Generic prediction of crosstalk between shielded wires

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***Abstract*—A closed-form expression for near-end crosstalk between an unshielded and a double shielded wire is derived. Analysis in the frequency domain of such crosstalk expressions leads to generic crosstalk predictions. These predictions contain regions of different frequency dependencies separated by transition frequencies.**

Keywords—Near-end crosstalk; multi-conductor transmission line; shielding; low-frequency analysis

I. Introduction

Unintentional electromagnetic coupling between wires interconnecting electronic systems is usually called crosstalk. This phenomenon plays an important role in design rules for routing of cables in aircraft and automobiles. Crosstalk can be subdivided into inductive and capacitive coupling due to magnetic or electric fields respectively. Depending on the impedances of connected systems the magnitudes of inductive and capacitive coupling change. In [1] closed-form expressions were derived for crosstalk between single wires and wire pairs that explicitly reveal these dependencies. Of several solutions to minimise crosstalk, shielding is regarded an effective way to reduce capacitive crosstalk.

This paper presents a derivation of closed-form expressions for near-end crosstalk (NEXT) for doubly shielded wires, relating crosstalk levels to every model parameter. The presented methodology can be used to derive expressions for a variety of shielding configurations. Moreover, the expressions show interesting behaviour of the shields with respect to frequency. For double shielded wires five different frequency regions are observed. The presence of one shield gives rise to a region where crosstalk remains constant with frequency. The addition of a second shield could cause another such interval. Analysis of the transition frequencies and the crosstalk levels leads to generic predictions of crosstalk in the frequency domain. In a very practical way this can be advantageous for the design of cabling. For instance cable manufacturers can analyse effects of several model parameters on crosstalk levels. The generic prediction could be used to optimise shielding parameters with the aim to achieve desired crosstalk levels in certain frequency domains.

For the inclusion of shields in crosstalk models we extend the theory of Paul [2]. His book describes the derivation and solution of equations for multi-conductor transmission lines (MTLs). The results of Paul were used in [1] to derive closed-form expressions for near-end differential mode crosstalk between wire pairs close to a ground plane. An important aspect in solving the MTL equations is the determination of the per-unit-length (PUL) parameters for the cross section of shielded wires. Paul considers a method to include a single shield around a single wire into the PUL parameters and the MTL equations in several papers [3], [4]. In the present paper we extend this method for other shielding configurations. We derive a closed-form expression for common mode crosstalk in a configuration where two wires are situated parallel to an infinite, perfectly conducting ground plane. One of these wires carries a double shield, of which the electromagnetic interactions between in- and outside are modelled by two separate transfer impedances.

In section II the properties of the double shield are included into the PUL parameters for inductance, capacitance and resistance. For the case of a double shielded wire and an unshielded wire above a ground plane the PUL parameters are described by four-dimensional matrices. By substituting these matrices into the MTL equations closed-form expressions can be obtained for the voltages and crosstalk levels. To fully derive the effects of the shields we first solve the shield currents from the corresponding MTL equations. Next, we substitute these obtained expressions into the remaining two equations. This results into an augmented two-dimensional system for two wires above a ground plane. Like in [1], we can obtain analytical expressions for several shielding configurations by solving this augmented system by use of Taylor expansions in the frequency domain. This procedure leads to results corresponding with [3] when applied to the case where the double shield is replaced by a single shield.

The knowledge about effects of shielding is increased by analysing generic predictions of crosstalk in the frequency domain for several shielding configurations. In the final section of this paper we compare generic results for three different shielded situations versus one with two unshielded wires. It illustrates the practical use of analysing crosstalk levels with the introduced approach.

II. Transmission line characteristics

Consider a situation with two wires parallel to an infinite, perfectly conducting ground plane where one of the wires is surrounded by a double shield. The cross section is illustrated in Fig. 1 and the termination in Fig. 2. In the following we describe the modelling of this wiring configuration.

