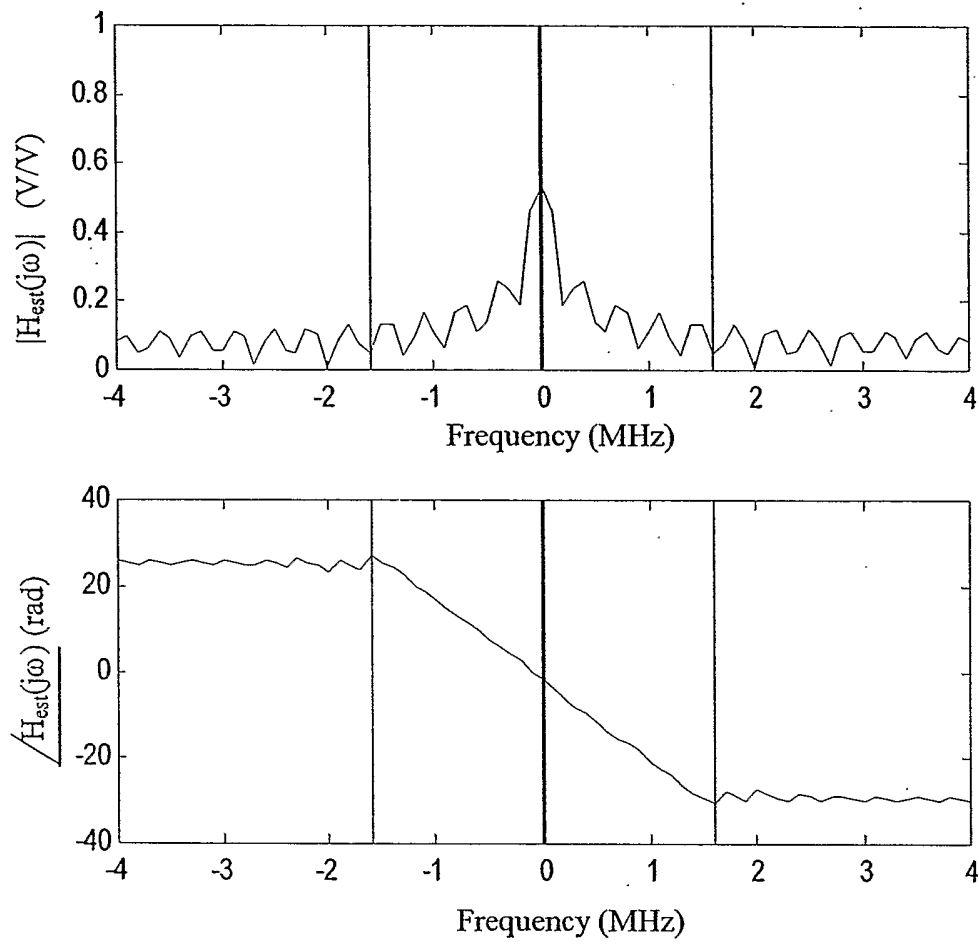


FIG. 32

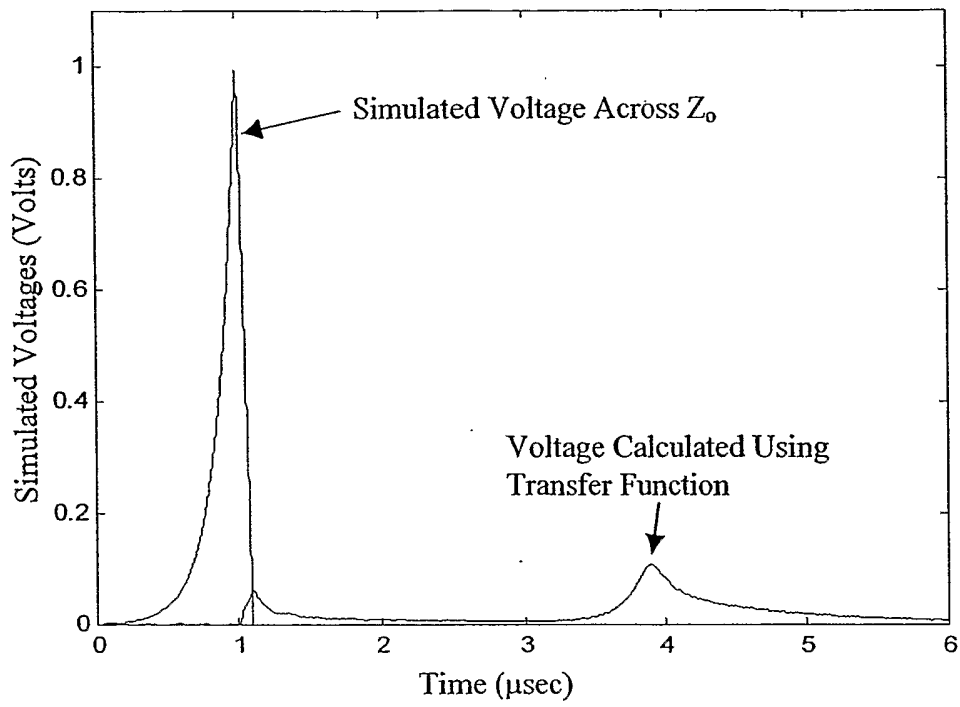


FIG. 33

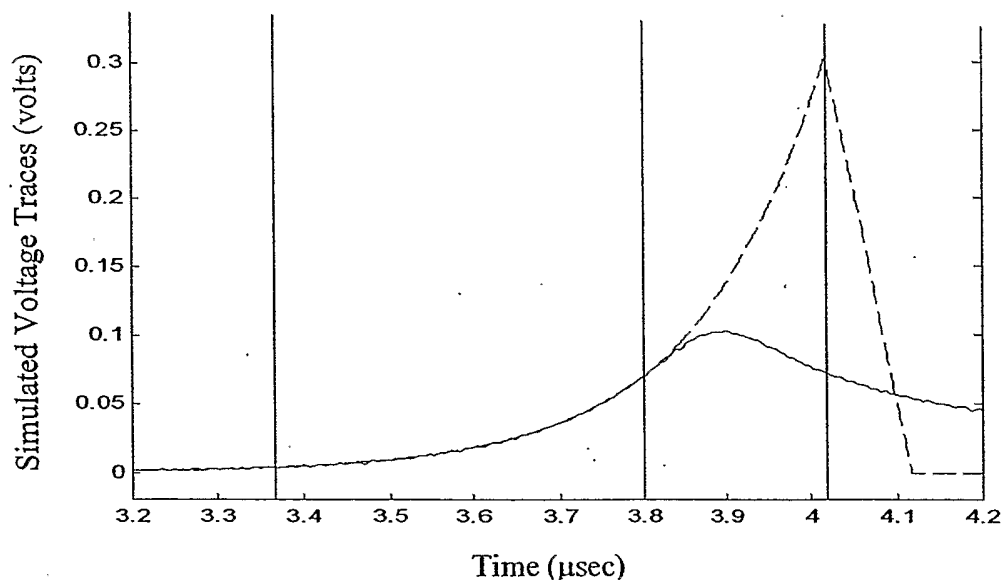


FIG. 34

