Numerical modelling of the transfer impedances and admittances of a braided coaxial cable

A.Gavrilakis Student Member IEEE
M.M.Al-Asadi Member IEEE
A.P.Duffy Member IEEE
De Montfort University, Leicester, UK

K.G.Hodge **A.J.Willis**
Brand Rex Ltd.
Glenrothes, Fife, Scotland, UK

Principle email contact: alex@dmu.ac.uk

Abstract

This paper presents the numerical simulation of the transfer characteristics of a coaxial cable using the Transmission Line Matrix (TLM) Method. The geometry of a braided shield is examined and the coupling across the braided shield is incorporated into the numerical model. The main advantage of the proposed method is that it can simulate the effect of the length of the cable and electromagnetic interference with spatial variation along the length of the cable shield. The model takes into account any cable terminations and can locate any resonances and standing waves present on the line. Comparison with measurements shows very good agreement.

Keywords

Transfer impedance, braid shield, TLM, coaxial cable.

INTRODUCTION

As technology advances, the role of electromagnetic compatibility becomes more dominant in the design of distributed electronic systems. In high performance applications the leakage of energy into and out of coaxial cables is an important source of interference; it is widely acknowledged that the cables are often the common 'weakest link' of any system [1]. The use of braided shields is very common as they generally provide a high level of shielding, which can be optimized, while retaining mechanical flexibility.

In this paper the transmission-line theory is presented together with a theoretical model for the transfer impedance calculation. Afterwards the TLM theory is presented and issues such as synchronization are discussed. Then, several excitation types that can be applied to the shield are introduced, and the results of the simulations for different termination conditions are presented.

TRANSMISSION LINE THEORY

Transmission-line theory uses lumped elements to characterize the signal propagation on a line. This is valid for short electrical lines where the separation between the conductors is much smaller than the conductors' length.

In this paper the first condition is met by using a suitable step length for the TLM [1,2] and the second condition is usually valid for communication cable measurements. Hence, this method involves representing each transmission line by an equivalent circuit, which takes into account the primary parameters of the lines (e.g. capacitance, inductance etc.).

The equivalent circuit of the system consists of the circuits in Figures 1 and 2. In the case of the shield equivalent circuit, the lumped parameters are; the shield resistance R_s, the shield inductance L_s (which is a result of the balance between the braid and the leakage inductance), and the shield capacitance C_s which is the capacitance between the shield and the ground. The shield conductance G_s is usually omitted as its value is negligible.

In the case of the inner conductor circuit; apart from the line resistance R, inductance L conductance G and capacitance C, there are also present a voltage source V_T and a current source I_T, both describing the transfer characteristics of the shield and obtained directly from the shield model.

Due to Faraday's law,

$$V_T = -Is \cdot Z_T \tag{1}$$

$$I_T = -Vs \cdot Y_T \tag{2}$$

Z_T is calculated here by a theoretical model proposed by Katakis [4] which is a modification of the model proposed by Tyni [5]. The two inductances present are the braid and the leakage inductances. The braid inductance L_b, arises from the woven nature of the braid, and the leakage inductance L_a, is caused by the holes of the braid. These are given by,

$$L_b = \frac{-\mu_o h}{4\pi D_m}\left(1-\tan^2\alpha\right) \tag{3}$$

$$L_a = \frac{\mu_o 2N}{\pi\cos\alpha}\left(\frac{b}{\pi \cdot D_m}\right)^2 \cdot e^{\left(\frac{-\pi \cdot d}{b}-2\right)} \tag{4}$$

$$D_m = D_o + 2d + h \qquad (5)$$

Where α is the braid angle, D_m is the mean braid diameter, D_o is the diameter over the dielectric, N is the total number of belts (spindles), b is the hole width, d is the braid-wire diameter, h is the radial spindle separation and μ_o is the permeability of free space.
The transfer impedance is then obtained from the approximation:

$$Zts \approx j\omega(L_b - L_a) \qquad (6)$$

Y_T is the transfer admittance of the shield [3].
Is is the current flowing on the shield and Vs is the voltage between the shield and the external environment. *Is* may exhibit longitudinal or temporal variation.