This research has partly been performed under contract with Fokker Elmo.
978-1-4799-6616-5/15/$31.00 ©2015 IEEE

A. *Per-Unit-Length (PUL) parameters*

The electromagnetic analysis of transmission lines requires the determination of per-unit-length parameters: capacitance C, inductance L, resistance R and conductance G. The latter is assumed to be zero throughout this paper; the medium in all our cable configurations is free space and therefore homogeneous and lossless. In setting up the MTL equations it is a necessity that the cross section of the line is uniform along the transmission line. As a consequence the PUL parameters should be equal along its length ℓ, since these contain all cross sectional information.

Modelling of shields in the PUL parameters is based upon papers by Paul [3], [4], in which PUL matrices were derived for a situation with one single shield. Here we extend these matrices to a double shielded configuration (see Fig. 1, which also gives numbering of conductors). Afterwards, in the next section we derive crosstalk expressions for this configuration.

1) *Inductance*

Fig. 2 shows that all conductors use the ground as return path for currents, thus self-inductances are all defined between wire (or shield) and ground. Proceeding on Paul's approach we take the inductance between the circuit containing wire 2 and that of the inner (or outer) shield equal to the self-inductance of that shield. Moreover we assume the mutual inductance between culprit transmission line and the inner (or outer) shield with ground return to be equal to the mutual inductance between the culprit and victim circuits. The mutual inductance between the two shields equals the self-inductance of the outer shield. Finally we assumed that the thickness of the shield is significantly smaller than its radius. All this results in:

$$\mathbf{L} = b\begin{bmatrix} \ln\left(\frac{2h}{r}\right) & \frac{\ln(1+x)}{2} & \frac{\ln(1+x)}{2} & \frac{\ln(1+x)}{2} \\ \frac{\ln(1+x)}{2} & \ln\left(\frac{2h}{r}\right) & \ln\left(\frac{2h}{r_{S,1}}\right) & \ln\left(\frac{2h}{r_{S,2}}\right) \\ \frac{\ln(1+x)}{2} & \ln\left(\frac{2h}{r_{S,1}}\right) & \ln\left(\frac{2h}{r_{S,1}}\right) & \ln\left(\frac{2h}{r_{S,2}}\right) \\ \frac{\ln(1+x)}{2} & \ln\left(\frac{2h}{r_{S,2}}\right) & \ln\left(\frac{2h}{r_{S,2}}\right) & \ln\left(\frac{2h}{r_{S,2}}\right) \end{bmatrix}. \quad (1)$$

Here $x = 4h^2/d^2$, $b = \mu_0/2\pi$ and μ_0 is the permeability of free space. The geometrical parameters are defined in Fig. 1.

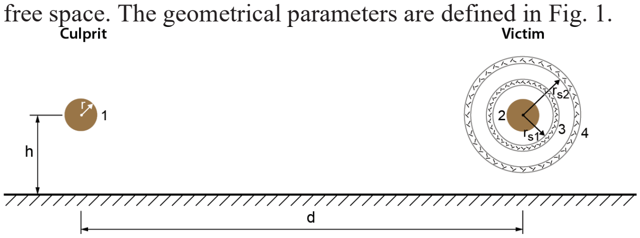

Fig. 1. Unshielded versus double shielded wire above an infinite, perfectly conducting ground plane. Both wires have equal radius *r* and height *h*.

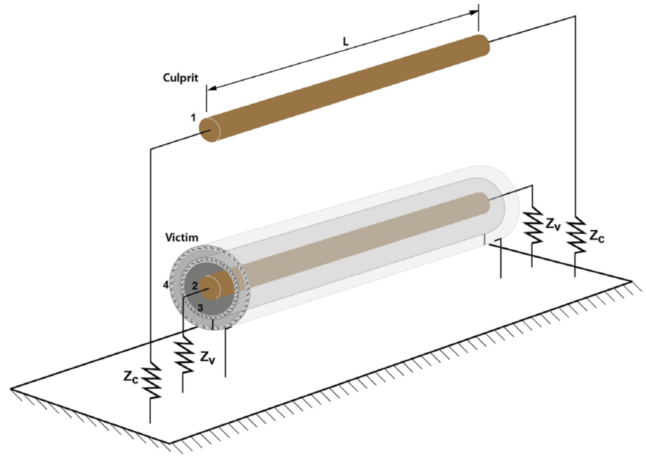

Fig. 2. Finishing of the configuration with one unshielded wire and one double shielded wire.