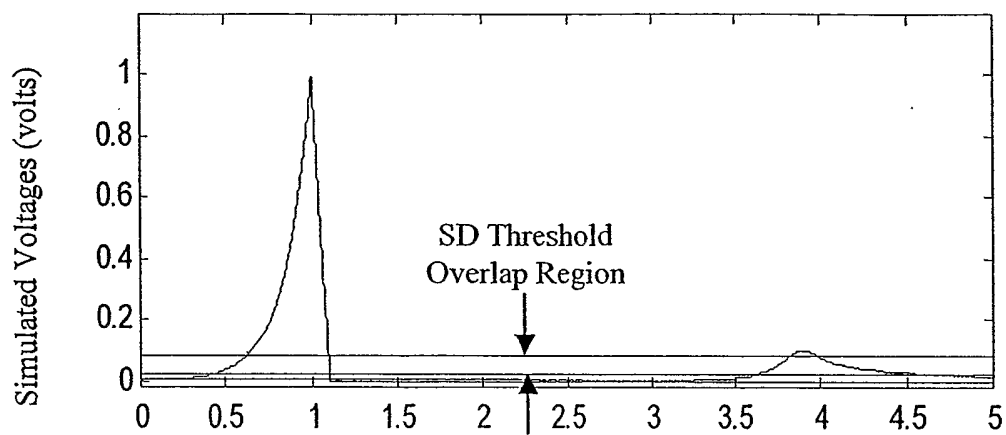


FIG. 35A

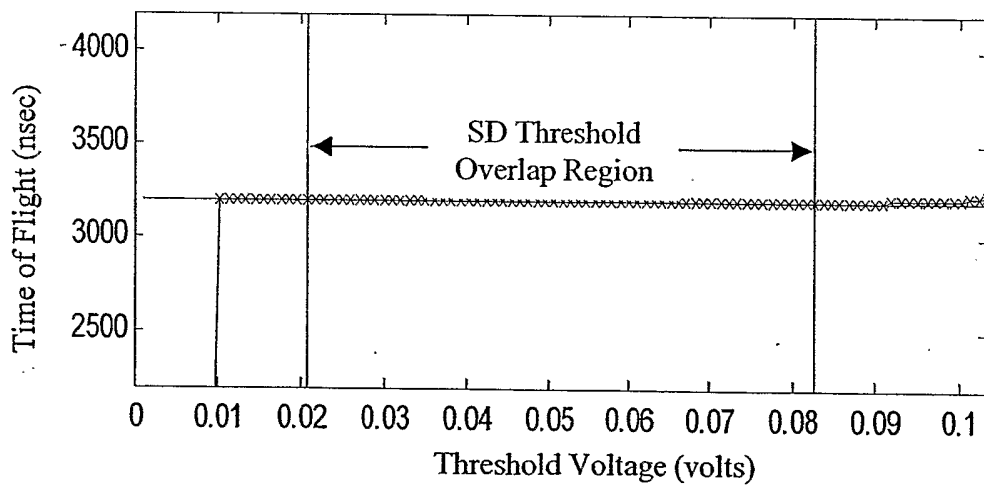


FIG. 35B

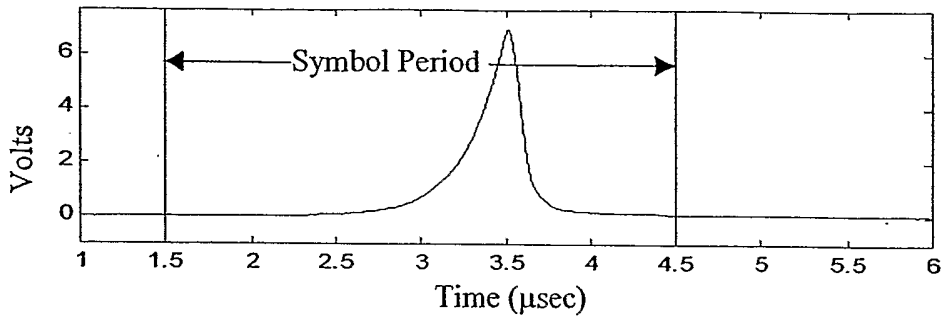


FIG. 36A