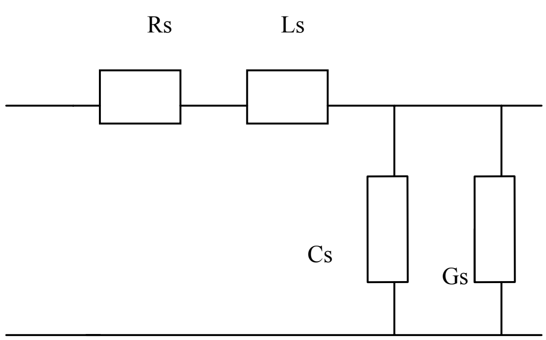

Figure 1. Equivalent circuit of the shield

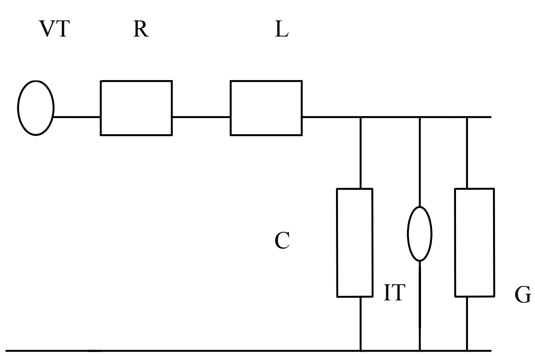

Figure 2. Equivalent circuit of inner conductor

TRANSMISSION LINE MODELLING

The Transmission Line Matrix (TLM) technique is mainly used in 2-D and 3-D applications, however there are a small number of publications for 1-D applications [6,7]. In the method presented in this paper, a 1-D approach is used to minimize computational resource requirements and because geometrical issues, such as dielectric inhomogeneities etc. are not significant for most coaxial cable designs.