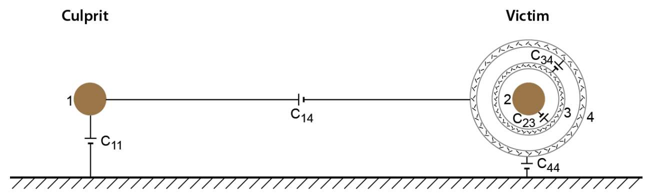

Fig. 3. Capacitances for the configuration of one unshielded wire and one double shielded wire.

2) *Capacitance*

Both shields are directly connected to the ground. Therefore the capacitances between the culprit wire and the inner conductors (referenced by number 2 and 3 in Fig. 3) are neglected. In Fig. 3 all nonzero capacitances of our model are given. All of them represent an entry in the capacitance matrix. The diagonal entries also contain the sum of the other entries in that matrix row. This gives:

$$\mathbf{C} = \begin{bmatrix} (c_{11}+c_{14}) & 0 & 0 & -c_{14} \\ 0 & c_{23} & -c_{23} & 0 \\ 0 & -c_{23} & (c_{23}+c_{34}) & -c_{34} \\ -c_{14} & 0 & -c_{34} & (c_{44}+c_{14}+c_{34}) \end{bmatrix}. \quad (2)$$

Here it is assumed that the well-known relation between capacitance and inductance for homogeneous media holds for the culprit wire versus outer shield, resulting in:

$$\begin{bmatrix} c_{11}+c_{14} & -c_{14} \\ -c_{14} & c_{44}+c_{14} \end{bmatrix} = \mu_0\varepsilon_0 \begin{bmatrix} l_{11} & l_{14} \\ l_{14} & l_{44} \end{bmatrix}^{-1}.$$

Here ε_0 equals permittivity of free space. For the capacitances between the victim wire and inner shield respectively inner and outer shield we have:

$$c_{23} = \frac{2\pi\varepsilon_0}{\ln\left(r_{S,1}/r\right)}, \qquad c_{34} = \frac{2\pi\varepsilon_0}{\ln\left(r_{S,2}/r_{S,1}\right)}.$$

3) Resistance

Since shields reduce the coupling between wires, but will not eliminate it completely, they cannot be modelled as perfect conductors. Therefore we have positive resistivity matrix entries at the diagonal positions that represent the shields, giving $\mathbf{R} = diag\begin{pmatrix}0 & 0 & Z_{T,1} & Z_{T,2}\end{pmatrix}$. It contains the transfer impedance of the shields, given by $Z_{T,k}$ where *k* is the number of the shield. In literature accurate models for this quantity can be found [5], though here for illustration purposes we use simplified models:

$$Z_{T,k} = R_{T,k} + j\omega l_t, \quad \text{with} \quad l_t = 2/\pi \ nF/m. \tag{3}$$

Here the transfer impedance of shield *k* equals a combination of the constant DC resistance part of the transfer impedance $R_{T,k}$ and its inductive part l_t.

B. Impedance matrices

Besides the PUL parameters also termination networks determine the behaviour of transmission lines. In Fig. 2 it is seen that for this wiring situation the terminations of the transmission lines are straightforward. Moreover it is equal at both sides. This results in a four by four impedance matrix $\mathbf{Z} = diag\begin{pmatrix}Z_c & Z_v & 0 & 0\end{pmatrix}$ in which the only nonzero entries are given by Z_c and Z_v, respectively the culprit and victim termination impedance.