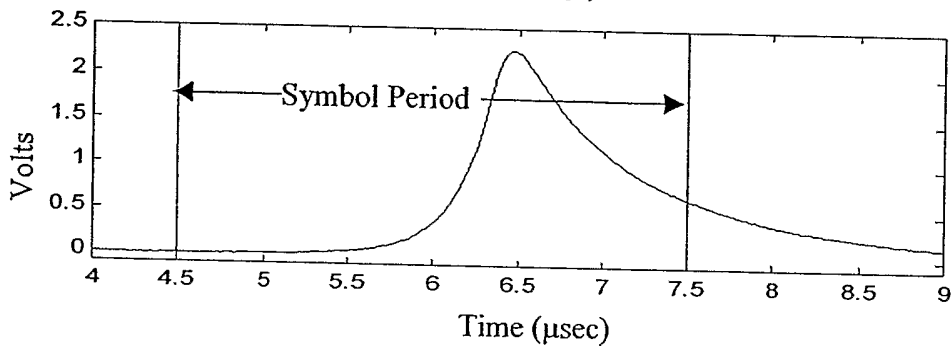


FIG. 36B

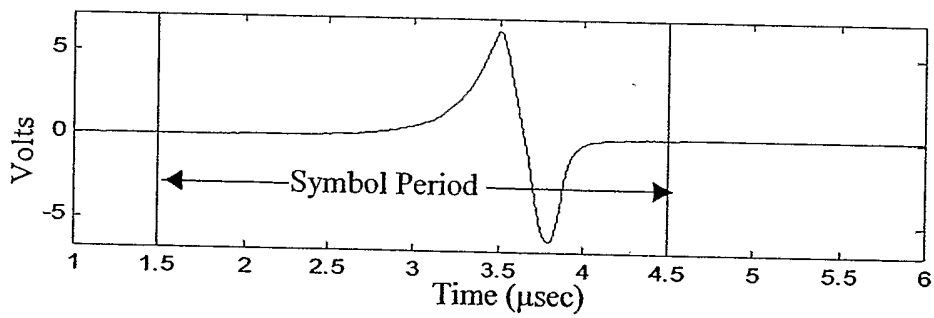


FIG. 37A

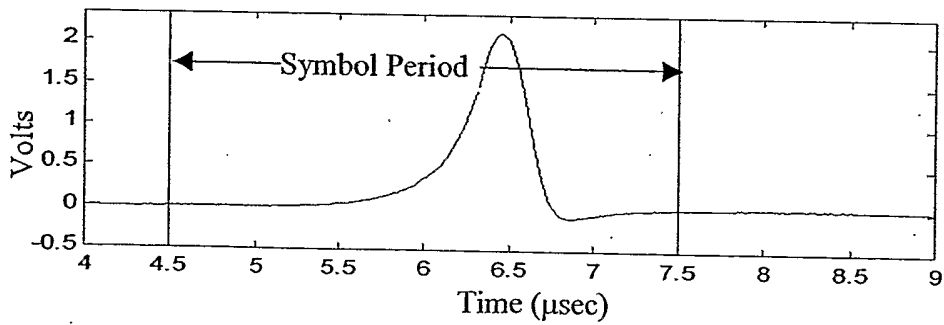


FIG. 37B

