Twisted Pair Cable Design Analysis and Simulation

Mohammed M Al-Asadi[!], Alistair P Duffy[!], Kenneth G Hodge[*] and Arthur J Willis[*]

[!]Department of Engineering and Technology, De Montfort University. Leicester, LE1 9BH UK.

[*]Brand-Rex Ltd., Glenrothes, Fife, KY6 2RS, Scotland, UK.

Principle Contact: email alasadi@dmu.ac.uk

Abstract

In this paper, a family of equations has been developed to describe the behaviour of twisted pair cables as functions of cable dimensions, basic material parameters and frequency of operation. These equations allow the prediction of secondary parameters without the need to extrapolate from measurements. Further a simulation method has been developed which requires dimensions, materials and frequency only and enables investigation of the behaviour of the cable under the influence of periodic, sporadic or random deformities, such as those arising through poor installation practice or production process variability. Both analytical and simulation approaches complement the more historical approach of iterative development and practical extrapolation. The set of equation derived and the simulation approach were then used to predict the performance of a twisted pair cable where the dimensions and the materials are known. The results obtained were then validated against measurements. Excellent agreement is in all cases demonstrated.

1. Introduction

Data communication is of increasing importance to society, with the information revolution showing no sign of slowing. Applications such as the Internet and mobile communications are becoming an integral part of everyday life. A common component in all the applications is the communication channel. This may be transmission line and cable, optical fibre, wave-guide, 'free space' propagation, etc. This paper is concerned with twisted pair cables.

Twisted pair cables have been in use for a long time. Early telephone signals were sent over twisted pair cable and now virtually all networked buildings use twisted pair cable for telephone and computer communication signals. Despite the fact that coaxial cables and optical fibre cables were developed to handle higher frequency signals with minimal interference, twisted pair cables have been developed so that they can carry increasingly higher frequency signals; The proposed category 8 cable will be specified to 1.2 *GHz*.

Different varieties of twisted pair cables are available for different needs and applications. First is the basic telephone cable, which is known as direct-inside wire (DIW), is still in use in some countries. Developments over the years, such as improvements in the individual wire sheaths and variations in the twist have led to the development of standard-compliant Category 3 UTP cables specified for frequencies up to 16 *MHz*. Further improvement led to the development of Category 4 for specifications up to 20 *MHz* and Category 5 for specifications up to 100 *MHz*. The next generations of Category 6 and Category 7 cables are now emerging in the open market. The specifications of those Categories reach 200 *MHz* and 600 *MHz* respectively. Category 6 cable can be, and Category 7 cable is, shielded.

With the emergence of new technologies in high frequency communication system, the employment of multi-conductor twisted pair cable has increased. The primary parameters, R, L, G and C, for a such cables at high frequencies, are not difficult to calculate [1]. The calculation of the secondary parameters, such as attenuation constant α, the phase constant β, the propagation constant γ, and the characteristic impedance Z_o, relies on the calculation of the primary parameters [1,2].

Cable manufacturers and design engineers are keen to understand the behaviour of communication cables, specifically twisted pair cables, at high frequencies. They need to understand the effects of changing the dimensions of the cable and the material, particularly the dielectric, on cable performance at the frequency of operation. Therefore there is a need for a set of design equations to relate all the secondary parameters to the zeroth order parameters: cable dimensions, material parameters and the frequency of operation. Hence, the first aim of this paper is to develop a family of equations to describe the behaviour of twisted pair cables as function of cable dimensions, basic material parameters and frequency of operation.

For a regular communication channel, where the cable is uniform and the matching between the communication channel components is perfect, the calculation of secondary parameters can be obtained using the developed set of equations. But what will happen when the cable used suffers an irregularity ? And what would be the way of measuring or calculating the secondary parameters of cable affected by such variations along itself or by miss-handling the cable properly? This paper addresses cable irregularities and the calculation of secondary parameters. The answer could be difficult to find using measurements at higher frequencies [3]. Hence, the need for further analytical development and the development of a modelling technique, capable of tackling the problems arising in testing cables at very high frequencies.

The second approach illustrated in this paper is the development of a computer-modelling tool, using the Transmission Line Matrix, TLM, method [4], capable of simulating the behaviour of both regular and irregular twisted pair cables. The model is able to incorporate other parameters of the communication channel, such as connectors and different termination, to analyse their contribution to the performance of the communication channel. The next section includes the derivation of the secondary parameter equations of a twisted pair cable.

2. Secondary Parameters Derivation

This section presents the derivation of equation for the calculation of attenuation constant, phase constant and hence propagation constant for unshielded twisted pair cable (UTP). It also presents the derivation of an equation for the calculation of the attenuation constant of a shielded twisted pair cable (STP). Equations for the calculation of both amplitude and phase of the characteristic impedance of UTP cables are also derived and presented in this section.

2.1 UTP Cables Attenuation Constant Calculation

The attenuation constant (α_u), can be obtained by calculating the real part of the propagation constant (γ), where γ is given by [1]:

$$\gamma = \alpha_u + j\beta_u \tag{1}$$

and β_u is the phase constant. Letter u is used to differentiate between the unshielded formula and the shielded formula presented later (where letter s is used). The propagation constant is also given as:

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} \tag{2}$$

Where R, L, G and C are the distributed primary parameters per unit length of one segment of a single pair cable and given as [4]:

$$R = \frac{2R_s}{\pi \ d}\left[\frac{D/d}{\sqrt{(D/d)^2 - 1}}\right] \quad (\Omega/\mathrm{m}) \tag{3}$$

$$L = \frac{\mu_d}{\pi}\cosh^{-1}\left(\frac{D}{d}\right) \quad (H/m) \tag{4}$$

$$G = \frac{\pi \ \sigma_d}{\cosh^{-1}\left(\frac{D}{d}\right)} \quad (S/m) \tag{5}$$

$$C = \frac{\pi\varepsilon}{\cosh^{-1}\left(\frac{D}{d}\right)} \quad (F/m) \qquad 6)$$

where D is the distance between the centres of the two conductors of the cable, d is the diameter of each conductor and R_s is the surface resistance of the conductor. For low loss cables at relatively high frequency, it can be assumed that $R \ll \omega L$. Using the binomial expansion of [5]:

$$(1+b)^{\pm\frac{1}{2}} = 1 \pm \frac{b}{2} \qquad \text{For} \qquad b \ll 1$$

Equation .2 can be simplified to the following form:

$$\gamma = \frac{1}{2}\left(R\sqrt{\frac{C}{L}} + G\sqrt{\frac{L}{C}}\right) + j\left(\omega\sqrt{LC} - \frac{RG}{4\omega\sqrt{LC}}\right) \tag{7}$$

Comparing the above equation with equation 1, the attenuation constant can be given as:

$$\alpha = \frac{1}{2}\left(R\sqrt{\frac{C}{L}} + G\sqrt{\frac{L}{C}}\right) \tag{8}$$

Substituting the values of the primary parameters given in equations 3→ 6 in the last equation with some mathematical simplifications gives:

$$\alpha = \frac{1}{2}\left[\frac{2R_s}{\pi d} \times \left(\frac{D/d}{\sqrt{(D/d)^2 - 1}}\right) \times \frac{\pi}{\cosh^{-1}(D/d)}\sqrt{\frac{\varepsilon}{\mu_d}} + \sigma_d\sqrt{\frac{\mu d}{\varepsilon}}\right] \tag{9}$$

The surface resistivity for any material can be given as ($R_s = R_k \cdot f^{1/2}$), where R_k is a constant related to the conductivity of the conductor material, σ, and the skin depth, δ. For example, for copper conductors, the constant R_k is given in reference [1] as $2.61x10^{-7}$. Substituting R_s into equation 9, and further simplifying, gives:

$$\alpha = \frac{R_k f^{1/2}}{d\cosh^{-1}(D/d)}\left(\frac{D/d}{\sqrt{(D/d)^2 - 1}}\right)\sqrt{\frac{\varepsilon}{\mu_d}} + \frac{\sigma_d}{2}\sqrt{\frac{\mu_d}{\varepsilon}} \tag{10}$$

It is easier to visualise 'cause and effect' if the equation is a function of the dielectric thickness around each conductor and the conductor diameter. Therefore, if τ is the thickness of the dielectric then, for the two-wire cable, $D = 2\tau + d$. Hence, the term (D/d) in the last equation can be written in terms of the dielectric thickness and the conductor's diameter as ($D/d = 2\tau/d + 1$) Meanwhile, the conductivity of the dielectric can be given as [1] ($\sigma_d = \omega\varepsilon \ = 2\pi f\varepsilon$). To obtain the attenuation constant in decibels, the last equation should be multiplied in 8.686 [1]. Hence equation 10 can be further simplified to:

$$\alpha_u = \left\{\frac{8.686R_k}{d\cosh^{-1}\left(\frac{2\tau}{d}+1\right)}\left(\frac{\left(\frac{2\tau}{d}+1\right)}{\sqrt{\left(\frac{2\tau}{d}+1\right)^2 - 1}}\right)\sqrt{\frac{\varepsilon}{\mu_d}}\right\}f^{1/2} + \left\{8.686\pi\varepsilon''\sqrt{\frac{\mu_d}{\varepsilon}}\right\}f \tag{11}$$

Assuming that:

$$\alpha_{ua} = \left\{\frac{8.686R_k}{d\cosh^{-1}\left(\frac{2\tau}{d}+1\right)}\left(\frac{\left(\frac{2\tau}{d}+1\right)}{\sqrt{\left(\frac{2\tau}{d}+1\right)^2 - 1}}\right)\sqrt{\frac{\varepsilon}{\mu_d}}\right\} \tag{12}$$

and

$$\alpha_{ub} = \left\{8.686\pi\varepsilon''\sqrt{\frac{\mu_d}{\varepsilon}}\right\} \tag{13}$$

The attenuation can be written in a simple form as:

$$\alpha_u = \alpha_{ua} f^{1/2} + \alpha_{ub} f \tag{14}$$

It is clear that α_{ua} is mostly a dimension dependant factor while α_{ub} is a material dependant factor. For a copper cable where D = 0.94 *mm* and d = 0.53 *mm*, and attenuation is in *dB*/100*m* and frequency is in *MHz*, both factors given in equations 13 and 14, are calculated and found as; α_{ua} = 1.7259, α_{ub} = 0.01453. Equation 14 is similar to the equation commonly used by manufacturers, which is in the following form [6]:

$$\alpha_u = af^{1/2} + bf + cf^{-1/2} \tag{15}$$

The terms a, b and c of equation 15 are reported as [6]; a = 1.967, b = 0.023 and c = 0.1. It can be seen that for the high frequencies, the 3rd term of equation 15 will be very small and can be ignored. However, it does not appear in equation 14, which was obtained through direct analysis (using high frequency assumption).