III. Near-End Crosstalk

Like in [1] we define $\mathbf{V}_0$ as the vector of voltage differences between each conductor and the reference plane. The vector $\mathbf{I}_0$ contains the corresponding currents flowing in each conductor. Each pair of conductors forms a transmission line and between each pair of transmission lines electromagnetic coupling might occur which causes crosstalk. We model the common mode near-end crosstalk between two transmission lines as follows:

$$\gamma_{NE} = \frac{\mathbf{U}_2^T \mathbf{V}_0}{\mathbf{U}_1^T \mathbf{V}_0}. \tag{4}$$

Here $\mathbf{U}_1 = \begin{pmatrix}1 & 0 & 0 & 0\end{pmatrix}$ and $\mathbf{U}_2 = \begin{pmatrix}0 & 1 & 0 & 0\end{pmatrix}$ serve to respectively select the culprit and victim wires from $\mathbf{V_0}$.

For deriving explicit, analytic crosstalk expressions we use the matrix formulation derived from the MTL equations by Paul in its book on multiconductor transmission lines [2]. Equation (4) requires the solution of voltages which can be obtained from equation (4.90) of [2]:

$$\begin{aligned}&\left[\mathbf{Z}\phi_{22}(\ell) - \phi_{12}(\ell) + \phi_{11}(\ell)\mathbf{Z} - \mathbf{Z}\phi_{21}(\ell)\mathbf{Z}\right]\mathbf{I}_0 \\ &= \left[\phi_{11}(\ell) - \mathbf{Z}\phi_{21}(\ell)\right]\mathbf{V_S},\end{aligned} \tag{5}$$

in combination with:

$$\mathbf{V}_0 = \mathbf{V_S} - \mathbf{Z}\mathbf{I}_0. \tag{6}$$

Here $\mathbf{V}_S$ is the vector containing all voltage sources. We assume that a voltage source (see Fig. 2) is present only at one side of the culprit transmission line (thus $\mathbf{V}_L = 0$). Then:

$$\mathbf{V}_S = V_S \mathbf{U}_1.$$

Expressions for the chain matrices ϕ_{ij} are given in [2].

A. One double shielded versus one unshielded wire

For the given wiring configuration we use the following first order approximation for the chain matrices [3]:

$$\begin{aligned}\phi_{11} &= \mathbf{I}_4 \\ \phi_{12} &= -j\omega\ell\mathbf{L} - \ell\mathbf{R} \\ \phi_{21} &= -j\omega\ell\mathbf{C} \\ \phi_{22} &= \mathbf{I}_4.\end{aligned} \tag{7}$$

Here $\mathbf{I}_4$ is the 4 by 4 identity matrix and ω is the angular frequency. Via these chain matrices the transfer impedance appears into the MTL equations thereby including the behaviour of the shield.

To solve the matrix equation (5) we first solve the two equations for the shield currents. If subsequently these currents are substituted along with the chain matrices into the other two equations of (5), we find a two-dimensional matrix equation given by:

$$\left[2\mathbf{Z} + j\omega\ell\left[\mathbf{L} + \mathbf{ZCZ} - \mathbf{S}\right]\right]\mathbf{I}_0 = \left[\mathbf{I}_2 + j\omega\ell\mathbf{ZC}\right]\mathbf{V_S}. \tag{8}$$

In order to drop two dimensions in the already defined PUL and impedance matrices, all row or column entries concerning the shields are deleted. The matrix $\mathbf{S}$ is given by:

$$\mathbf{S} = j\omega\ell \cdot g \cdot \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22}\end{bmatrix}, \tag{9}$$

in which:

$$\begin{aligned}S_{11} &= l_{12}^2 Z_{T,2}\ell + l_{12}^2\left(Z_{T,1}\ell + j\omega\ell \cdot l_{33}\right) - l_{12}^2 l_{44} j\omega\ell \\ S_{12} &= l_{12} l_{33}\left(Z_{T,2}\ell + j\omega\ell \cdot l_{44}\right) + l_{12} l_{44} Z_{T,1}\ell - l_{12} l_{44}^2 j\omega\ell \\ S_{21} &= S_{12} \\ S_{22} &= l_{33}^2\left(Z_{T,2}\ell + j\omega\ell \cdot l_{44}\right) + l_{44}^2 Z_{T,1}\ell - l_{33} l_{44}^2 j\omega\ell,\end{aligned}$$

and:

$$g = \left[\left(Z_{T,1}\ell + j\omega\ell l_{33}\right)\left(Z_{T,2}\ell + j\omega\ell l_{44}\right) - \left(j\omega\ell l_{44}\right)^2\right]^{-1}.$$

Here we write l_{ij} for element (i,j) of the inductance matrix.

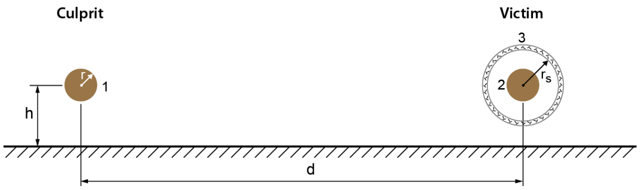

Fig. 4. One unshielded wire and one wire carrying a single shield, both parallel above an infinite, perfectly conducting ground plane.

By using a low-frequency approximation to (8) together with equation (6) we obtain the voltage vector:

$$\mathbf{V}_0 \approx \frac{V_S}{2}\left[\mathbf{I}_2 + \frac{1}{2} j\omega\ell\left(\mathbf{L}\mathbf{Z}^{-1} - \mathbf{Z}\mathbf{C} - \mathbf{S}\mathbf{Z}^{-1}\right)\right]\mathbf{U}_1. \quad (10)$$

Finally we compute the numerator and denominator of (4):

$$\begin{aligned} \mathbf{U}_1^T \mathbf{V}_0 &= \frac{V_S}{2} + O(\omega\ell) \\ \mathbf{U}_2^T \mathbf{V}_0 &= \frac{V_S}{4} Z_c^{-1} j\omega\ell\left(l_{21} - S_{21}\right) + O\left(\omega^2\ell^2\right), \end{aligned} \quad (11)$$

which leads to the following closed-form expression for the near-end crosstalk between two wires, of which one carries a double shield:

$$\gamma_{NE} \approx \frac{j\omega\ell \cdot b\ln\left(1+4h^2/d^2\right) Z_{T,1} Z_{T,2}}{4Z_c\left[Z_{T,1}^* Z_{T,2}^* - \left(j\omega b\ln\left(2h/r_{s,2}\right)\right)\right]^2}, \quad (12)$$

in which the extended impedance $Z_{T,k}^*$ is given by $Z_{T,k}^* = Z_{T,k} + j\omega b\ln\left(2h/r_{s,k}\right)$ for $k \in \{1,2\}$. Inspection of (11) and (12) reveals that the near-end crosstalk is governed by inductive coupling only. Capacitive crosstalk vanishes because our model assumes that the shields are grounded.

Fig. 5 shows that there is a good comparison between MTL simulations and the expression given by (12). For simulations we used the following values for the model parameters: h=1.67mm, r=0.16mm, d=10mm, ℓ=2m, $r_{s,1}$=0.3mm, $r_{s,2}$=0.52mm and Z_v=Z_c=100 Ω. Finally the transfer impedances are equal to $Z_{T,1}$=0.005+$j\omega l_t\ \Omega/m$ and $Z_{T,2}$=0.1+$j\omega l_t\ \Omega/m$.

