Measurement and causal modelling of twisted pair copper cables

Jason J. Yoho1 | Sedki M. Riad2 | Ali H. Muqaibel3

1 Hyberlabs inc, 13830 SW Rawhide Ct, Beaverton, Oregon, USA
2 The Bradley Department of Electrical and Computer Engineering, Virginia Polytechnic Institute and State University, Blacksburg, Virginia, USA
3 Electrical Engineering Department, King Fahd University of Petroleum and Minerals (KFUPM), Dhahran, Saudi Arabia

Correspondence
Ali H. Muqaibel, Electrical Engineering Department, King Fahd University of Petroleum & Minerals (KFUPM), Dhahran 31261, Saudi Arabia. Email: muqaibel@kfupm.edu.sa

Abstract
The modelling of twisted pair copper cables has been under investigation in recent years with the emergence of several forms of digital subscriber line broadband access technology including G.fast. Designers of these communication systems implement these models to prototype early system architectures as well as simulate feasibility trials on the existing subscriber loop infrastructure. Previous attempts at modelling these cables have been limited because of the challenges associated with acquiring physical measurement and the inaccuracies of simulators to model effects such as the skin effect, which is significant when using these cables in high-speed systems. These previous modelling attempts employed empirical formulas to match the measured primary transmission line parameters (RLCG) of the cables and generally failed to meet the Kramer–Kronig causality relationships. This paper presents a measurement setup for the electrical properties of twisted pair copper cables as well as a realizable RLCG model for the cables. This model is based on electromagnetic theory considering the model causality, the skin effect, and the twisting of the cable.

1 | INTRODUCTION

The demand for high data rate access rapidly grew from Digital Subscriber Line (xDSL) to G.fast technology to newer standards from the International Telecommunication Union targeting 1 Gb/s over short copper loops [1]. Due to their many advantages related to security and robustness, cabled system will always coexist with wireless systems. Different broadband access technologies have a high degree of spectrum overlap with home networking technologies. The characteristics of the cables determines the electromagnetic coupling between other technologies and twisted pairs. Accurate evaluation of the impact in terms of both average and worst-case impact on data rate requires accurate modelling [2] to predict crosstalk within a bundle of twisted-wire pairs [3]. Other important applications which utilize twisted pair cables include smart vehicles and aircraft systems [4]. These new technologies require accurate parametric cable models for simulation, design, and performance evaluation tests.

Recent research in twisted pairs modelling include the work of Gustavsen et al. to include the 3-D effects imposed by twisting wire screens in cable impedance calculations [3]. Accounting for wire coating in the modelling of field coupling to twisted-wire pairs was performed in [6]. Modelling of twisted pairs usually assumes ideal condition of lossless and uniform media, ignoring the length of the twisted section, these non-idealities were addressed by Sun et al. [7]. A new crosstalk estimation method for non-uniform pitch twisted pair was recently proposed in [8]. Current models of twisted pair cables used in high data rate technologies include practical modelling of some form of measured data [9, 10]. In spite the fact that these empirical line parameters formulations of the cable can be employed in simulations to calculate frequency characteristics like transfer functions and crosstalk, they fall short to represent physically realizable formulas and generally ignore model causality [1]. A system where the impulse response has considerable energy during the propagation delay is non-causal. Accurate models also allow for accurate localization and characterization of wiring faults [11]. Considering the difficulties of performing measurements on these cables at low frequencies, the empirical models also include the inaccuracies of fitting the models to questionable measurements results obtained at low frequencies [12]. Since cables measurements at high frequency are considered the most accurate, resulting from the lower bandwidth limitations of the measurement equipment, employing a broadband, physically realizable model to the high frequency measurements would yield a model that covers the whole spectrum.
This paper develops physically realistic models of twisted pair cables that consist of the geometric dimensions and the material parameters of the cable. The discussion begins with a presentation on the high frequency approximation of the RLCG parameters for two-wire transmission lines. The frequency limitation of this model originates on the assumption that the diameter of the wires is great compared to the skin depth, which is not valid at lower frequencies. As a result, the broadband distribution of the current stemming from the skin effect in a round wire are then developed and incorporated into the model. The integration of the twisting effects of the cables within the model is accomplished through typical representations of helical structures.

A broadband measurement setup to obtain the electrical properties of twisted pair copper cable is presented. This measurement setup implements a technique of terminating the cable with an open and a short circuit to obtain the RLCG parameters of the cable. Following the cable measurements, the resulting data is used to verify the cable modelling technique presented in the paper with applications in both the time and frequency domains.