2.2 UTP Cables Phase Constant Calculation

Following the same procedure, comparing equations 1 and 7, the phase constant of the UTP cable can be obtained as:

$$\beta = \left[\omega\sqrt{LC} - \frac{RG}{4\omega\sqrt{LC}}\right] \tag{16}$$

For a low loss cable, the phase constant may be represented by the 1st term of the above equation; this is common practice [1]. But, for accurate calculations, for any cable and over a large range of frequencies. It is more appropriate to use equation 16. Again, substituting equations 3$\rightarrow$ 6 in equation 16 and with some mathematical simplification the phase constant can be obtained as:

$$\beta = \beta_a f^{1/2} + \beta_b f \tag{17}$$

where:

$$\beta_a = \frac{-R_k \varepsilon''}{2d\sqrt{\mu_d \varepsilon}\cosh^{-1}\left(\frac{2\tau}{d}+1\right)}\left[\frac{\frac{2\tau}{d}+1}{\sqrt{\left(\frac{2\tau}{d}+1\right)^2 - 1}}\right] \tag{18}$$

and

$$\beta_b = 2\pi\sqrt{\mu_d \varepsilon} \tag{19}$$

It is clear that equation 17 is obtained in a similar form to that of equation 14 as a function of cable dimensions, material properties and frequency

2.3 UTP Cables Propagation constant calculation

For the unshielded twisted pair cable, substituting equations 14 and 17 in equation 1, a simple formula for the propagation constant can be obtained as:

$$\gamma = (\alpha_a + j\beta_a) f^{1/2} + (\alpha_b + j\beta_b) f \tag{20}$$

which can again be rewritten as:

$$\gamma = \gamma_a f^{1/2} + \gamma_b f \tag{21}$$

where,

$$\gamma_a = \alpha_a + j\beta_a \tag{22}$$

and

$$\gamma_a = \alpha_b + j\beta_b \tag{23}$$

It can be seen that as both attenuation and phase constants are functions of cable dimensions, material properties and frequency, subsequently, the propagation constant is also obtained as a function of the cable dimensions, materials properties and frequency. The next secondary parameter discussed is the characteristic impedance of the cable that is analysed in the following sub-section.

2.4 UTP Cables Characteritic Impedance calculation

It is well known that the characteristic impedance, Z_o can be given as [1]:

$$Z_o = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \tag{24}$$

As it is a complex quantity, both impedance amplitude and phase are derived as functions of cable dimensions, materials properties and frequency of operation.

2.4.1 Amplitude Calculations:

Squaring both sides of equation 24 and taking the magnitude gives:

$$\left|Z_o^2\right| = \left(\frac{R^2 + \omega^2 L^2}{G^2 + \omega^2 C^2}\right)^{0.5} \tag{25}$$

Therefore the magnitude of the characteristic impedance Z_o can be given as:

$$\left|Z_o\right| = \left(\frac{R^2 + \omega^2 L^2}{G^2 + \omega^2 C^2}\right)^{0.25} \tag{26}$$

Using equations 3 $\rightarrow$ 6, and replacing σ_d by ($\sigma_d = \omega\varepsilon'' = 2\pi f\varepsilon''$), the above equation can be mathematically simplified as a function of cable dimensions, material properties and frequency to:

$$\left|Z_o\right| = \frac{\sqrt{\left(\cosh^{-1}\left(1 + 2\tau/d\right)\right)}}{\pi} \cdot \left(\frac{\frac{\left(2.61\times10^{-7}\right)^2 (2\tau + d)^2}{4\pi^2 d^2 f^{1.75}\tau \; (\tau + d)} + \mu^2\left(\cosh^{-1}\left(1 + 2\tau/d\right)\right)^2}{\left(\varepsilon''^2 + \varepsilon^2\right)}\right)^{0.25} \tag{27}$$

To enhance clarity equation 27 can be rearranged in the form:

$$|Z_o| = a\left(\frac{b+c}{e}\right)^{0.25} \quad (28)$$

where:

$a = \dfrac{\sqrt{\left(\cosh^{-1}\left(1+2\tau/d\right)\right)}}{\pi}$ is a function of dimensions,

$b = \dfrac{\left(2.61\times10^{-7}\right)^2 (2\tau+d)^2}{4\pi^2 d^2 f^{1.75}\tau \ (\tau+d)}$ is a function of dimensions and frequency, but note the 10^{-14} term,

$c = \mu^2\left(\cosh^{-1}\left(1+2\tau/d\right)\right)^2$ is a function of materials and dimensions, and

$e = \left(\varepsilon''^2 + \varepsilon^2\right)$ is a function of material properties only.

At high frequencies, assuming that $b << c$ and $\varepsilon'' << \varepsilon$ then equation 27 can be simplified as:

$$|Z_o| = \frac{\sqrt{\left(\cosh^{-1}\left(1+2\tau/d\right)\right)}}{\pi}\left(\frac{\mu^2\left(\cosh^{-1}\left(1+2\tau/d\right)\right)^2}{\varepsilon^2}\right)^{0.25}$$

$$= \frac{\sqrt{\left(\cosh^{-1}\left(1+2\tau/d\right)\right)}}{\pi}\sqrt{\frac{\mu\left(\cosh^{-1}\left(1+2\tau/d\right)\right)}{\varepsilon}}$$

Hence, at high frequencies, the magnitude of the characteristic impedance of the cable may be simplified to the following form:

$$|Z_o| = \frac{\cosh^{-1}\left(1+2\tau/d\right)}{\pi}\sqrt{\frac{\mu}{\varepsilon}} \quad (29)$$

Equation 29 is similar to the equation reported by [1] at high frequencies.

