Equivalent cable harness method generalized for predicting the electromagnetic emission of twisted-wire pairs

**Introduction.** In this paper, the equivalent cable harness method is generalized for predicting the electromagnetic emissions problems of twisted-wire pairs. The novelty of the proposed work consists in modeling of a multiconductor cable, in a simplified cable harness composed of a reduced number of equivalent conductors, each one is representing the behavior of one group of conductors of the initial cable. **Purpose.** This work is focused on the development and implementation of simplified simulations to study electromagnetic couplings on multiconductor cable. **Methods.** This method requires a four step procedure which is summarized as follows. Two different cases, of one end grounded and two ends grounded configurations can be analyzed. **Results.** The results had shown that the model complexity and computation time are significantly reduced, without, however, reducing the accuracy of the calculations. References 20, tables 1, figures 8.

**Key words:** electromagnetic emission, asymmetric digital subscriber line, crosstalk, equivalent cable harness method, multiconductor transmission line network, power loss, twisted-wire pairs.

**Вступ.** У цій статті метод еквівалентного кабельного джгути використовується для прогнозування значень електромагнітного випромінювання кручених пар дротів. Новизна запропонованої роботи полягає в моделюванні багатожильного кабелю в спроценому джгути проводі, що складається зі зміненої кількості еквівалентних провідників, кожен з яких репрезентує певну однієї групи провідників вихідного кабелю. **Мета.** Розробка зосереджена на розробці і реалізації спроценого моделювання для дослідження електромагнітних зв'язків у багатожильних кабелях. **Методи.** Цей метод вимагає чотирьохступінчастої процедури, яка коротко описана у статті. Можна проаналізувати два різних випадки: конфігурації із заземленням одного кінця та заземлення двох кінців. **Результати.** Результати показали, що складність моделі та тривалість обчислень значно знизуються, проте без зниження точності обчислень. **Ключові слова:** електромагнітне випромінювання, асиметрична цифрова абонентська лінія, перехресні перешкоди, метод еквівалентного кабельного джгути, мережа багатопровідних ліній передачі, втрати потужності, крученні пар дротів.

**Introduction.** For transmission signal in complex telecommunication network one of principle resources of noise that affect the signal quality is due to the crosstalk coupling. The crosstalk among twisted-wire pairs (TWPs) is commonly classified into near end crosstalk (NEXT) and far end crosstalk (FEXT) [1–4]. Furthermore, we distinguished two types of coupling crosstalk, the inductive coupling and the capacitive coupling, the dominant coupling at an arbitrary configuration is due to the termination impedance effects [5, 6].

The survey of literature shows that reducing the number of wires is not a new task. An efficient simplification technique called the equivalent cable harness method (ECBM) has been proposed for modelling electromagnetic (EM) common-mode currents on complex cable bundles for telecommunications networks applications with arbitrary loads [7]. This method, allows, using the theory of Multiconductor Transmission Line Network (MTLN), to take into account the phenomena of propagation and couplings on and between all the wires of the network.

In recent years, the increase in the frequency of disturbances electromagnetic that can potentially attack the wiring network encourages the research to extend digital immunity to high frequencies. This will unfortunately come up against the limitation in MTLN frequency.

In high frequencies, the use of three-dimensional methods, solving the Maxwell equations in space then proves to be obligatory. However, they require a fine discretization of each conductor of the beam in segments whose length is usually less than one tenth of the wavelength. The computation times then become prohibitive as soon as the number of conductors of the beam is important.

It must therefore find simplifying hypotheses allowing the complexity to be reduced wiring harness without, however, reducing the accuracy of the calculations [8, 9]. The goal final is to generalize the feasibility of such an approach for twisted-wire pairs.

**Purpose.** The work is focused on the development and implementation of simplified simulations to study electromagnetic couplings on multiconductor cable.

**Presentation of the equivalent cable harness method proposed of complex twisted-wire pairs.** Description of the main assumption adopted in this modified procedure for a coupling problem in twisted-wire pairs cable will be detailed in this section.

The proposed geometry consists of a cable composed of twisted-wire pairs, both ends or only one end of the each pair circuit can be grounded.

Note that the twisted-wire pairs used here are connected to the ground plane. We would expect common mode current to be dominant. Therefore, the main assumption of the original ECBM is unchanged [10], we make approach that the impedance loads in such case of twisted-wire pairs can be considered such as a common mode loads and the differential loads in one end of the pairs can be neglected.