2 | HIGH FREQUENCY TWO-WIRE MODEL

The high frequency RLCG approximation of the two-wire transmission line in a homogeneous dielectric [13] is presented in this section as a basis for the twisted pair model design. The inhomogeneous dielectric of the two-wire topology in Figure 1 is taken into account by the introduction of an effective complex permittivity. Modifications of this models is made in the following sections to produce a broadband twisted pair cable model. The per unit length series resistance for the two-wire line in a homogeneous dielectric is derived as follows:

\[
R(\omega) = \frac{2R_s}{\pi d} \cdot \frac{D}{d} \cdot \left[\left(\frac{D}{d}\right)^2 - 1\right]^{-\frac{1}{2}}, \quad \text{where} \quad R_s = \frac{1}{\sigma \delta s} = \left(\frac{\omega \mu_0}{\sigma \delta s}\right). \quad (1)
\]

\(D\) is the distance between the axis of the wires and \(d\) their diameter. \(\sigma\) is the conductivity and \(\mu\) is the permeability of the wire material, \(\delta_s\) is the skin depth, and \(\omega\) is the angular frequency.

The series inductance per unit length is given by

\[
L(\omega) = L_s + \frac{\mu_0}{\pi} \cos^{-1} \left(\frac{D}{d}\right), \quad (2)
\]

where \(L_s = \frac{R_s}{\omega}\) is the internal inductance owing to the penetration of the magnetic field into the conductor according to the skin effect, and \(\mu_0\) is the permeability of free space. The shunt conductance per unit length is

\[
G(\omega) = \frac{\omega \varepsilon' \pi}{\cosh^{-1} \left(\frac{D}{d}\right)} \quad (3)
\]

with \(\varepsilon'' = \sigma_d/(\omega \varepsilon_0)\), and \(\sigma_d\) is the conductivity of the dielectric of the cable and \(\varepsilon_0\) the permittivity of free space. Finally,

\[
C(\omega) = \frac{\varepsilon'(\omega) \pi}{\cosh^{-1} \left(\frac{D}{d}\right)} \quad (4)
\]

describes the shunt capacitance per unit length. \(\varepsilon'\) is the real part of the complex permittivity \(\varepsilon = \varepsilon' - j\varepsilon''\) and is related to the relative permittivity \(\varepsilon_r\) of the dielectric by \(\varepsilon' = \varepsilon_r\varepsilon_0\).

3 | BROADBAND TWISTED PAIR CABLE MODEL

The broadband current distribution for a round wire needs to be developed [14] to extend the present equations for the high frequency two-wire topology into the lower frequencies. Following this investigation, a presentation of the effect of twisting the parallel wires will be discussed to relate the parallel two-wire line model to the twisted pair cables.

To analysis the current density distribution for round wires with good conductors, it is noted displacement current density for good conductors is insignificant compared to the conduction current density, resulting in the following Maxwell’s equation in phasor form:

\[
\nabla \times \vec{H} = \jmath = \sigma \vec{E}. \quad (5)
\]

In phasor form, Faraday’s law is given by:

\[
\nabla \times \vec{E} = -j\omega \mu \vec{H}. \quad (6)
\]

From Equations (5) and (6) the differential equation for the current density is given as:

\[
\nabla^2 \vec{j} = -j\omega \mu \sigma \vec{j}. \quad (7)
\]

The \(\zeta\)-axis of the cylindrical coordinate system \((r, \varphi, \zeta)\) is located on the symmetry axis of the infinitely long round wire. Because of the symmetry of the wire, only the \(\zeta\)-component of the current density has to be taken into account which
where the boundary condition of the magnetic field for

tance are given by:

\[ \frac{d^2 J_z}{dr^2} + \frac{1}{r} \frac{dJ_z}{dr} + T^2 J_z = 0, \text{ where } T = \mu \omega \sigma = \frac{1}{2} \sqrt{\frac{2}{\delta}}. \]  

(8)

The result in Equation (8) is the Bessel differential equation
and its general solution is:

\[ J_z = AJ_0(Tr) + BY_0(Tr). \]  

(9)