2.4.2 Phase Calculations: Again, squaring both sides of equation 24 and taking the phase gives:

$$\angle Z_o^2 = \tan^{-1}\left(\frac{\omega \ L}{R}\right) - \tan^{-1}\left(\frac{\omega \ C}{G}\right) \quad (30)$$

Using de Moivre's theorem [5], the phase constant of the characteristic impedance, Z_o, is given as:

$$\angle Z_o = \frac{1}{2}\left(\tan^{-1}\left(\frac{\omega \ L}{R}\right) - \tan^{-1}\left(\frac{\omega \ C}{G}\right)\right) \quad (31)$$

which can be rearranged to [5]:

$$\angle Z_o = \frac{1}{2}\left(\tan^{-1}\left(\frac{\omega \ L}{R}\right) + \tan^{-1}\left(-\frac{\omega \ C}{G}\right)\right) \quad (32)$$

Since [5]:

$$\tan^{-1}(\phi) + \tan^{-1}(\vartheta) = \left(\tan^{-1}\left(\frac{\phi+\vartheta}{1-\phi\vartheta}\right)\right) + \theta$$

where:

$$\theta = \begin{cases} \pi, & \phi\vartheta > 1, \phi > 0 \\ 0, & \phi\vartheta < 1 \\ -\pi, & \phi\vartheta > 1, \phi < 0 \end{cases}$$

Therefore equation 32 can be rewritten as:

$$\angle Z_o = \frac{1}{2}\left(\tan^{-1}\left(\frac{\frac{\omega \ L}{R} - \frac{\omega \ C}{G}}{1 + \frac{\omega \ L}{R}\cdot\frac{\omega \ C}{G}}\right) + \theta\right)$$

Since

$$\frac{\omega \ L}{R} - \frac{\omega \ C}{G} < 1$$

Therefore θ of the above equation goes to zero and hence, the phase of the characteristic impedance can be simplified to:

$$\angle Z_o = \frac{1}{2}\left(\tan^{-1}\left(\frac{\omega \ LG - \omega \ RC}{RG + \omega^2 LC}\right)\right)$$

Once again, substituting equations 3→6 in the above equation, the phase constant of the characteristic impedance will be given in terms of cable dimensions, materials and frequency, after some mathematical simplifications, in the following form:

$$\angle Z_o = \frac{1}{2}\tan^{-1}\left(\frac{a_\theta \varepsilon'' f^{1/2} - b_\theta \varepsilon}{a_\theta \varepsilon \ \ f^{1/2} + b_\theta \varepsilon''}\right) \quad (33)$$

where:

$$a_\theta = \pi\mu_d \quad (34)$$

And

$$b_\theta = \frac{2.61\times10^{-7}}{d\cosh^{1/2}(1+2\tau/d)}\left(\frac{\left(1+2\tau/d\right)}{\sqrt{4\tau/d\left(1+\tau/d\right)}}\right) \quad (35)$$

Equations 34 and 35 demonstrate that a_θ is a function of the cable materials and b_θ is a function of its dimensions, therefore equation 33 is a function of materials, dimension and frequency as required

At this stage, the only secondary parameter that is not discussed here is the return loss. This is investigated in detail in an associated paper [7]. The next subsection illustrates the derivation of a similar equation for the calculation of the attenuation constant of a shielded twisted pair cable (STP).

2.5 STP Cables Attenuation Constant Calculation

For this analysis, it will be assumed that the shield is a perfect cylinder. In this case, the inner diameter of the shield D_S, is twice the distance between the centres of the conductors, (i.e. $D_S = 2D$). The attenuation constant due to the conductors and the shielding material, α_c, is given by [1] as:

$$\alpha_c = \frac{R}{2Z_O} \tag{36}$$

R, in the above equation, is the resistance per unit length, Ω/m, and is given as [1]:

$$R = \frac{2R_{S2}}{\pi d}\left[1 + \frac{1+2p^2}{4p^4}\left(1-4q^2\right)\right] + \frac{8R_{S1}}{\pi D_S}q^2\left[1+q^2-\frac{1+4p^2}{8p^4}\right] \tag{37}$$

where $p = D/d$ and $q = D/D_S$. As $D = d+2\tau$, p and q can be written as:

$$p = \frac{d+2\tau}{d} = 1 + \frac{2\tau}{d} \tag{38}$$

$$q = \frac{D}{D_S} = \frac{D}{2D} = 0.5 \tag{39}$$

For a copper conductor, the surface resistivity is $R_{S2} = 2.61\times10^{-7} f^{1/2}$, and for an Aluminium shield, the surface resistivity is $R_{S1} = 3.26\times10^{-7} f^{1/2}$. However, to make the design equations more general, the surface resistivity for the conductors is taken as $R_{S2} = R_{k2} f^{1/2}$,and for the shield as $R_{S1} = R_{k1} f^{1/2}$. R_{k1} and R_{k2} are constants related to the conductivity, σ, and the skin depth, δ, of both the conductors and the shield respectively. Substituting equations 38 and 39 in equation 37, the resistance R will be given as a function of the conductor diameter d, the dielectric thickness τ, and the frequency of operation f. Equation 37 may now be further simplified to the following form:

$$R = \frac{2}{\pi}\left[\frac{R_{k2}}{d} + \frac{R_{k1}}{2(d+2\tau)}\left(1.25 - \frac{1+4\left(1+\frac{2\tau}{d}\right)^2}{8\left(1+\frac{2\tau}{d}\right)^4}\right)\right] f^{1/2} \tag{40}$$

The characteristic impedance Z_o, of equation 36 is given as [1]:

$$Z_O = \frac{\eta}{\pi}\left[\ln\left\{2p\left(\frac{1-q^2}{1+q^2}\right)\right\} - \frac{1+4p^2}{16p^4}\left(1-4q^2\right)\right] \tag{41}$$

Using the dielectric factor $\eta = \sqrt{\mu_d/\varepsilon}$ and equations 38 and 39, equation 41 can be further simplified to:

$$Z_O = \frac{1}{\pi}\sqrt{\frac{\mu_d}{\varepsilon}}\left[\ln\left\{1.2\left(1+\frac{2\tau}{d}\right)\right\}\right] \tag{42}$$