The determination of geometrical characteristics of the reduced cable is omitted here because the aim of the method is to predict crosstalk in the pairs that we are interested in his currents and voltages [11, 12].

The modified equivalent cable harness method for modeling crosstalk in frequency domain among twisted-
wire pair requires a four-step procedure [9], which is summarized as follows.

**Step I.** The goal of this step is to classify the pairs of the initial cable in different groups according to their termination loads at both ends of each pair. The culprit and victim pairs are classified into two groups separately and they hold its initial characteristics (including its positions, radius, and medium). The common-mode characteristic impedance $Z_{mc}$ is determined by modal transformation in the MTLN formalism in order to obtain the characteristics of all the modes propagating along the cable. Furthermore, the very important condition of the eigenvector associated with the common mode in the matrix $[T_k]$ is verified, this part is more detailed in the original ECBM [8, 10].

Then all the remaining pairs in the complete cable bundle are sorted into groups by comparing the value of the termination loads (near $Z_0$ and far $Z_L$) to $Z_{mc}$. The conductors are installed as pairs; each two conductors for one pair are in the same group because the modal analysis is made of twisted-wire pairs configuration. We define the four groups (may be less than four) which can be determined in each configuration in Table 1.

**Step II.** The determination of the per unit length (p.u.l) parameter matrices of an equivalent cable is based on the determination of p.u. parameter matrices of the pairs from which it is consisted.

We consider a short circuit between conductors of one group of «k» conductors (Fig. 1). This assumption allows: firstly to define the group current $I_{ec}$ for the equivalent cable and secondly the group voltage $V_{ec}$ for the equivalent cable.

![Fig. 1. Mode common current and voltage of k conductors in the same group](image)

To determine the inductance reduced matrix, we need two additional assumptions detailed in [9]:
1) the currents flowing along all the «k» conductors of a cable bundle are decomposed in common mode currents and a differential mode currents. The differential mode current can be neglected [5, 13, 14].

The current and the voltage of «k» conductors in the same group can be written by:
   \[ I_{ec} = I_1 + I_2 + \ldots + I_k \]  \hspace{1cm} (1)
   \[ V_{ec} = V_1 + V_2 + \ldots + V_k \]  \hspace{1cm} (2)

Thus, from the telegrapher's equation on the MTLN formalism for lossless wire can we obtain inductance matrix links the currents and the voltages on each conductor on an infinitesimal segment of length:

\[
\frac{\partial}{\partial z} \begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_k \end{bmatrix} = j\omega \begin{bmatrix} L_{11} & L_{12} & \cdots & L_{1k} \\ L_{21} & L_{22} & \cdots & L_{2k} \\ \vdots & \vdots & \ddots & \vdots \\ L_{k1} & L_{k2} & \cdots & L_{kk} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_k \end{bmatrix} . \]  \hspace{1cm} (3)

2) The common mode currents along all the conductors of a same group are identical:
   \[ I_{ec} = I_1 = I_2 = \ldots = I_k , \]  \hspace{1cm} (4)

where $I_k$ is the mode common current in the $k$ conductor.

The common-mode characteristic impedance of the harness can be obtained:

\[ V_{ck} = \frac{V_{ec}}{kI_{ec}} = \frac{Z_i}{k} , \]  \hspace{1cm} (5)

where $Z_i$ is the common-mode characteristic impedance of each conductor in the group (Fig. 2).

![Fig. 2. Cross section view of complete and reduced cable](image)

These assumptions may allow finding the final system matrix; we index the conductors of each group as follows:
- $N_1$ conductors of the first group indexed 1 to $x$;
- $N_2$ conductors of the second group indexed $x+1$ to $y$;
- $N_3$ conductors of the third group indexed $y+1$ to $z$;
- $N_4$ conductors of the fourth group indexed $z+1$ to $N$.