\( A \) and \( B \) are integration constants, \( J_z \) is the zeroth order Bessel function of the first kind and \( Y_0 \) is the zeroth order Bessel function of the second kind. \( B = 0 \) because \( Y_0 \) tends to infinity for \( r = 0 \) whereas \( J_z \) is finite. \( A \) can be determined by use of the boundary condition of the magnetic field for \( r = r_0 = d/2 \) resulting in the following:

\[ J_z = \frac{\sigma E_0}{J_0(T r_0)} J_0(T r). \]  

(10)

It will be handy to break the complex Bessel function into the real part and the imaginary part:

\[ J_0 \left( \sqrt{-\frac{\omega}{\delta}} r \right) = \text{Ber}(r) + j \text{Bei}(r). \]  

(11)

The resulting current density is given by

\[ J_z = \frac{\sigma E_0}{J_0(T r_0)} \text{Ber} \left( \sqrt{-\frac{\omega}{\delta}} r \right) + j \frac{\sigma E_0}{J_0(T r_0)} \text{Bei} \left( \sqrt{-\frac{\omega}{\delta}} r \right). \]  

(12)

The impedance per unit length can be calculated from the electric field intensity at the surface of the wire. \( E(r_0) = J_z(r_0) / \sigma \) and the current distribution in Equations (12) results in the following impedance:

\[ Z_{\text{int}} = R + j \omega L_{\text{int}} = \frac{\sigma R_t}{\sqrt{2\pi r_0}} \left[ \frac{\text{Ber}(q) + j \text{Bei}(q)}{\text{Ber}'(q) + j \text{Bei}'(q)} \right], \]  

(13)

where \( q = \sqrt{\frac{\omega}{\delta}} r_0 \). The resulting resistance and internal inductance are given by

\[ R = \frac{R_t}{\sqrt{2\pi r_0}} \left[ \frac{\text{Ber}(q) \text{Ber}'(q) - \text{Bei}(q) \text{Ber}'(q)}{[\text{Ber}'(q)]^2 + [\text{Bei}'(q)]^2} \right], \]  

(14)

and

\[ L_{\text{int}} = \frac{R_t}{\omega \sqrt{2\pi r_0}} \left[ \frac{\text{Ber}(q) \text{Ber}'(q) + \text{Bei}(q) \text{Bei}'(q)}{[\text{Ber}'(q)]^2 + [\text{Bei}'(q)]^2} \right]. \]  

(15)

The next step in the developing the cable model is the incorporation of the twisting effects into the model. As in a similar

method used to model helical antenna structures [6], we investigate the presentation of the self-inductance of a circular conducting wire loop. The external inductance of the loop shown in Figure 2 is given by [15]:

\[ L_{\text{coil}_{\text{ext}}} = \mu_0 \frac{D^2}{8\pi} \frac{\ln(\frac{8D}{d} - 2)}{\int_0^2 \frac{1}{2} \cos(\theta_1 - \theta_0) \theta \theta_0 \theta_0 \theta_0 \cos(\theta_1 - \theta_0) \theta \theta_0 \theta_0 \theta_0 \cos(\theta_1 - \theta_0)} \]  

(16)

Through a change of variables and the use of elliptic integrals, the following result is obtained:

\[ L_{\text{coil}_{\text{ext}}} = \frac{\mu_0D^2}{2} \left( \ln \frac{8D}{d} - 2 \right). \]  

(17)

Through the use of Equation (17) and the assumption that the turns of the helix are distant enough apart so that the distribution of the current in the interior of the conductor is not effected by the twisting, the inductance equations will increase by the product of the number of turns per unit length and the loop inductance in Equation (17).

The remaining investigation to complete the twisted pair model is the relationship between the real part, \( \epsilon_r' \), and the imaginary part, \( \epsilon_r'' \), of the complex permittivity. Looking at the Kramers–Kronig relationships for the complex permittivity gives the following relationships [16]:

\[ \epsilon_r'(\omega) = \epsilon_r''(\infty) + \frac{2\pi}{P} \int_0^\infty \frac{\epsilon_r''(\xi) \xi}{\xi^2 + \omega^2} d\xi, \]  

(18)

\[ \epsilon_r''(\omega) = \frac{2\pi}{P} \int_0^\infty \frac{\epsilon_r''(\xi) \xi - \epsilon_r''(\infty)}{\xi^2 + \omega^2} d\xi, \]  

(19)

where \( \epsilon_r''(\infty) \) is the relative permittivity at infinite frequency. The measurements described in the following section show that...
the series resistive component dominates the loss in the transmission line model, which resulting in the loss caused by the shunt conductance to be undetectable. These investigations have shown that the shunt conductance parameter is negligible; therefore, forcing the imaginary part of the permittivity in Equation (19) and with it the conductance to zero results in a constant value for the real part of the permittivity in Equation (18).