Using the values of R and Z_o, given in equations 40 and 42 respectively, the attenuation due to the conductor given in equation 36, can then be written as:

$$\alpha_c = \frac{\frac{2}{\pi}\left[\frac{R_{k2}}{d} + \frac{R_{k1}}{2(d+2\tau)}\left(1.25 - \frac{1+4\left(1+\frac{2\tau}{d}\right)^2}{8\left(1+\frac{2\tau}{d}\right)^4}\right)\right] f^{1/2}}{\frac{2}{\pi}\sqrt{\frac{\mu_d}{\varepsilon}}\left[\ln\left\{1.2\left(1+\frac{2\tau}{d}\right)\right\}\right]} \tag{43}$$

which can be simplified to the following form:

$$\alpha_c = \frac{\left[\frac{R_{k2}}{d} + \frac{R_{k1}}{2(d+2\tau)}\left(1.25 - \frac{1+4\left(1+\frac{2\tau}{d}\right)^2}{8\left(1+\frac{2\tau}{d}\right)^4}\right)\right]}{\left[\ln\left\{1.2\left(1+\frac{2\tau}{d}\right)\right\}\right]}\sqrt{\frac{\varepsilon}{\mu_d}} f^{1/2} \tag{44}$$

The attenuation due to the dielectric α_d, is also given by [1] as:

$$\alpha_d = \frac{\sigma\eta}{2}$$

Substituting both σ and η, by their values, the last equation can be simplified to the following form:

$$\alpha_d = \frac{\varepsilon''\omega}{2}\sqrt{\frac{\mu_d}{\varepsilon}} = \pi\varepsilon''\sqrt{\frac{\mu_d}{\varepsilon}} f \tag{45}$$

Using equations 44 and 45 and simplifying, the total attenuation α_s in dB/m can be given as:

$$\alpha_s = 8.686\left\{\frac{\left[\frac{R_{k2}}{d} + \frac{R_{k1}}{2(d+2\tau)}\left(1.25 - \frac{1+4\left(1+\frac{2\tau}{d}\right)^2}{8\left(1+\frac{2\tau}{d}\right)^4}\right)\right]}{\left[\ln\left\{1.2\left(1+\frac{2\tau}{d}\right)\right\}\right]}\sqrt{\frac{\varepsilon}{\mu_d}}\right\} f^{1/2} + \left\{8.686\pi\varepsilon''\sqrt{\frac{\mu_d}{\varepsilon}}\right\} f \tag{46}$$

Given that:

$$\alpha_{sa} = 8.686\left\{ \frac{\left[\left[\frac{R_{k2}}{d} + \frac{R_{k1}}{2(d+2\tau)}\right]\left(1.25 - \frac{1+4\left(1+\frac{2\tau}{d}\right)^2}{8\left(1+\frac{2\tau}{d}\right)^4}\right)\right]}{\left[\ln\left\{1.2\left(1+\frac{2\tau}{d}\right)\right\}\right]}\sqrt{\frac{\varepsilon}{\mu_d}} \right\} \tag{47}$$

and

$$\alpha_{sb} = \left\{8.686\,\pi\varepsilon''\sqrt{\frac{\mu_d}{\varepsilon}}\right\} \tag{48}$$

Equation 46 can now be simplified and written in the following form:

$$\alpha_s = \alpha_{sa}\, f^{1/2} + \alpha_{sb}\, f \tag{49}$$

This is similar to equation 14 developed earlier for the calculation of the attenuation constant for unshielded twisted pair cable. It should be noted that the f term α_{sb} is exactly equivalent to the f term of a UTP cable in equation 14. The only variation will arise from different effective permittivity due to presence of the shield. In here, an equation for the calculation of the attenuation constant of a shielded cable has been developed. A set of similar design equations for the calculations of the rest of the secondary parameters can also be developed in.

The next section of this paper illustrates the development of a modelling approach for the simulation of both regular and irregular twisted pair cables (UTP), which can incorporate other communication channel components such as connectors into the model. The modelling method used is the Transmission Line Matrix (TLM) modelling method.

3. TLM Model of a UTP cable

The Transmission Line Matrix (TLM) modelling method is a time domain numerical analysis technique that was originally developed at Nottingham University in the early 1970s [8]. It is based on the representation of the modelled physical phenomena by a network of transmission lines. Any model developed using a TLM approach consists of many transmission-line segments connected together [4]. For the purpose of this paper, TLM may be understood via a description of a twisted pair cable model. One elemental section (of length Δl) of a lossy, single pair cable is shown in figure 1.

As described in [9] and [10], the combination of L and C can be replaced by a link line of impedance ($Z_o=(L/C)^{1/2}$), and the time step for a wave to propagate one TLM node is ($\Delta t=(L.C)^{1/2}$). R and G, as they are not time dependent parameters, are left in the TLM model without change. For a regular cable, Z_o would replace the LC combination at every segment along the cable. If the cable is not regular, then L and C are not the same along the cable, hence to keep a constant time step, Δt, for iterative TLM method, the minimum inductance, L_{min} and minimum capacitance C_{min} are computed and combined in a link line of impedance ($Z_{min}=(L_{min}/C_{min})^{1/2}$).The remaining parts of both L and C at each node are calculated and replaced by stub transmission line models, having their impedances as in [9,10]. The time step in this case is calculated as ($\Delta t_{min}=(L_{min}.C_{min})^{1/2}$).

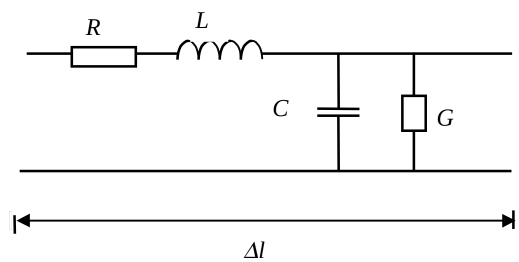

Figure 1. One segment of transmission line showing its primary parameters

The TLM model of two neighbouring segments can then be obtained and their Thevenin equivalent circuit as in figure 2.