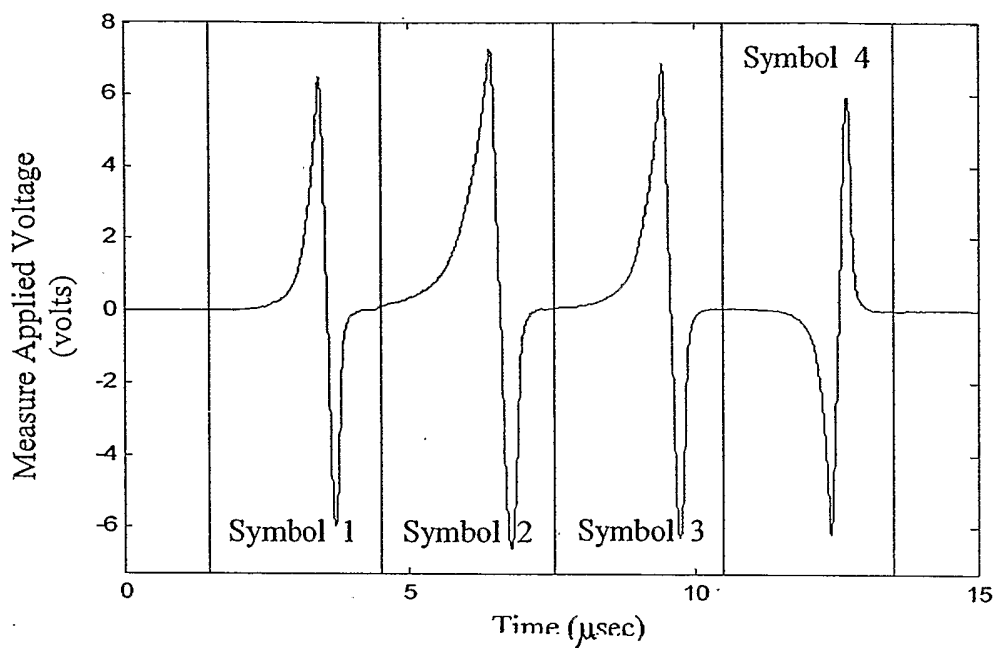


FIG. 38

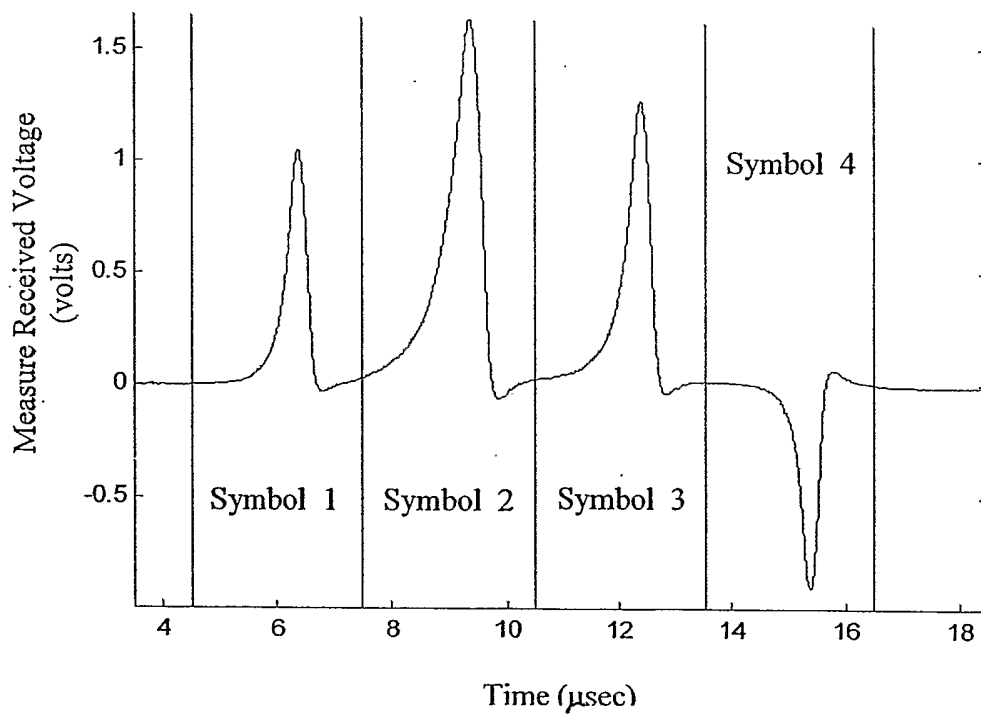


FIG. 39

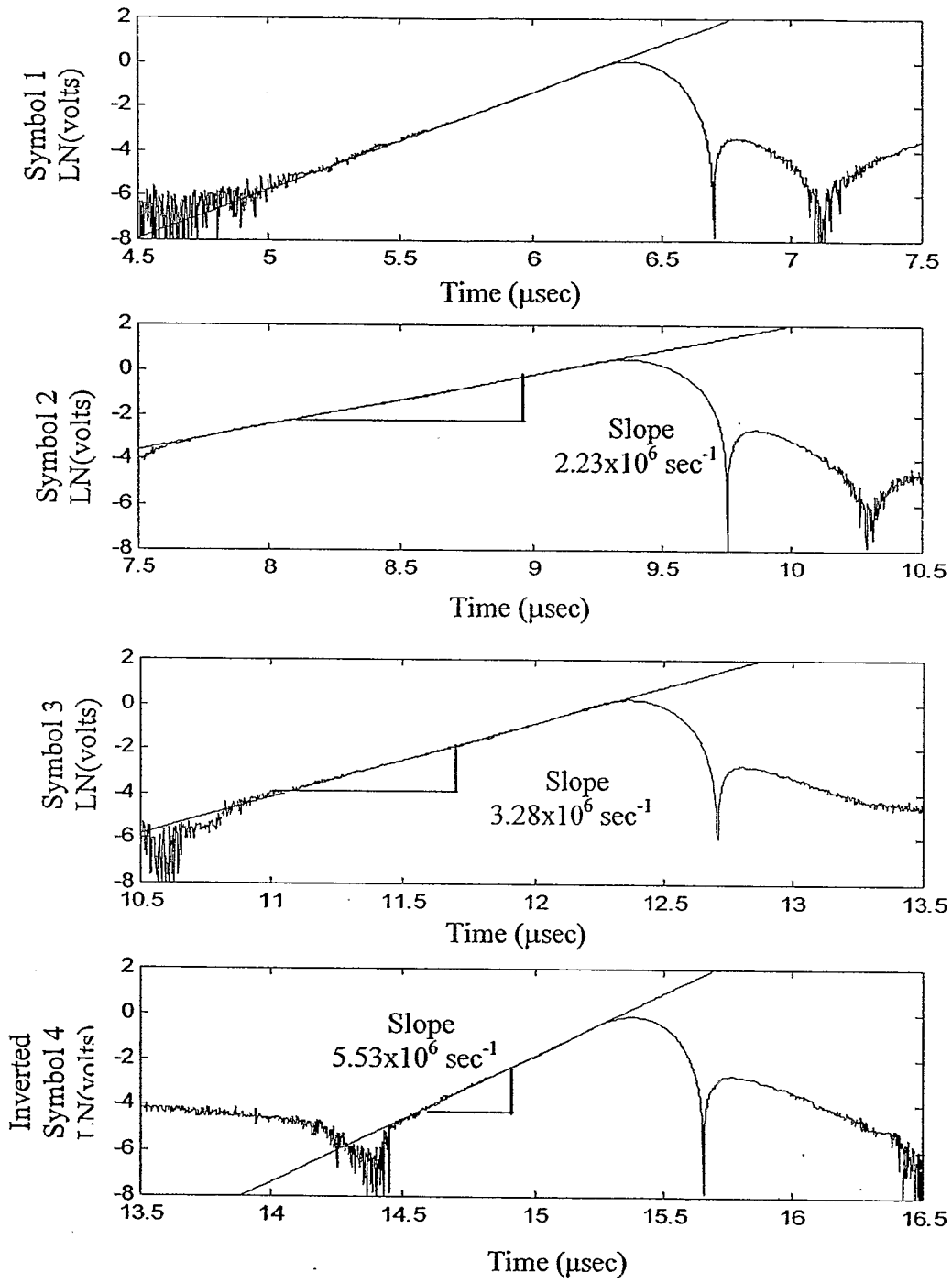


FIG. 40

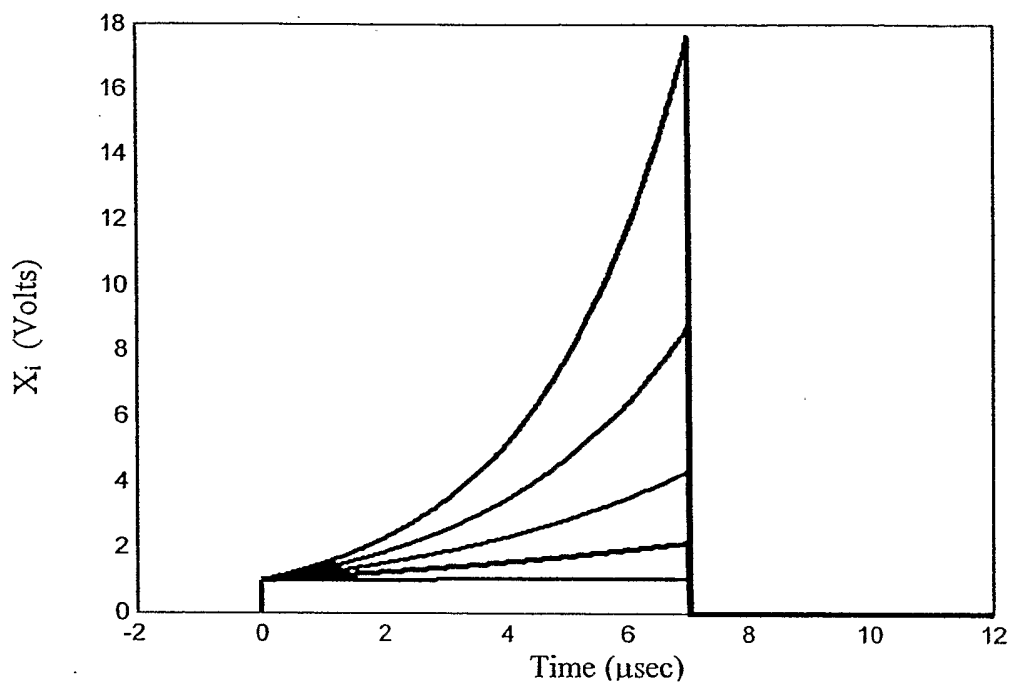