The resulting shunt conductance and capacitance of the transmission line model resulting from the two-wire line equations with regards to the Kramers–Kronig relationships are as follows:

\[ C = \frac{\pi \varepsilon''}{\cosh^{-1} \left( \frac{D}{d} \right)} = \text{Constant and} \]

\[ G = \frac{\pi \omega \varepsilon''}{\cosh^{-1} \left( \frac{D}{d} \right)} = 0. \tag{21} \]

4 TWISTED PAIR CABLE MODEL

**SUMMARY**

The per unit length series resistance of the transmission line model was established and is given by:

\[ R_s = \frac{2R_s}{\sqrt{2\pi} \varepsilon_0} \left[ \frac{\text{Ber}(q)\text{Ber}'(q) - \text{Bei}(q)\text{Bei}'(q)}{\text{Ber}'(q)^2 + \text{Bei}'(q)^2} \right] \frac{D}{d}, \tag{22} \]

where

\[ R_s = \frac{1}{\delta_{\sigma}} = \sqrt{\frac{\omega \mu}{2\sigma}} \qquad q = \sqrt{\frac{2\pi}{\delta_{\sigma}}} \qquad r_0 = \frac{d}{2} \quad \text{and} \quad \delta_{\sigma} = \sqrt{\frac{2}{\omega \mu \sigma}}. \]

\(r_0\) = wire conductor radius (m) 
\(d\) = wire conductor diameter (m) 
\(D\) = wire diameter including dielectric material (m) 
\(\sigma\) = conductivity of the metallic conductor (S/m) 
\(\mu\) = permeability of the conductor material (H/m) 
\(\omega\) = angular frequency (rad/s)

The series inductance per unit length of the transmission line model was developed is:

\[ L_s = L_{rod} + L_{int} + nL_{coil} \tag{23} \]

with

\[ L_{rod} = \frac{\mu \delta}{\pi} \ln \left( \frac{8D}{d} \right) - 2, \tag{24} \]

\[ L_{int} = \frac{2R_s}{\omega \sqrt{2\pi} \varepsilon_0} \left[ \frac{\text{Ber}(q)\text{Ber}'(q) + \text{Bei}(q)\text{Bei}'(q)}{\text{Ber}'(q)^2 + \text{Bei}'(q)^2} \right] \frac{D}{d}, \tag{25} \]

where \(n\) is the number of turns per unit length. Mutual inductance and crosstalk analysis showed that the crosstalk between cable pairs had little effect on the parameter measurements. The capacitance per unit length is assumed to be a constant over the required bandwidth which results in the next equation:

\[ C = \frac{\pi \varepsilon'_S}{\cosh^{-1} \left( \frac{D}{d} \right)} = \text{Constant} \tag{27} \]

where \(\varepsilon'_S\) is the real part of the effective permittivity describing the inhomogeneous dielectric of the cable.

\[ G = \frac{\pi \omega \varepsilon''_S}{\cosh^{-1} \left( \frac{D}{d} \right)} = 0 \tag{28} \]

describes the conductance per unit length, where \(\varepsilon''_S\) is the imaginary part of the effective permittivity.

5 TWISTED PAIR CABLE MEASUREMENTS

This section summarizes the procedures and techniques used to model the SEALPIC® twisted-pair communication cables produced by Superior Essex. The primary line parameters (RLCG per unit length) were extracted from single-port measurements performed in frequency domain. The measurements were completed on 22 AWG and 24 AWG cables in the frequency band from 100 kHz to 10 MHz using the Keysight 4195A network analyser. The bandwidth limitation of the Keysight impedance test set (41951-61001) dictated the lower frequency limit of the measurements.

These single-port impedance measurements are performed over a given frequency range. Two measurements with different terminations for the tested are required to extract the RLCG parameters: an open-circuited and a short-circuited termination impedance measurement. The numerical techniques to obtain the line parameters from these two measurements will be described in detail along with the measurement system.