B. *Comparison to literature: one single shielded wire*

The configuration with two wires in which one of the wires carries a single shield has been covered in literature before (see for instance [3]). An illustration of this situation is given in Fig. 4. If we follow the procedures explained above, we find that the near-end crosstalk in this situation is equal to:

$$\gamma_{NE} \approx \frac{j\omega\ell b Z_t \ln\left(1+4h^2/d^2\right)}{4Z_c\left[Z_t + j\omega b\ln\left(2h/r_s\right)\right]}. \quad (13)$$

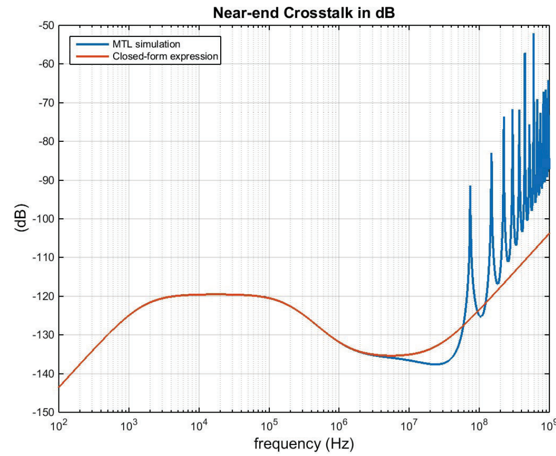

Fig. 5. MTL simulations (blue) and the closed-form expression in (12) (red) of near-end crosstalk between an unshielded and a double shielded wire.

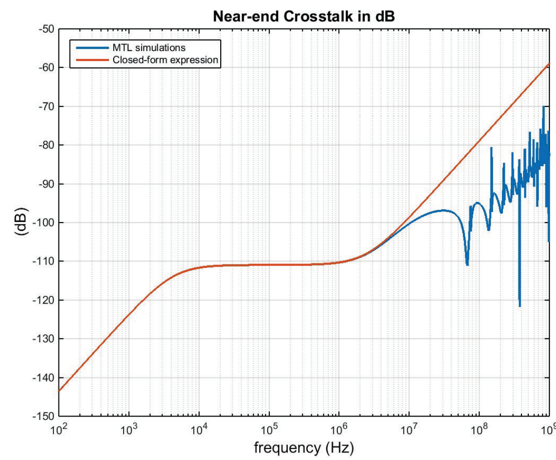

Fig. 6. MTL simulations (blue) and the closed-form expression in (13) (red) of near-end crosstalk between an unshielded and a single shielded wire.

This expression coincides with results found in literature. Fig. 6 again shows a good comparison between (13) and MTL simulations, in which we used the same parameter values as for Fig. 5 (the single shield has radius equal to the outer shield in the previous situation).

IV. Generic Crosstalk Predictions

By analysing results of the previous section several regions with different kinds of crosstalk behaviour can be observed. This behaviour can be retraced and explained by further analysis of the closed-form expressions of near-end crosstalk. Apparently several transition frequencies occur where the dependency on frequency changes. Evidently this is where certain parts of shield inductances and transfer impedances start to contribute. To find these dependencies and transitions we use (12) and (3). This analysis leads to Fig. 7 that illustrates the generic prediction of crosstalk with respect to frequency, as well as shielding parameters.

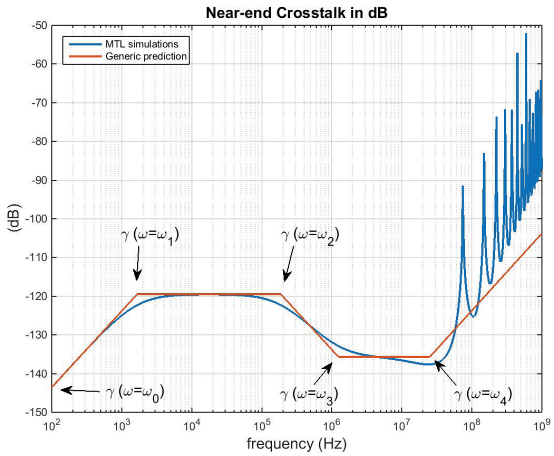

Fig. 7. Generic prediction of crosstalk in the frequency domain (red) along with MTL simulations (blue) for the configuration of one unshielded versus one double shielded wire.