FIG. 41

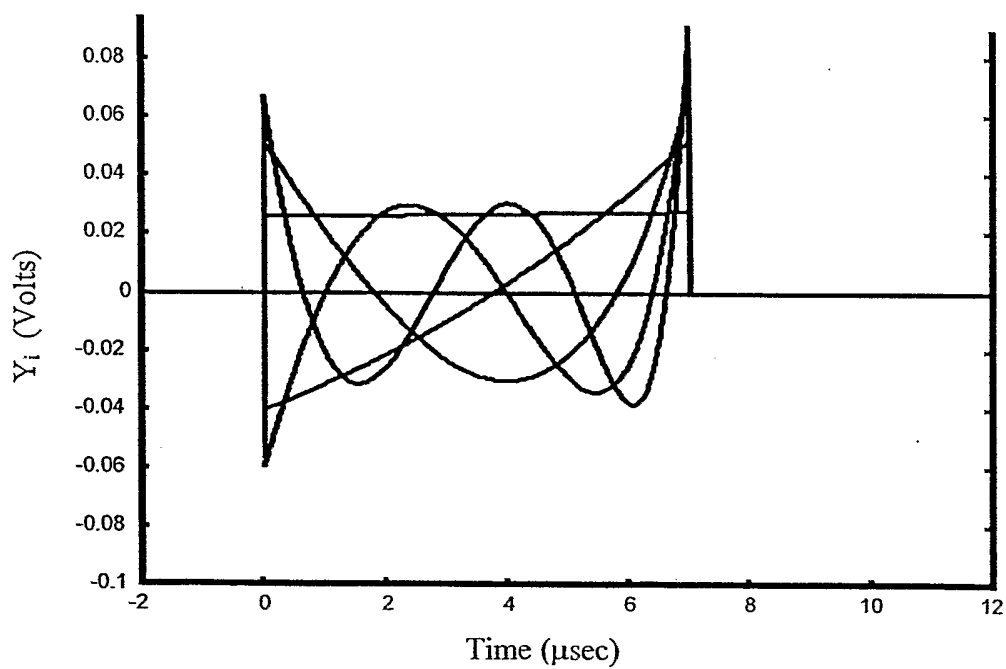


FIG. 42

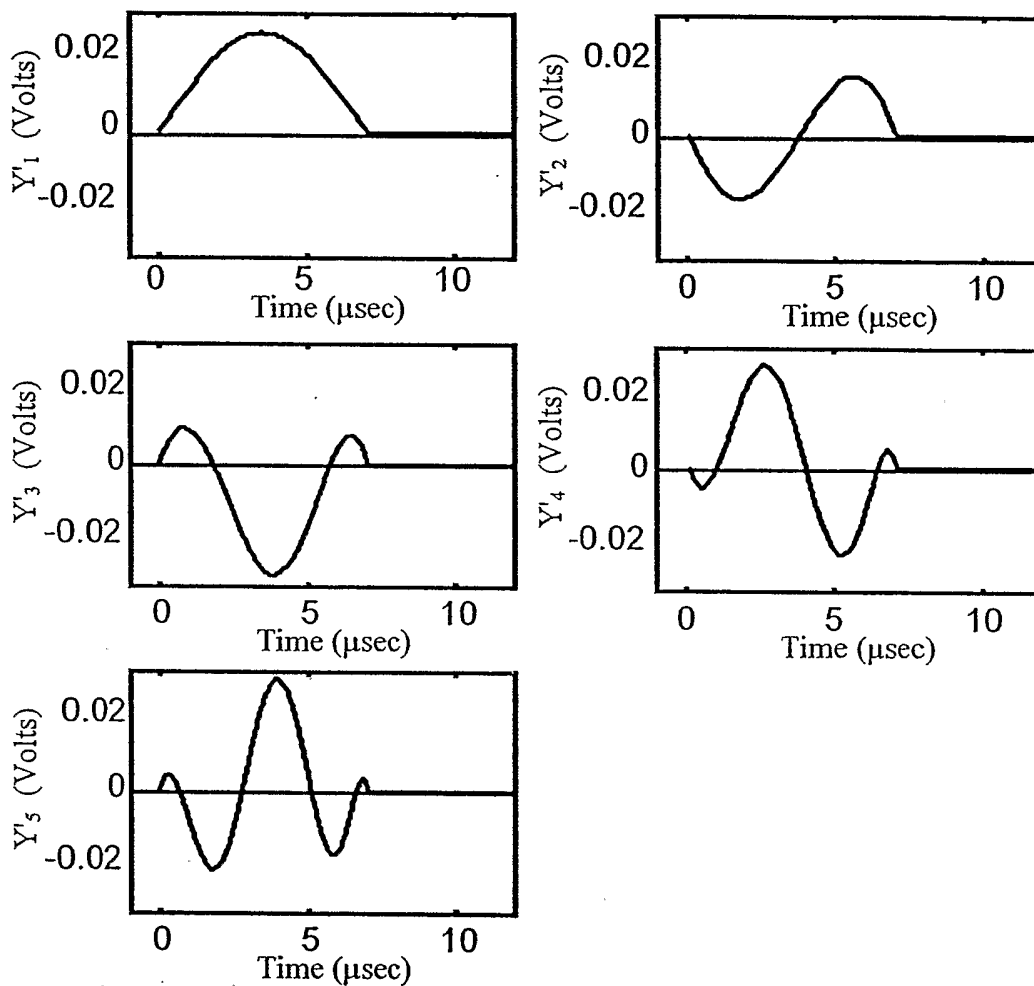


FIG. 43

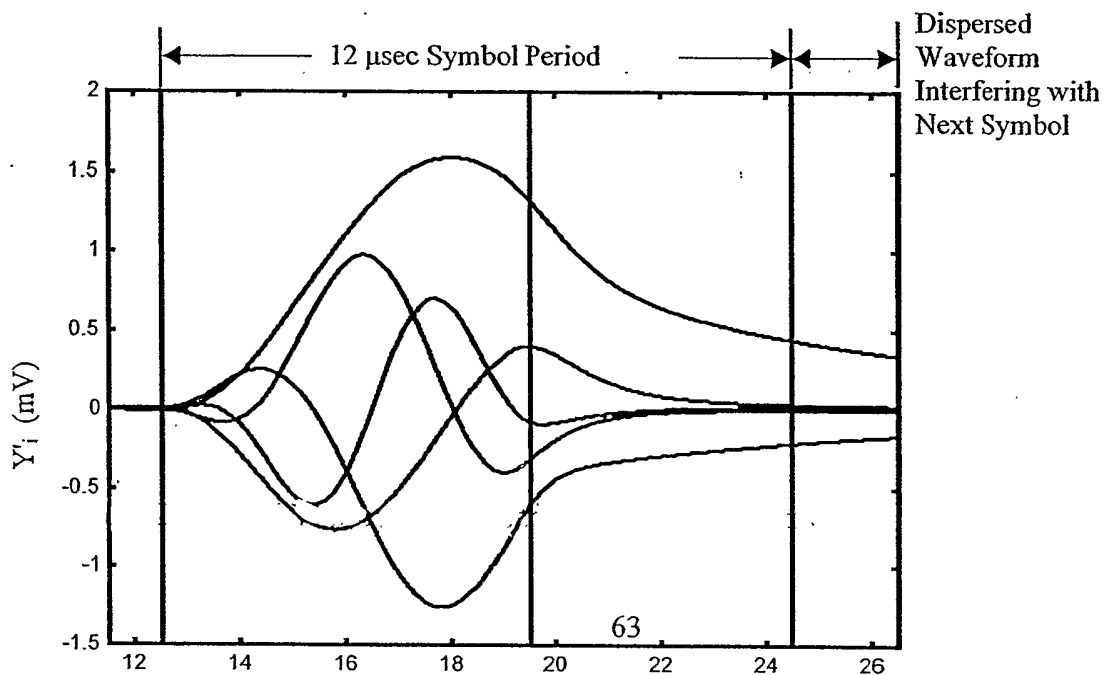