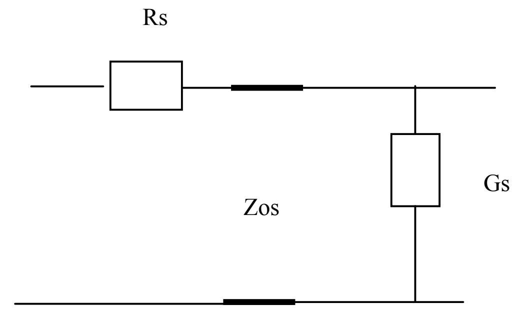

Figure 3. TLM Equivalent circuit -shield

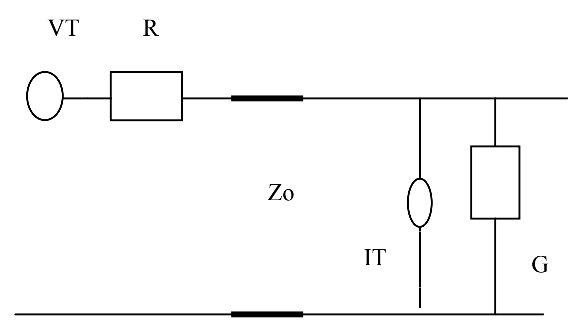

Figure 4. TLM Equivalent circuit- inner conductor

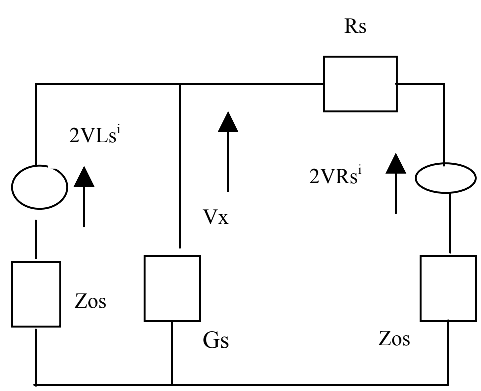

Figure 5. Thevenin equivalent circuit - shield

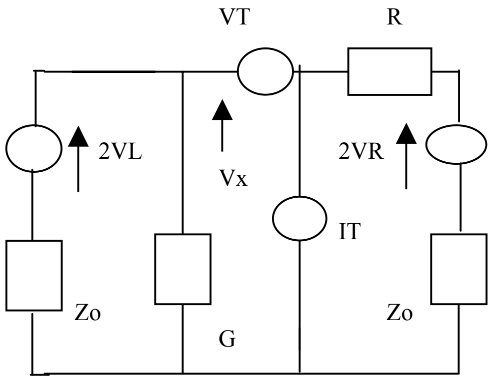

Figure 6. Thevenin equivalent circuit- inner conductor

When the Thevenin equivalent circuits are derived for the inner conductor, the combination of inductance L and capacitance C has been replaced by an equivalent transmission line of $Zo = \sqrt{\frac{L}{C}}$ [8]. The analysis involves splitting the cable into a number of nodes of short length so reflections and transmissions can been accommodated and any time domain behavior can be modeled. The corresponding time step Δt, can be calculated as $\Delta t = \sqrt{L \cdot C}$. In order to achieve time synchronization for the wave propagation between the inner conductor and the shield, the shield characteristic impedance is derived as $Zos = \frac{L_s}{\Delta t}$ where Ls is the inductance of the shield. The shield capacitance is then obtained as $Cs = \frac{\Delta t^2}{L_s}$. Measurements have shown that this calculated value for Cs is very close to the actual value [3,9] for practical cases. If that is not the case then capacitive stubs have to be introduced into the TLM analysis [8].

The voltage at a point x on the shield (Figure 5) is given as,

$$ {}_kVx_n = \frac{\frac{2\,{}_kVLs^i{}_n}{Zos} + \frac{2\,{}_kVRs^i{}_n}{Rs + Zos}}{\frac{1}{Zos} + \frac{1}{Rs + Zos} + Gs} \qquad (6) $$

Respectively, the voltage at a arbitrary point x on the inner conductor (Figure 6) is ,

$$ {}_kVx_n = \frac{{}_kITx \cdot Zo \cdot (Zo + R)}{2 \cdot Zo + R} + \frac{\frac{2\,{}_kVL^i{}_n}{Zo} + \frac{2\,{}_kVR^i{}_n - {}_kVT_n}{R + Zo}}{\frac{1}{Zo} + \frac{1}{R + Zo} + G} \qquad (7) $$

TLM calculates the currents and voltages in each node and then manipulates them through the ‘scatter and connect’ algorithm, the details of which are dependent on the circuit parameters, to obtain values for the next time step.

EXCITATION TYPES

The excitation is applied directly to the shield. The simplest case is a single pulse applied at the near end of the shield, which then propagates along the shield. However in most cases an incident field (e.g. from an radiating antenna, or as a result of cross-talk) is present. In which case, due to Lenz’ law, the incident magnetic field H^i in the z direction and the resulting induced voltage source Vs(x) will be related by [10],

$$ V_s(x) = j\omega\mu_o \int_{y=0}^{s} H^i{}_n dy = -j\omega\mu_o \int_{y=0}^{s} H^i{}_z dy \qquad (8) $$

where s is the distance between the inner conductor and the shield in the y direction. The induced current source $I_s(x)$ due to the transverse electric field E^i will be given as,

$$ I_s(x) = -j\omega c \int_{y=0}^{s} E^i{}_t dy = -j\omega c \int_{y=0}^{s} E^i{}_y dy \qquad (9) $$

However, in the latter excitation case, the formulation of the shield transmission line equations is similar to that of the inner conductor, as the excitation is not applied at one point but across the shield length.

SIMULATIONS

The advantage of this method over purely theoretical approaches is that the model can be used to observe the effects of any terminations (e.g. matched, short, open).