For all expressions derived in the previous section it holds that the leading order term in our low-frequency approximation equals the inductive coupling between two single wires above a ground plane (see [1]). This holds for frequencies where both transfer impedance and shield inductance are negligible. In this first region of relatively low frequencies the behaviour is linear with respect to ω and the shields have no influence on inductive crosstalk levels. The shields only cause the capacitive crosstalk to vanish. In the following we will derive the other regions of different frequency dependence. Therefore we assume the values used in simulations before, $5\cdot10^{-3}\Omega = R_{T,1} < R_{T,2} = 10^{-1}\Omega$. Moreover $l_t = 2/\pi\, nF/m$ by which the self-inductance of the shield will play a role for lower frequencies than the inductance part of the transfer impedance. If values are different and the order of these inequalities changes the figures and transition frequencies below will also change, though the analysis procedure remains the same.

A. One double shielded versus one unshielded wire

Analyses of (12) and Fig. 5 reveals five different regions of crosstalk behaviour, which are separated by the following four transition frequencies:

$$\omega_k = \begin{cases} \dfrac{2\pi R_{T,1}}{\mu_0 \ln\left(2h/r_{s,1}\right)} & \text{for } k=1 \\ \dfrac{2\pi R_{T,2}\ln\left(2h/r_{s,1}\right)}{\mu_0 \ln\left(r_{s,2}/r_{s,1}\right)\ln\left(2h/r_{s,2}\right)} & \text{for } k=2 \\ \dfrac{R_{T,k-2}}{l_t} & \text{for } k\in\{3,4\}. \end{cases} \quad (14)$$

These transition frequencies are also shown in Fig. 7. The values of ω_1 to ω_4 separate the following crosstalk regions:

1. $\boxed{\omega < \omega_1}$

For all shielding configurations this first region obtains linear behaviour with respect to frequency and crosstalk levels are equal to inductive coupling when no shield is present. Any point on this line can be calculated by:

$$\gamma_{NE} = \frac{j\omega\ell}{4Z_c}\cdot b\ln\left(1+4h^2/d^2\right). \quad (15)$$

2. $\boxed{\omega_1 < \omega < \omega_2}$

In this region the self-inductance of the shield with the lowest transfer impedance starts to contribute. This frequency dependence cancels against the linear behaviour of the first region and therefore gives rise to a constant region in the crosstalk graph. Its value is equal to:

$$\gamma_{NE} = \frac{\ell}{4}\cdot\frac{R_{T,1}}{Z_c}\cdot\frac{\ln\left(1+4h^2/d^2\right)}{\ln\left(2h/r_{s,1}\right)}. \quad (16)$$

3. $\boxed{\omega_2 < \omega < \omega_3}$

For these frequencies also the inductance of the second shield interacts. Then the crosstalk becomes inversely proportional to ω. The values of this region can be calculated via:

$$\gamma_{NE} = \frac{\ell R_{T,1}R_{T,2}}{4bZ_c}\cdot\frac{\ln\left(1+4h^2/d^2\right)}{j\omega\cdot\ln\left(r_{s,2}/r_{s,1}\right)\ln\left(2h/r_{s,2}\right)}. \quad (17)$$

4. $\boxed{\omega_3 < \omega < \omega_4}$

Here the contribution of inductance of the transfer impedance of one of the shields becomes large in comparison to its resistance part. Then the frequency dependence in the numerator is cancelled and again a constant value appears, which is equal to:

$$\gamma_{NE} = \frac{\ell l_t R_{T,2}}{4bZ_c}\cdot\frac{\ln\left(1+4h^2/d^2\right)}{\ln\left(r_{s,2}/r_{s,1}\right)\ln\left(2h/r_{s,2}\right)}. \quad (18)$$