FIG. 44

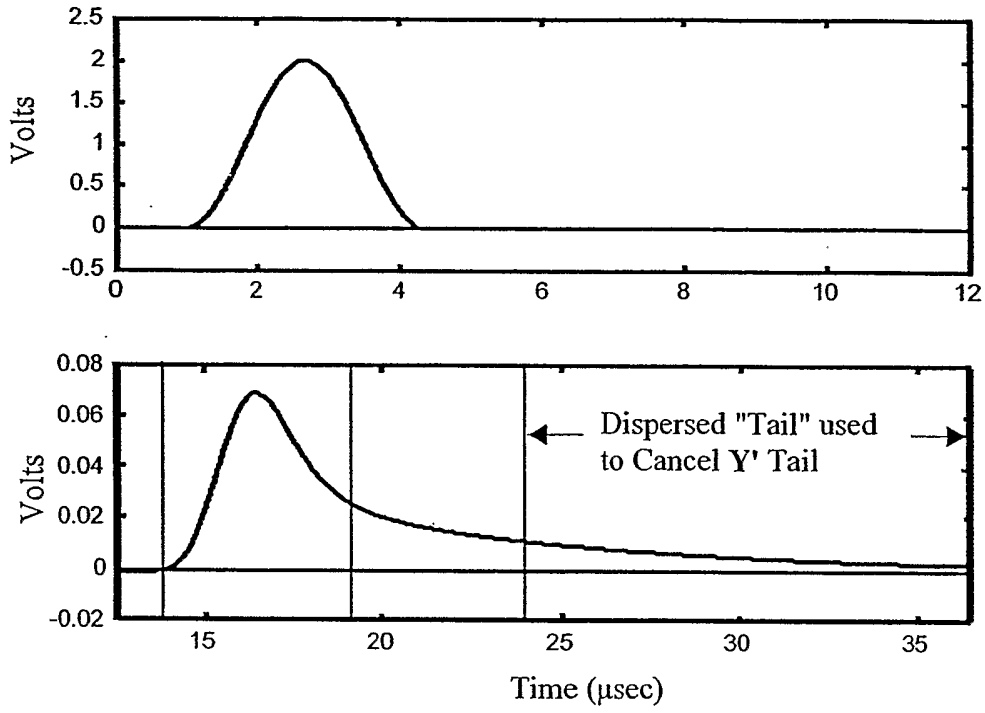


FIG. 45

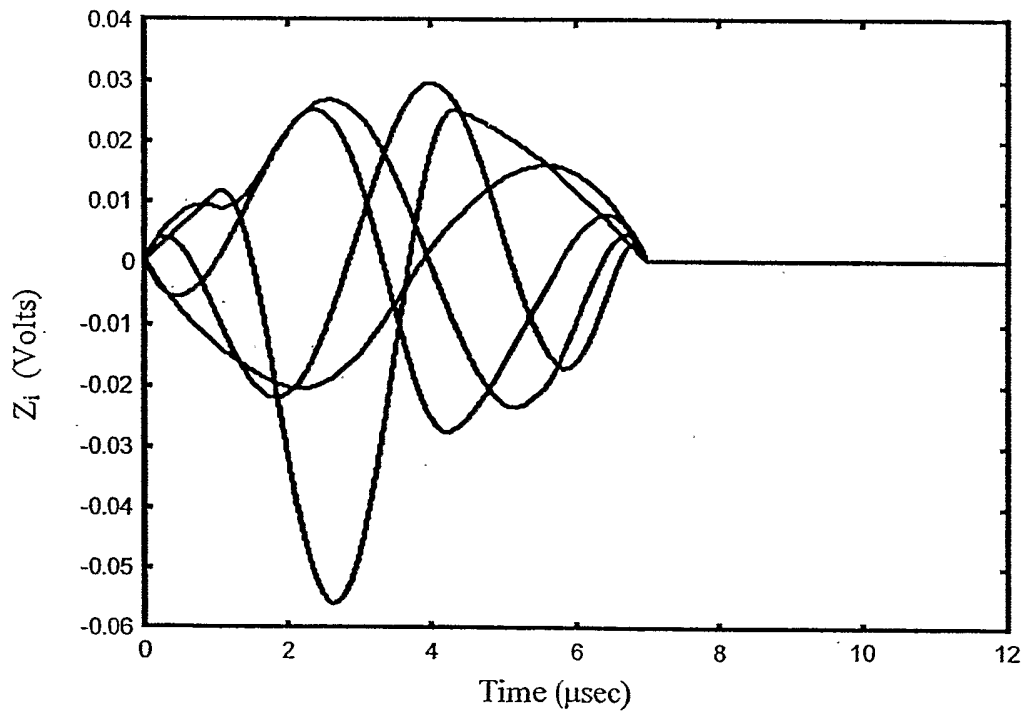


FIG. 46

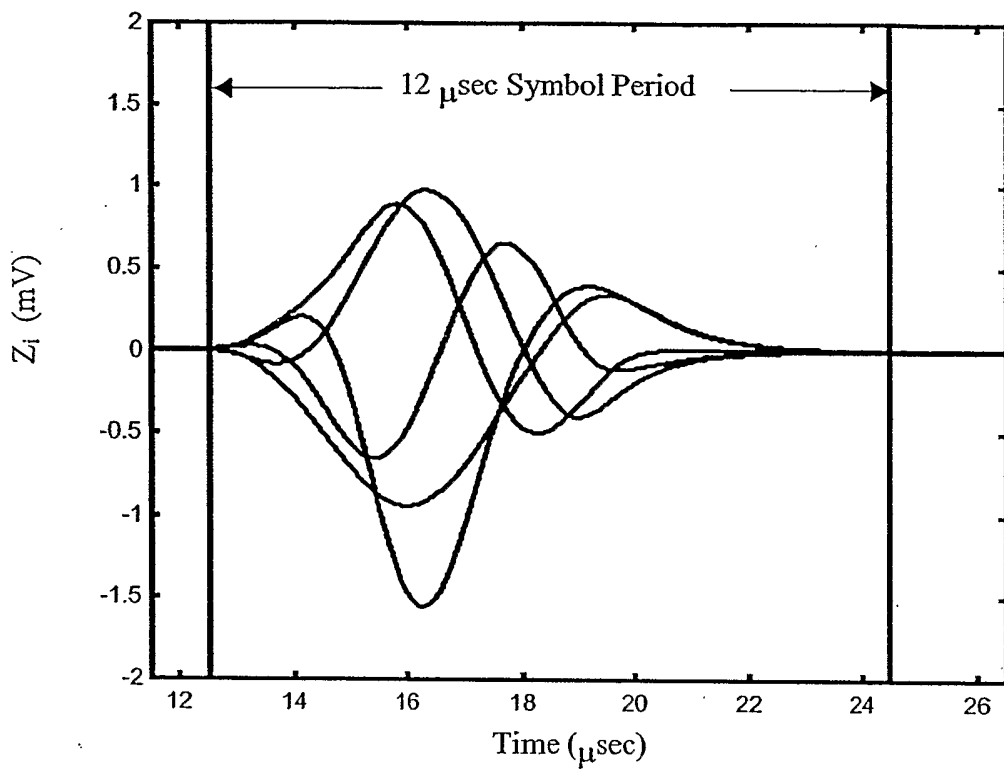