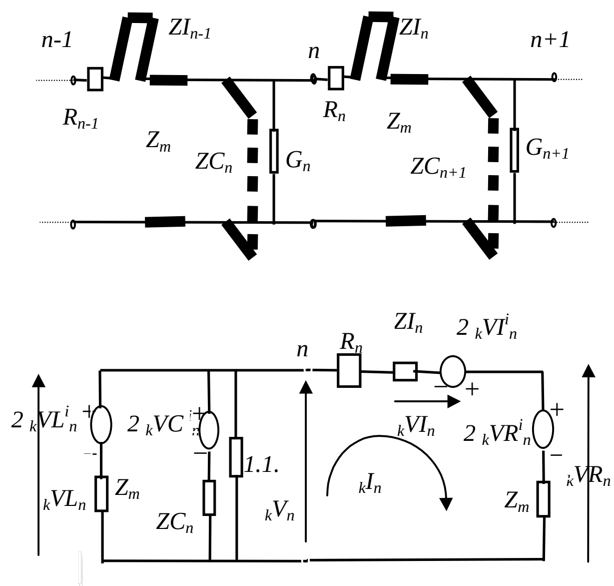

Figure 2. TLM model of two segments of the cable and its Thevenin equivalent circuit.

The variation of both L and C may come about due to regular or random variations in the distance between the centres of the conductors (changing the thickness of the dielectric or dielectric non-uniformity). The benefit of the developed TLM approach is that the stubs may be ignored in the case of modelling a regular cable. This process reduces the overall computing time required to run the TLM program. Hence the same TLM model can actually be used for the simulation of both regular (uniform) and irregular (non-uniform) twisted pair cables. It should be mentioned that the twist length is taken into account by calculating the actual length of the cable to be used for the calculation of the number of the TLM nodes (segments) that represent the whole cable.

Using the parallel generator theory, the nodal voltage of the above circuit can be obtained as:

$$ {}_kV_n = \frac{\dfrac{2{}_kVL_n^i}{Z_m} + \dfrac{2{}_kVC_n^i}{ZC_n} + \dfrac{2{}_kVR_n^i - 2{}_kVI_n^i}{R + Z_m + ZI_n}}{\dfrac{1}{Z_m} + \dfrac{1}{ZC_n} + \dfrac{1}{R + Z_m + ZI_n} + G_n} \qquad (50) $$

Using the process described in [8,9], incident and reflected voltages on, and from, the stubs and the TLM links can be computed and used in the iterative solution of the wave propagation along the cable. Further details about the TLM method can be obtained from [4] and about the twisted pair cable model from [9,10].

In order to compute the secondary parameters of the cable, namely attenuation and return loss, S-parameters of the cable should be obtained. Knowing both near and far end incident and reflected voltages of the TLM model of the cable in the time domain, their values in the frequency domain and at the frequency of operation can be obtained with the use of a Fourier Analysis program.

4. Implementation and Results

The first sub-section validates both approaches against measurements. And the second sub-section is to report the implementation of the TLM model for the calculation of secondary parameters, under different working conditions of the cable.

4.1 Validation

In order to validate the analysis and modelling approaches, the attenuation constant of UTP cable is calculated using equation 14 and computed using the TLM approach where the dimensions of the cable are: D = 0.94 *mm*,d = 0.53 *mm* and the lay length ll = 20 *mm*.

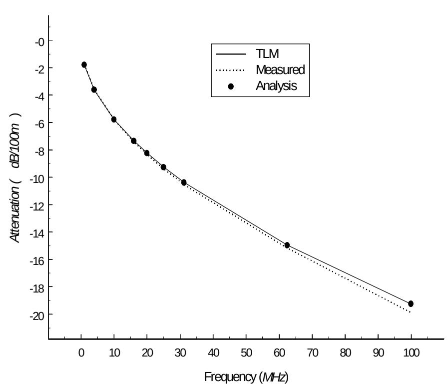

Figure 3. Calculated, modelled and measured attenuation for 100m cable.

Then results obtained using both approaches were plotted along with the measured data for the same cable as function of the frequency as in figure 3.

For the same cable, both phase and propagation constants were calculated using equations 17 and 23 respectively. Both sets of results were validated against standards equations obtained from [1] and plotted as in figures 4 and 5 respectively.

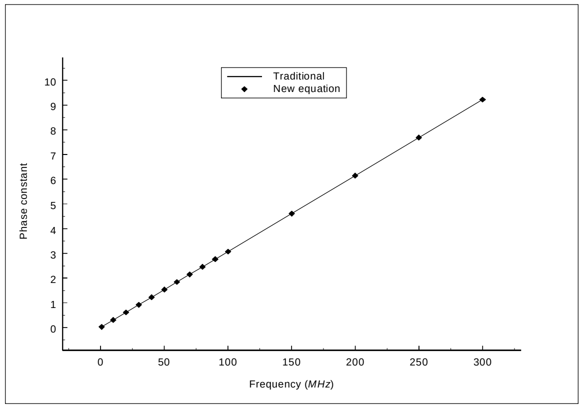

Figure 4. Calculated phase constant as a function of the frequency.

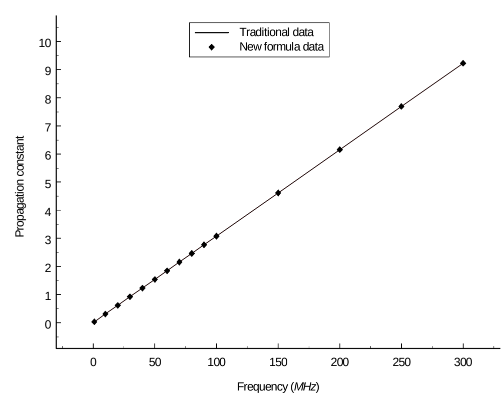

Figure 5. Calculated propagation constant as a function of frequency.