The single-port measurement setup and the cable under test are depicted in Figure 3. The network analyser is linked to balun through an impedance test set using a section of 50 \(\Omega\) semi-ridged coaxial cable. In the single-port measurements, the balun transformer is utilized in both the calibration and measurement stages as a changeover between the unbalanced coaxial line and the balanced two-wire cable. The balun transformer also transforms the characteristic impedance of the coaxial line to the characteristic impedance of a two-wire line of 100 \(\Omega\). To match the twisted-pair cable with a
characteristic impedance of about 120–100 Ω two-wire line, two high precision 10 Ω resistors are connected between the balun transformer and the cable under test as shown in Figure 3.

The used single-port full calibration option involves short, open, and load measurement standards. To reduce most effects due to the connections and the small mismatches with the balun, the calibration procedure is conducted at the reference plane of the measurement setup. The calibration is also preformed over the desired frequency band from 100 kHz to 10 MHz.

As mentioned above, the cable under test is connected to the two-wire line at the reference plane and to open- and short-circuited loads at the termination plane. The two tests comprising of the short- circuited termination impedance measurement and the open-circuited termination impedance measurement are executed resulting in the impedances \( Z_o \) and \( Z_{oc} \), respectively.

\( Z_o \) and \( Z_{oc} \) are used to compute the characteristic impedance, \( Z_o \), and the propagation constant, \( \gamma \), by solving Equations (29) and (30) simultaneously for each frequency, where \( \ell \) is the length of the cable under test.

\[
Z_{oc} = Z_o \tanh (\gamma \ell) \quad \text{(29)}
\]

\[
Z_{sc} = Z_o \coth (\gamma \ell) = \frac{Z_o}{\tanh(\gamma \ell)} \quad \text{(30)}
\]

Following these computations, the primary transmission line parameters can be calculated as follows:

\[
R = \text{Re}\{Z \gamma\} \quad \text{(31)}
\]

\[
L = \frac{1}{\omega} \text{Im}\{Z \gamma\} \quad \text{(32)}
\]

\[
G = \text{Re} \left\{ \frac{\gamma}{Z_o} \right\} \quad \text{(33)}
\]

\[
C = \frac{1}{\omega} \text{Im} \left\{ \frac{\gamma}{Z_o} \right\} \quad \text{(34)}
\]

Note that the LCR meter, the impedance analyser, and the vector network analyser are different type of measurement instruments with different theory of operation and hardware characteristics. These instruments have different upper limitation of measurement frequency range. It is shown that all instruments provide almost the same measurement performance in the range of 1 MHz up to 100 MHz [17].

### 6 MEASUREMENT RESULTS AND MODEL COMPARISON

Comparisons of the developed models with the model presented in the ANSI T1E1.4/96 [9] fit to our measured data are shown graphically in this section. The relationships of the models in both the time and frequency domain are presented.

The two models were fit to the measured obtained by the methods described earlier in this paper. For a 3 m section of 24-AWG cable, the models are contrasted as in Figure 4. A problem with resonances occurs at longer cable lengths within the frequency range of interest. For the model to avoid structural behaviour, multiple cable lengths between 1.8 and 12.8 m were initially measured to be able to piecwise the data together surrounding the different resonances. After this investigation, it was found that the shorter length cables give accurate low frequency results (\( \sim 100 \) kHz) and a single length cable was chosen for all measurements. The provided per unit length parameters allow the model to be use for extended range of lengths. The length of 3 m was selected since the resonance frequency is higher than the frequency band of interest (\( f_r > 10 \) MHz) and the low frequency results are analogous to longer cables. The low frequency measurements are not available due to the low frequency constraints of the measurement setup. Therefore, the two models are not the same in fitting the low frequency data. The propagation constants and characteristic impedances generated by the two models are displayed in Figure 5 and show excellent correlation, except for the real part of the propagation constant, \( \alpha \). It is this difference that makes our model causal. In the ANSI T1E1.4/96, the value of \( \alpha \) increases exponentially with frequency. The discrepancy of the created model to the data reported in the ANSI standard can be due to few reasons. One prospect is that the data documented in the standard is not applicable over the entire frequency range because of the equipment bandwidth and accuracy limitations which complicates the measurements of the low frequencies’ primary transmission line parameters.

Typical model parameters of the devised models for the twisted pair lines of 22 and 24 AWG cables are given in Table 1. Measured physical parameters of these cables are as follows: (22 AWG) \( d = 5.080 \times 10^{-4} \) m, \( D = 8.890 \times 10^{-4} \) m; (24 AWG) \( d = 6.350 \times 10^{-4} \) m, \( D = 10.9220 \times 10^{-4} \) m; Typical value for the Conductivity of copper is \( \sigma = 5.813 \times 10^7 \) S/m. The measured data were averaged over multiple measurements. There is agreement between the measured and typical distance between the axis of the wires. For the specific used cables, there is relatively less agreement when we compare the values for the diameters.