5. $\boxed{\omega > \omega_4}$

In this last region also the inductance of the transfer impedance of the second shield becomes larger than its resistance. This ensures a linear increase of crosstalk with respect to frequency in the fifth region. The crosstalk formula now becomes:

$$\gamma_{NE} = \frac{j\omega\ell}{4b}\cdot\frac{l_t^2}{Z_c}\cdot\frac{\ln\left(1+4h^2/d^2\right)}{\ln\left(r_{s,2}/r_{s,1}\right)\ln\left(2h/r_{s,2}\right)}. \quad (19)$$

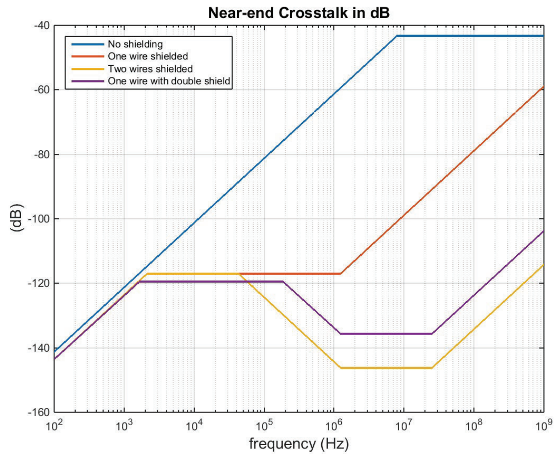

Fig. 8. Comparison of the generic prediction of crosstalk in four different wiring configurations.

In Fig. 7 results of MTL simulations are shown together with the generic prediction given above. The results correspond very well, except for the fifth frequency dependence. This is a limitation to our model, since the chain matrices introduced in (7) are invalid for such high frequencies. This causes the fact that (8) already fails to hold.

B. Comparison of shielding configurations

With the analysis shown above we have derived closed-form expressions and corresponding generic predictions of four different shielding configurations:

1. Two unshielded wires above a ground plane
2. One unshielded wire and one single shielded wire above a ground plane
3. Two single shielded wires above a ground plane
4. One unshielded wire and one double shielded wire above a ground plane.

Fig. 8 compares the generic crosstalk prediction of these four wiring configurations.

As stated in the previous section, the low-frequency behaviour of all situations is equal, except for a possibly significant part of capacitive crosstalk in the situation without shielding. The chosen loads result in mostly inductive crosstalk, though larger values for Z_v and Z_c could have made a bigger difference.

For all shielded situations even the first constant regions are quite similar. The only difference in levels is caused by a smaller shield radius of the inner shield in the double shielded situation compared to the single shields in the other two configurations. This gives a slightly lower constant value.

The introduction of a second shield even causes a temporary linear decrease and a second flat region before eventually crosstalk levels increase just like the configuration with one shield. The differences in transition frequency and constant value of situations 3 and 4 are caused by the difference in inductive coupling between the two shields. In the third case this is negligible by the fact that the separation distance of the wires is significantly larger than the radius of the shield. With the double shield this coupling is not significantly smaller than the self-inductances of the shields, which causes the differences observed in Fig. 8. Evidently, the third and fourth transition frequencies are equal for these two situations, since the transfer impedances of the shields are chosen equal.

V. Conclusions

A method to include transfer impedance parameters of multiple shields in crosstalk models has been described. The low-frequency solution of the MTL equations leads to closed-form expressions for near-end crosstalk. These expressions explicitly contain geometrical parameters, as well as termination impedances and transfer impedance parameters. For the situation of one unshielded versus one double shielded wire the expression is given by (12). The analysis of this expression reveals regions of different frequency dependencies. For these regions generic crosstalk predictions have been presented, which describe in a practical way dependencies of the different regions and their crosstalk levels on shield resistance, shield inductance, termination impedance and geometrical parameters.

Finally we compared generic crosstalk predictions of four different wiring configurations. This results in knowledge about the effects of shielding wires. The inclusion of a second shield causes crosstalk to decrease with frequency on a certain interval, which is followed by a second constant crosstalk level. Generic crosstalk prediction of different wiring configurations can be used to optimise shielding parameters to obtain certain crosstalk levels on specified frequency intervals.

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