For the same cable, the amplitude of the characteristic impedance is calculated using equation 28 and plotted along with results obtained using the standard equation 24 and plotted as a function of the frequency as in figure 6.

For a shielded twisted pair cable where the cable dimensions are:

D = 1.52 *mm*. d = 0.574 *mm*. ll = 20 *mm*

The screen tape thickness was 0.062 *mm*. It consisted of 2 layers. The outside layer was of 0.05 *mm* aluminium alloy and the inner side layer was of 0.12 *mm* polyester. Equation 49 was used for the calculation of the attenuation constant. Results are plotted along with the measured data as illustrated in figure 7.

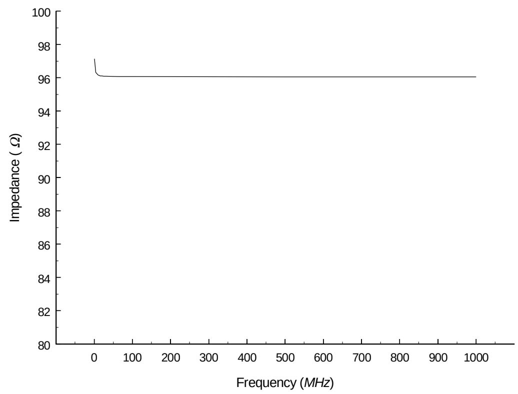

Figure 6. Calculated amplitude of Z_o as a function of frequency.

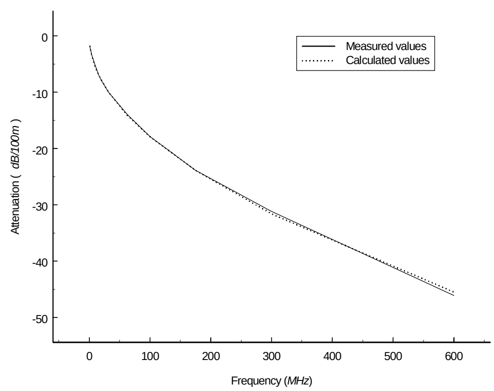

Figure 7. Calculated and measured attenuation constant for STP cable

4.2 TLM model applications

The TLM model described earlier is used for the calculation of both attenuation constant and return loss for the UTP cable investigated in this paper. Results for the attenuation of a regular cable were reported earlier along with those obtained using the design equations for validation purposes. Results for the return loss of a 1m link terminated with a matched load at the far end are illustrated with those obtained for irregular cable for comparison purposes in the following sub-sections. The TLM model is then used to investigate the irregularity effects on both time domain and frequency domain responses of the cable.

4.2.1 Handling effects: For the same category 5 UTP cable investigated here, exhibiting a cyclical variation in the separation of individual wires of the pair by ±100 microns, with five such cycles per meter length. A voltage source of a step wave are applied and the serial resistance of the source is 10 mΩ, the attenuation constant is calculated and plotted as a function of frequency along with the regular cable results as illustrated in figure 8.

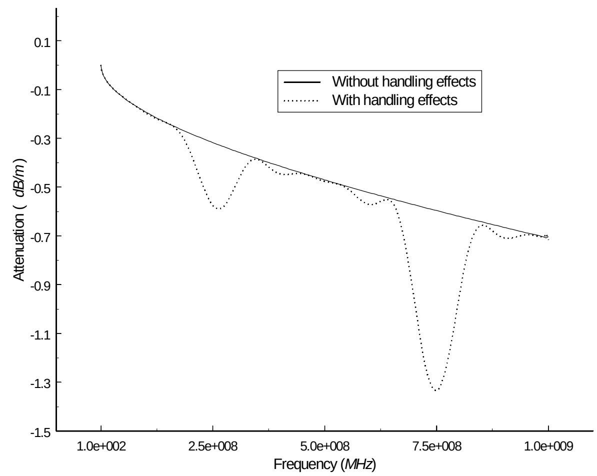

Figure 8. Calculated and measured attenuation for a deformed cable.

The return loss of the cable is also computed for both cases of regular cable and a deformed cable. Both sets of results are plotted in figure 9.

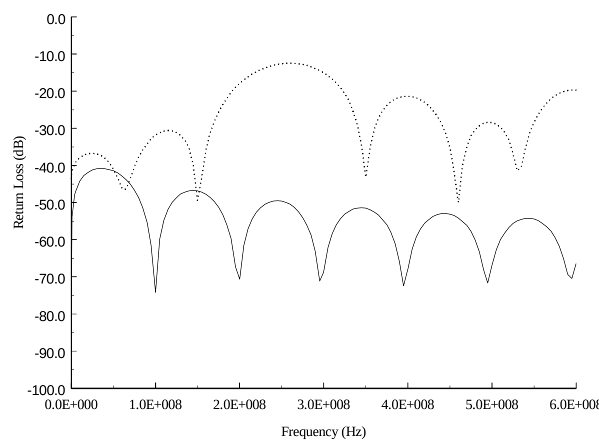

Figure 9. Calculated and measured return loss for a deformed cable.

4.2.2 Pig-tailing effects: Again, both ends of the same cable were connected between a source and a load via a pig-tail connection illustrated in figure 10.

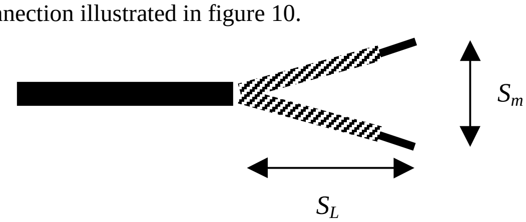

Figure 10. Illustration of pigtail connection.