FIG. 47

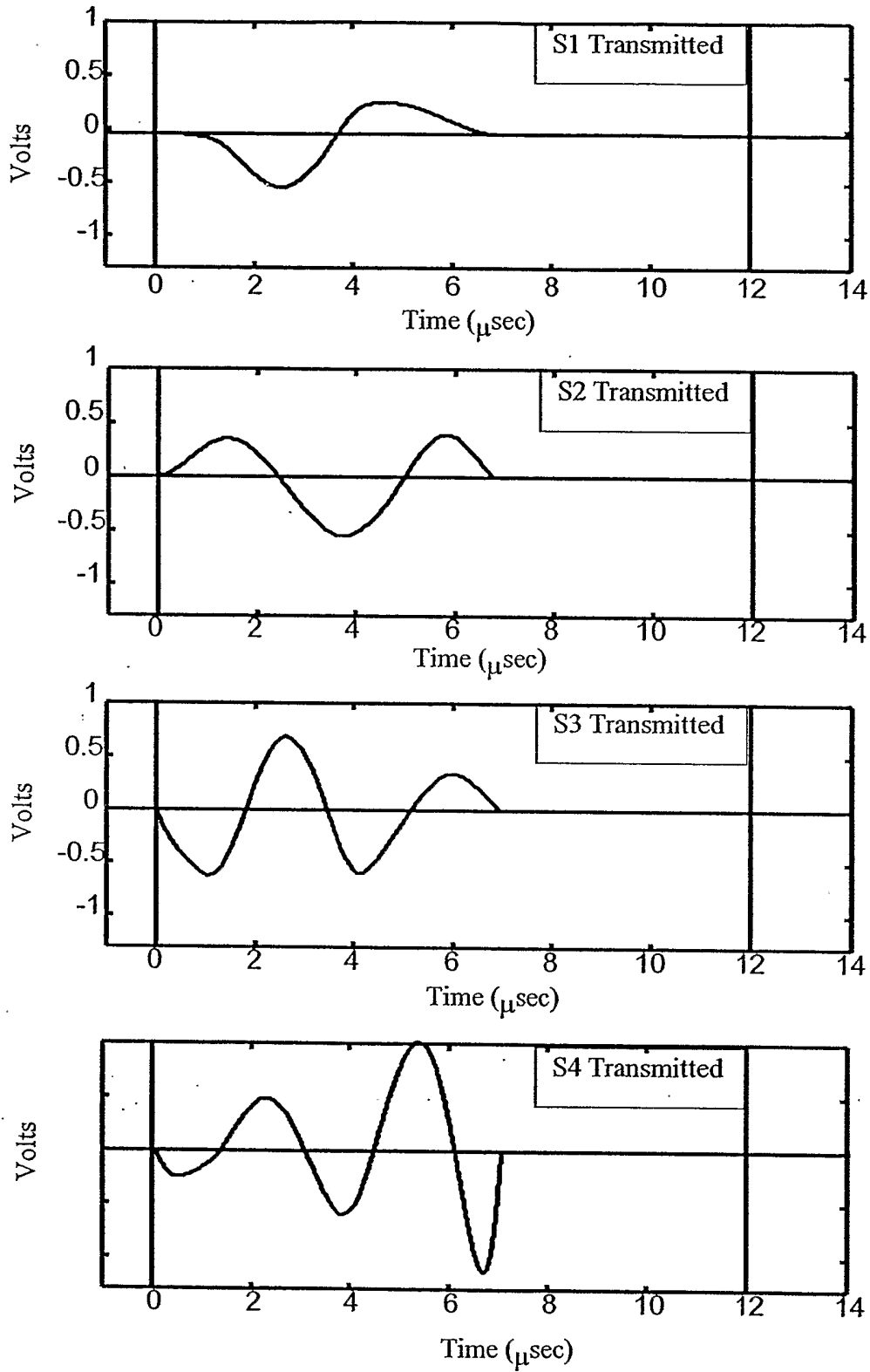


FIG. 48

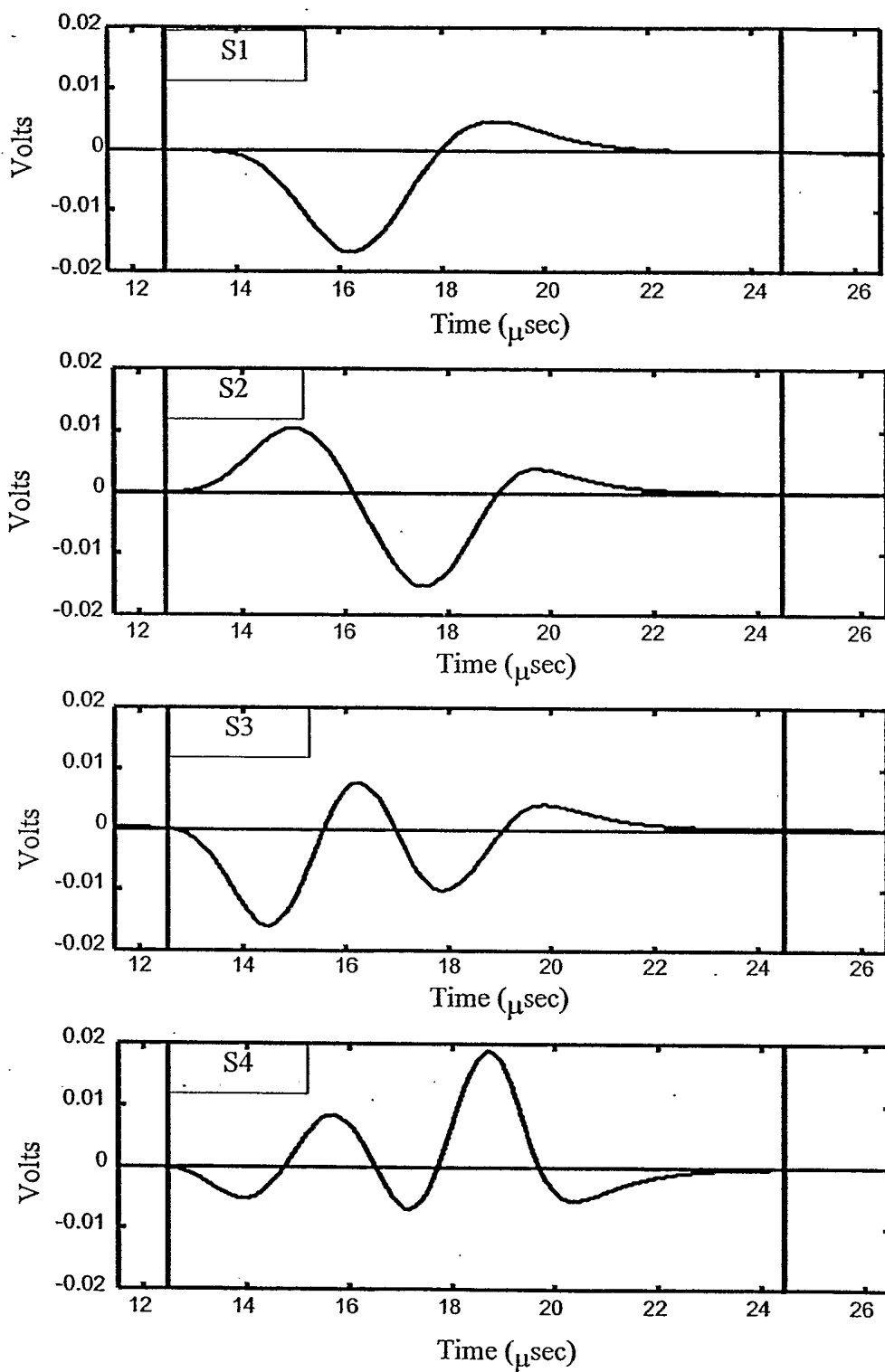


FIG. 49

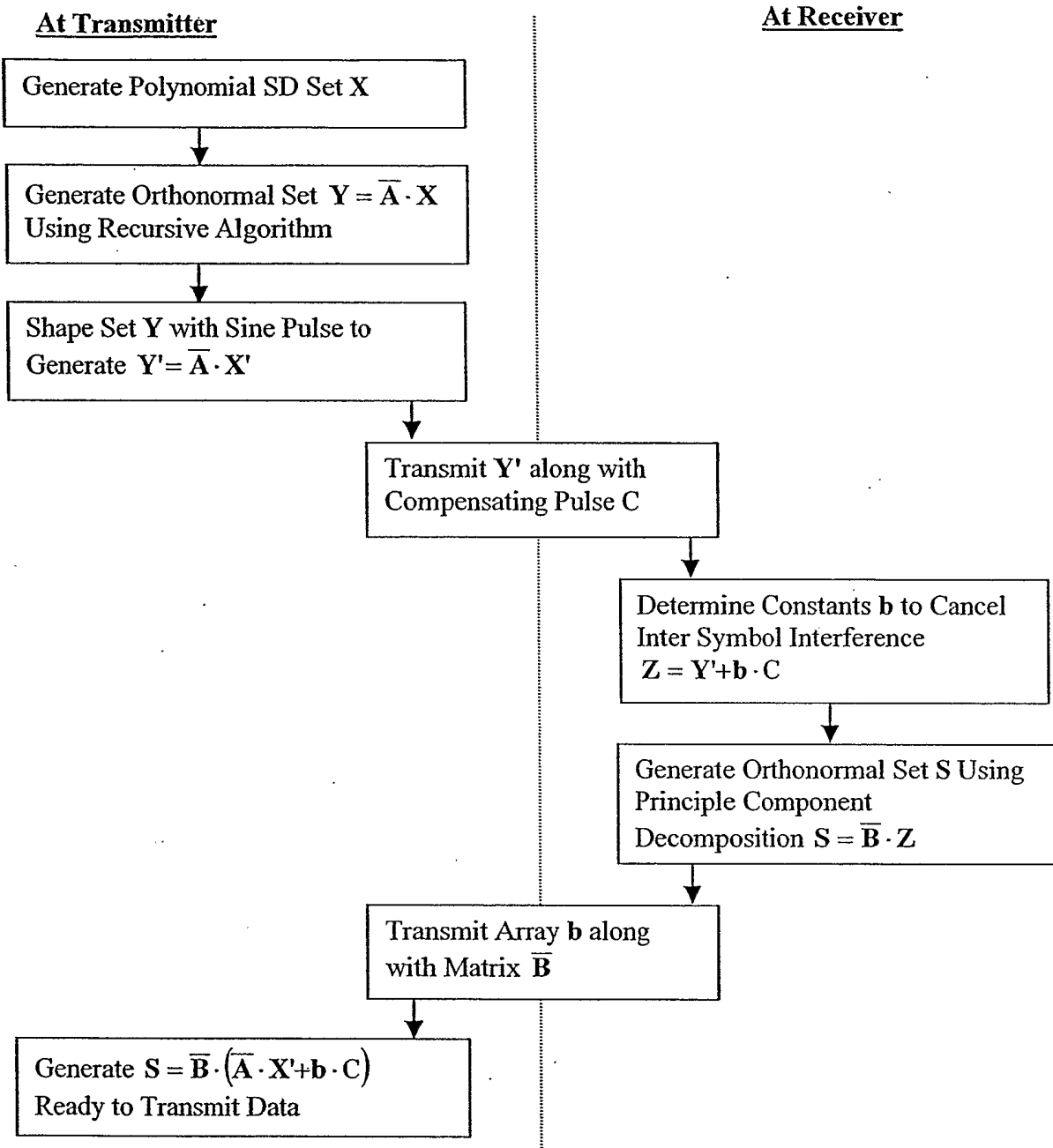