The two different conditions that will be examined are (i) inner conductor matched at both ends with braid short circuited at both ends and (ii) matched both inner conductor and shield.

The selection of the points is such that the voltages and currents reading will be able to determine the surface transfer impedance.

Short-matched case

In first case the shield is shorted at both ends and the inner conductors is terminated into matched loads (50 Ohms for a 1 meter RG58 cable). In Figures 7 and 8 the voltage spectra for the far end of the inner conductor and the current spectra for the near end of the shield are shown respectively. Due to the relative permittivity of the dielectric (er=2.26) the propagation velocity of the inner conductor is 66% of that of free space. Hence, this predicts a spacing of 100MHz between each resonance maxima which can be observed in Figure 7. However, because the shield is shorted, resonances are present on its surface (Figure 8). This phenomenon has been observed as well by Rumold and ter Haselborg [11] in work involving cable bundles. As the voltage present on the shield is due to the transfer characteristics of the shield, then by taking the ratio of the voltage at the far end of the inner conductor to the current flowing on the near end of the shield the transfer impedance illustrated in Figure 9 is produced. It can be clearly seen that the linearity is lost after about 50 MHz as resonances due to the length of the cable are present.

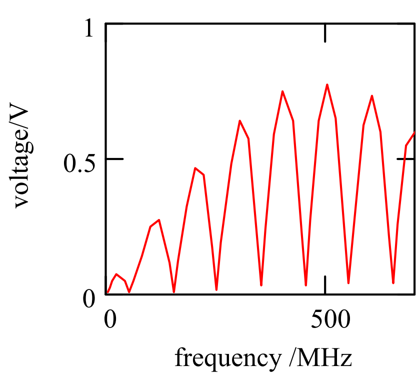

Figure 7. Voltage spectra at the far end of 1m RG58 inner conductor (Shorted shield,matched inner conductor)

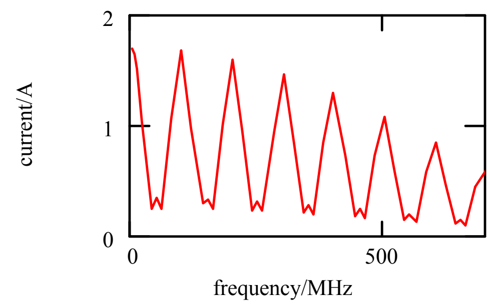

Figure 8. Current spectra at the near end of 1m RG58 shield (Shorted shield, matched inner conductor)

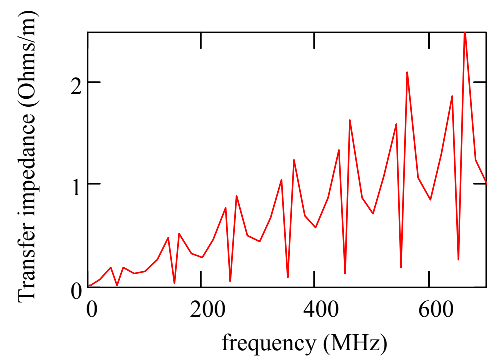

Figure 9. 'Transfer impedance' of 1m RG58 inner conductor (Shorted shield, matched inner conductor)

Match-matched case

In this case the only difference is that the shield is not short circuit anymore but terminated into the same loads as for the inner conductors (e.g. 50 Ohm network analyzer). The respective voltage spectra, current spectra and calculated transfer impedance as shown in Figures 10,11 and 12. As expected the magnitude of the voltage on the inner conductor in the case of Figure 10 is significantly smaller than that in Figure 9 due to the termination of the shield. However the mapping of the resonances is identical as it solely depends on the length of the conductor. However because the shield is terminated on the loads of the inner conductor the current on the near end is attenuated due to losses and mismatches present on the line, as shown in Figure 11.