For a 1m length of cable where S_m = 2.5 *mm* and S_L = 10 *mm*, the attenuation constant was calculated and is plotted as a function of the frequency along with the "no pig-tailing" case. Results are illustrated in figure 11.

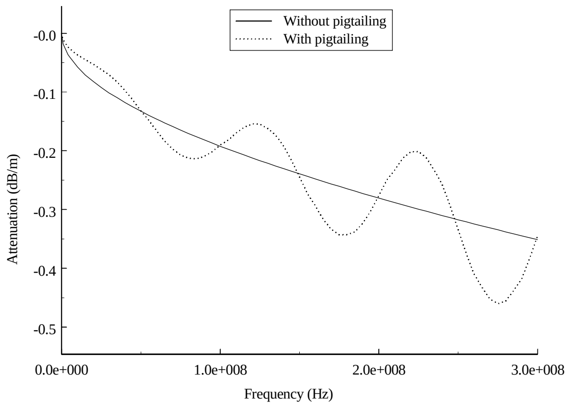

Figure 11.Attenuation computed for a pigtailed and uniform case.

While return loss for both cases are also computed and plotted as illustrated in figure 12

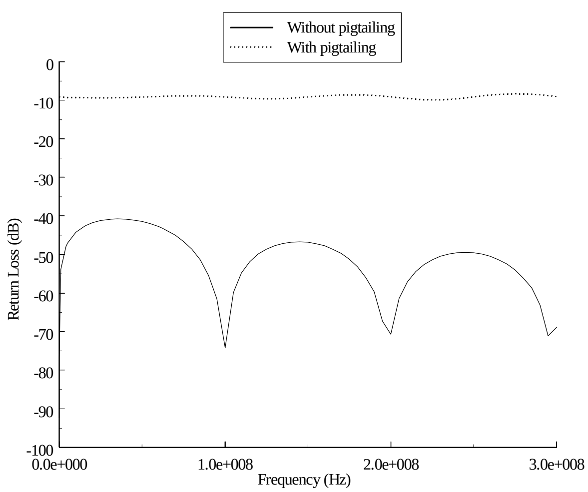

Figure 12.Return loss computed for a pigtailed and uniform case.

5. Conclusions

A family of design equations has been developed to describe the behaviour of twisted pair cables as functions of cable dimensions, materials properties, and frequency of operation. They are validated against measurements and an excellent agreement between both measured and calculated data have been demonstrated.

A TLM model for a twisted pair cable is developed. The model is capable of simulating both regular and irregular cables. It also incorporates the modelling of various communication channels elements, such as connectors. This provides a unique tool for the investigation of the performance of twisted pair channels under different working conditions.

Results obtained for a regular cable, using the TLM model show a very good agreement with both analytical and measured sets of data. The model is then used for the investigation of defected cable performance. Defects represented by changing the distance between the centres of the wires in some places, which provides an impedance discontinuity along the cable. Analytical solutions for such cable performance prediction are very difficult if not impossible to obtain. Results are plotted for both attenuation and return loss along with those for a regular cable. They illustrate the irregularity effects on the obtained results.

Another reason for impedance discontinuity is the 'pig-tailing' connection between both ends of cable and connectors at both source end and load end. Such conditions provide variable impedance values between the nominal impedance of the cable and may be 4 times its value. Again, attenuation and return loss calculations illustrate the fact that a good attention and care should be taken when installing short communication links.

The paper provides a simple set of secondary parameter equations that rely directly on the physical parameters of the cable and hence make them suitable to be used as a reference for future cable developments. It also provides a fast, flexible and accurate approach that can be used for the investigation of cable performance under different working conditions.

6. References

[1] S. Ramo, J. R. Whinnery and T. Van Douzer, *Fields and waves in communication electronics*, 2nd ed., Wiley, New york (1984).

[2] IEC/TC 46C, project number: IEC 61156-1 titled, "Test procedures for determining characteristic impedance (Zo) and return loss (RL)," (March 1996) and its amendment, (October 1997).

[3] Ke Lu, "An efficient method for analysis of arbitrary non-uniform transmission lines," IEEE Transaction on MTT, 45(1), 9-14, (January 1997).

[4] C. Christopoulos, *The Transmission Line Modelling method TLM*, IEEE Press, New York, (1995).

[5] L. Rade and B. Weestergren, *Mathematics Handbook for Science and Engineering*, 3rd edition, studentlitterature, (1995).

[6] Yin-Ying, "UTP cable attenuation constant", an article on a virtual textbook used in EE35 'Telecommunication' posted on the WWW, (14th February 1996), Worcester Polytechnic Institute, Worcester, Massachusetts.

[7] Mohammed M Al-Asadi, Alistair P Duffy, Keeneth G Hodge and Arthur J Willis, “Return Loss Prediction for Cascaded Systems”, paper no. 17-04, IWCS 49th, (November 2000).

[8] P. B. Johns and R. L. Beurle, “Numerical solution of 2-dimensional scattering problems using a transmission line matrix”, Proc. 118(9), 1203-1208, IEE, (September 1971).

[9] M. M. Al-Asadi, A. J. Willis, K. Hodge and A. P. Duffy, “Modelling as a tool for analysing handling effects in structured wire cabling”, 10th IEE Int. Conf. on EMC, 131-136, (September 1997).

[10] M. M. Al-Asadi, A. P. Dufy, A. J. Willis and K. Hodge, “Analysing Link Performance in High Frequency Transmission Systems”, 13th In. Symp. On EMC, Zurich 117-120, (February 1999).

Authors Biographies

Mohammed M. Al-Asadi, Alistair P. Duffy, Kenneth G. Hodge and Arthur J. Willis. For photos and biographies see "Return Loss Prediction For Cascaded Systems", paper 17-04 in these proceedings.