FIG. 50

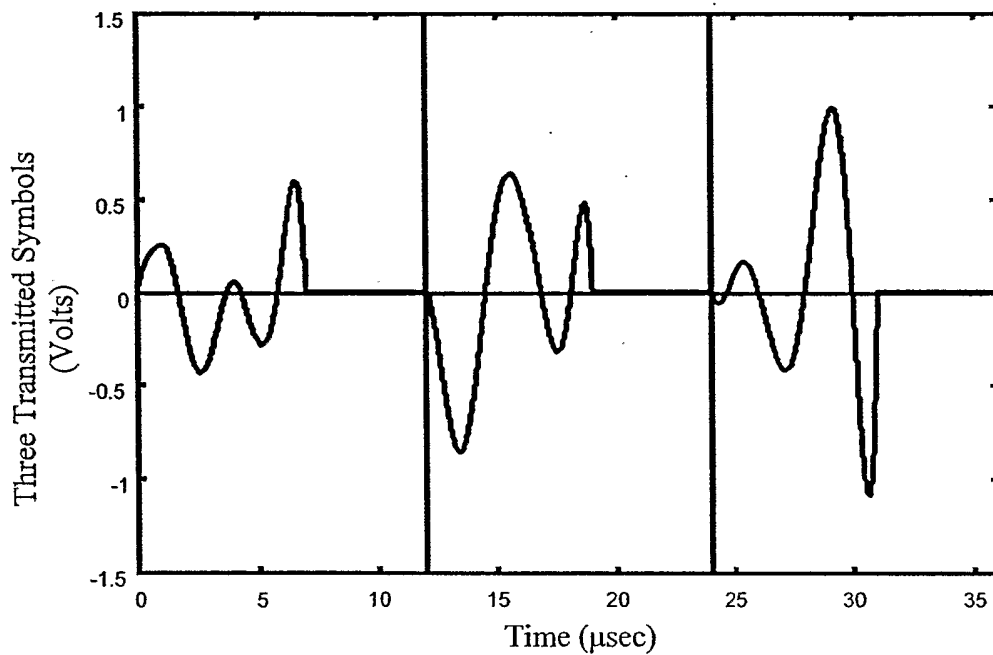


FIG. 51

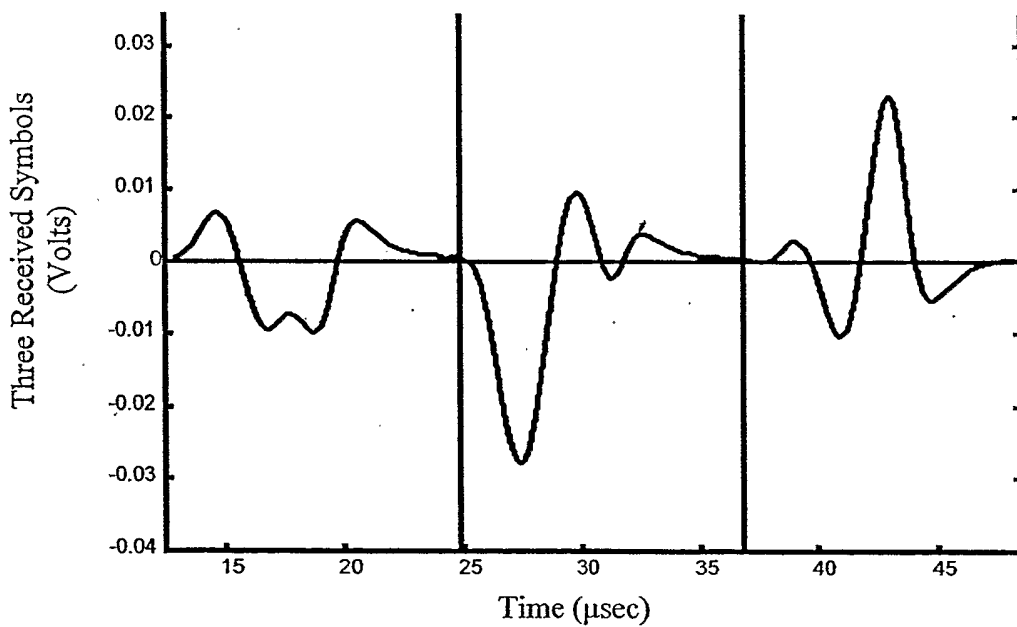


FIG. 52

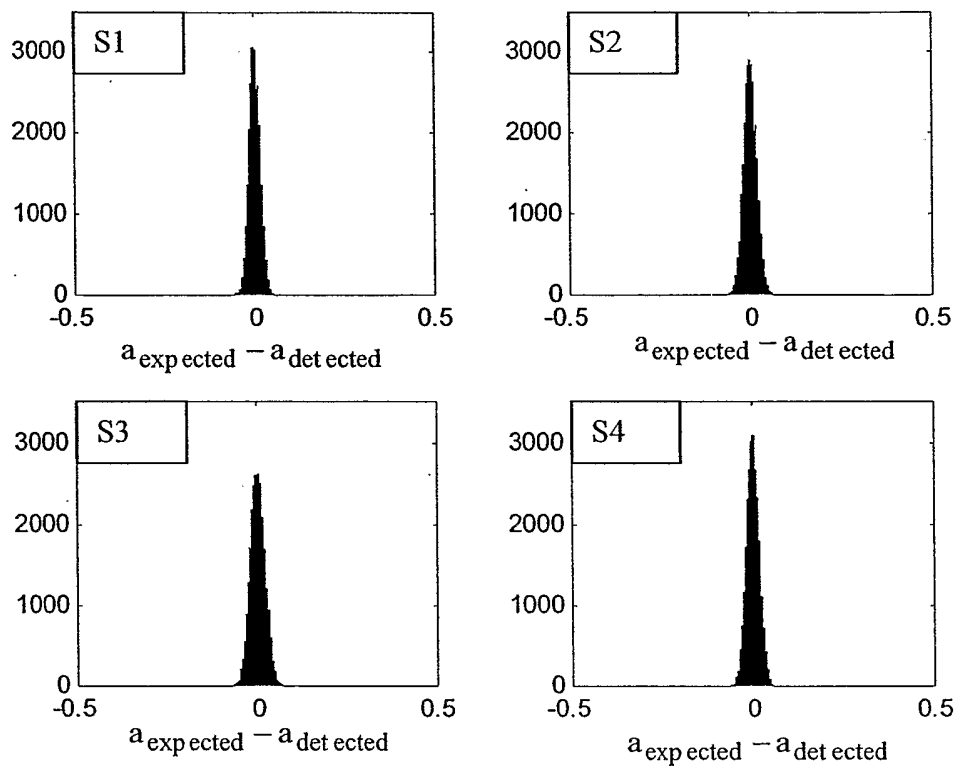


FIG. 53