The resulting number of turns per unit length is overestimated for the 24 AWG cable to obtain proper correlation; this
is most likely due to measurement difficulties or approximations tolerances of the higher gauge cables.

To analyse the sources of errors in the measurements, three full measurements (including calibration) were performed at different time intervals. The insignificant disagreements in the results, assured the repeatability of the measurement. When measuring the length of the cable under test, two potential causes of mistakes exist. First, differences between conductor pairs are possibly due to twisting rates and placement within the conducting shield. The resulting error is proportional to the twisting rate. The second error source is caused by the length estimation of the cut leads including the soldering and connectors. When cables are carefully prepared for measurements, the second error source would be consistent for the various cable pairs. A total of Six twisted pairs having a length of 31.63 m were measured. Due to twisting rate or placement within the cable, it is demonstrated that within a certain section of the cable, the length of the conductor pairs can be of slightly different. This length evaluation is shown to be a very credible explanation for some of the measurement variations.

In some situations, a non-causal model is not a problem. some results can be acquired by frequency-domain simulations, such as data rate estimations via channel capacity equations that depend on channel transfer function and power spectral densities. However, in other cases based on time-domain analysis, e.g., evaluation of the cyclic-prefix duration, noncausal models should be avoided as they can hinder the results interpretation [1]. Also, the use of Time Domain Reflectometry (TDR) to estimate the network topology requires comparing the measured response with the modelled response. In such a case the non-causality would result in wrong network reconstruction. It is realized that significant differences in the characteristic impedances cause large fluctuations in a time domain reflectometry (TDR) response [18]. Though, the characteristic impedances computed from each of the models are very comparable resulting in the inference that the non-causality of the empirical model is the reason of the differences in the TDR responses generated by the two models, depicted in Figure 6. The presented TDR responses was generated using a developed Matlab code that implement the presented model.
TABLE 1  Typical parameters for 22 AWG and 24 AWG cables

| Parameter | 24 AWG cable | 2 AWG cable |
|-----------|--------------|-------------|
| $d$ (m)   | $5.4459 \times 10^{-4}$ | $6.6472 \times 10^{-4}$ |
| $D$ (m)   | $8.8901 \times 10^{-4}$ | $10.9699 \times 10^{-4}$ |
| $\sigma$ (S/m) | $5.9225 \times 10^7$ | $5.8034 \times 10^7$ |
| $n$ (turns/m) | $171.60$ | $8.45$ |
| $\varepsilon_{\text{eff}}$ | $1.7823$ | $2.0326$ |
| $\varepsilon_{\text{eff}}$ | $0$ | $0$ |

The time domain simulation was obtained over a 40 $\mu$s sampling duration for a sampling frequency of 40 MHz. The input pulse was a 10 V raised-cosine pulse with a 2 $\mu$s duration having a source impedance of 120 $\Omega$. The pulse was input into a 1000 m segment of a 50 $\Omega$ lossless line followed by a 300 m segment of the modelled transmission line. Additionally, network loop nodes are assumed to be open-circuited [19].

7. CONCLUSION

This paper summarized the development of a physically realizable modelling approach for twisted pair copper cables. The model is based on electromagnetic theory and take into account the twisting effects, the skin effect, and the causality of the cable modelling. This model is more robust than the previous modelling approaches that used empirical models fit to questionable broadband RLCG measurement data of cables.

This paper also summarized the procedures used to model the SEALPIC twisted-pair communication cables 22 and 24 AWG manufactured by Superior Essex. The primary line parameters (RLCG) were obtained from frequency domain measurements on the cables.

Comparing the non-causal ANSI T1E1.4 model and our model demonstrated that the non-causality of the ANSI T1E1.4 models resulted in cumulative errors over the whole TDR response. Actual measurement are causal, so the models used to correlate the system for performance analysis or loop identification also need to be causal. The developed model characterizes...
FIGURE 6  TDR simulation of an example xDSL subscriber line using the ANSI T1E1.4 models and the models presented in this publication. The magnified window depicts the causality of the presented models

the cables over a sufficient wideband frequency range to obtain an accurate and causal TDR response.

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