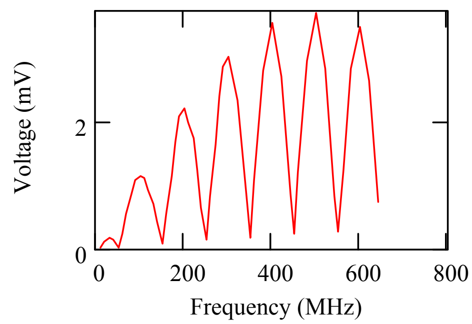

Figure 10. Voltage spectra at the far end of 1m RG58 inner conductor (Matched shield, matched inner conductor)

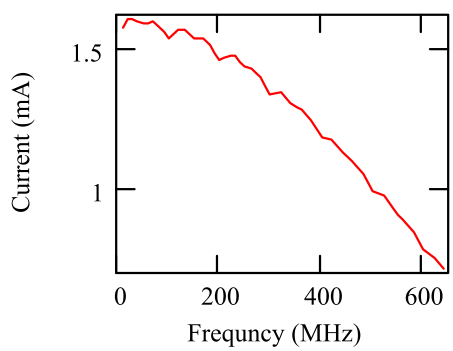

Figure 11. Current spectra at the near end of 1m RG58 shield (Shorted shield, matched inner conductor)

The next step was to compare this method with a measuring technique using a similar geometry [12] to extract the transfer impedance. It can be noticed from Figure 12, that the agreement of the maxima of the resonances is very good and a further validation of the theoretical prediction using Katakis' model has been included. It can been seen that the theoretical model can only describe the surface transfer impedance as a linear function of frequency and is unable to locate any resonances due to the length of the cable. On the other hand the magnitude agreement between theory, simulation and practice is good, which is encouraging.

Lastly, no open circuit analysis is presented here, because the introduction of the internal impedances of the measuring equipment (e.g. 50 Ohms for typical network analyzer) makes the analysis of perfect open circuits difficult, as suggested in the literature [12,13].

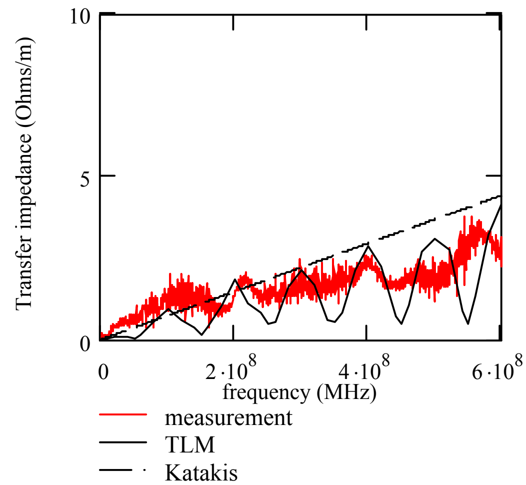

Figure 12. Comparison of the surface transfer impedance of a RG58 cable using theory, simulations and measurements.

DISCUSSION

The theory developed on this paper consists the basic block for analyzing many forms of communication cables [14, 15] by calculating the voltages and currents at any given point across their length. Although, in theory, the surface transfer impedance is just the ratio between the voltage induced into the inner conductor due to a current flowing to the shield over that current, implementation limitations have lead to different measurement methods [12,13] involving different termination conditions (matched, short, open etc). The resonances illustrated in the analysis of this paper are due to the long-line effect and mismatches and agree with other studies [11,16] involving analysis of multiconductor cables.

The problem that is addressed in this paper is that the theoretical model is not accurate enough to predict the surface transfer impedance after the first resonance (e.g. 50MHz in this paper). So, the TLM analysis of this paper can aid the prediction of the resonant performance and in developing a more comprehensive understanding of this problem. The most important limitation is that when considering electrically long lines, radiation effects may become significant [17] as the cable behaves like as an electrical dipole and lumped theory is not sufficient. However, in this paper, any electrically long line can be converted to an electrically short one by choosing a suitable node for TLM.

ACKNOWLEDGEMENTS

The authors of this paper would like to thank Alyse R Coates for providing the measurements for the RG58 cable using the Current Probe Method [12].

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