The final system that allows finding $L_{red}$ can be obtained:

\[
\frac{\partial V_1}{\partial x} = -j\omega \begin{bmatrix} \sum_{i=1}^{x} I_{1i} I_{ec1} + \sum_{i=x+1}^{y} I_{1i} I_{ec2} \\ \sum_{i=x+1}^{y} I_{2i} I_{ec3} + \sum_{i=z+1}^{N} I_{2i} I_{ec4} \\ \sum_{i=y+1}^{z} I_{3i} I_{ec1} + \sum_{i=x+1}^{N} I_{3i} I_{ec2} \\ \sum_{i=z+1}^{N} I_{4i} I_{ec3} + \sum_{i=1}^{N-1} I_{4i} I_{ec4} \end{bmatrix} ; \]  \hspace{1cm} (6)

\[
\frac{\partial V_N}{\partial x} = -j\omega \begin{bmatrix} \sum_{i=1}^{x} I_{N1} I_{ec1} + \sum_{i=x+1}^{y} I_{N1} I_{ec2} \\ \sum_{i=x+1}^{y} I_{N2} I_{ec3} + \sum_{i=z+1}^{N} I_{N2} I_{ec4} \\ \sum_{i=y+1}^{z} I_{N3} I_{ec1} + \sum_{i=x+1}^{N} I_{N3} I_{ec2} \\ \sum_{i=z+1}^{N} I_{N4} I_{ec3} + \sum_{i=1}^{N-1} I_{N4} I_{ec4} \end{bmatrix} . \]  \hspace{1cm} (7)

The voltages of each conductor belonging to the same group being equal (2), the insertion of this property in previous equations leads to the following system connecting the voltages and currents of the four groups of conductors [7]:

---

30  Electrical Engineering & Electromechanics, 2022, no. 2
\[ \frac{\partial V_{ec1}}{\partial x} = -j\omega \left[ \sum_{j=1}^{N_1} \frac{L_{jj}}{N_1^2} I_{ec1} + \sum_{j=1}^{N} \frac{L_{j+N}}{N_1 N_2} I_{ec2} \right] \]

\[ + \sum_{j=1}^{N} \frac{L_{N+y}}{N_1 N_3} I_{ec3} + \sum_{j=1}^{N} \frac{L_{N+y}}{N_1 N_4} I_{ec4} \]

\[ \frac{\partial V_{ec2}}{\partial x} = -j\omega \left[ \sum_{j=1}^{N_1} \frac{L_{jj}}{N_1^2} I_{ec1} + \sum_{j=1}^{N} \frac{L_{j+N}}{N_1 N_2} I_{ec2} \right] \]

\[ + \sum_{j=1}^{N} \frac{L_{N+y}}{N_1 N_3} I_{ec3} + \sum_{j=1}^{N} \frac{L_{N+y}}{N_1 N_4} I_{ec4} \]

\[ \frac{\partial V_{ec3}}{\partial x} = -j\omega \left[ \sum_{j=1}^{N_1} \frac{L_{jj}}{N_1^2} I_{ec1} + \sum_{j=1}^{N} \frac{L_{j+N}}{N_1 N_2} I_{ec2} \right] \]

\[ + \sum_{j=1}^{N} \frac{L_{N+y}}{N_1 N_3} I_{ec3} + \sum_{j=1}^{N} \frac{L_{N+y}}{N_1 N_4} I_{ec4} \]

\[ \frac{\partial V_{ec4}}{\partial x} = -j\omega \left[ \sum_{j=1}^{N_1} \frac{L_{jj}}{N_1^2} I_{ec1} + \sum_{j=1}^{N} \frac{L_{j+N}}{N_1 N_2} I_{ec2} \right] \]

\[ + \sum_{j=1}^{N} \frac{L_{N+y}}{N_1 N_3} I_{ec3} + \sum_{j=1}^{N} \frac{L_{N+y}}{N_1 N_4} I_{ec4} \]

\[ [L_{red}] = \begin{bmatrix}
  L_{11r} & L_{12r} & L_{13r} & L_{14r} \\
  L_{21r} & L_{22r} & L_{23r} & L_{24r} \\
  L_{31r} & L_{32r} & L_{33r} & L_{34r} \\
  L_{41r} & L_{42r} & L_{43r} & L_{44r}
\end{bmatrix} \]

\[ L_{11r} = \sum_{j=1}^{N_1} \sum_{y=1}^{N} L_{j+y} \]

The expression of the capacitance matrix for the reduced model \([C_{red}]\) can be obtained with the same reasoning in non-homogenous and polar medium.

For a weakly polar medium, one can ignore the dispersion of the dielectric constant and, accordingly, the phase velocity [11]:

\[ [C_{red}] = \frac{1}{\sqrt{\varepsilon_v}} [L_{red}]^{-1}, \]

where \( \varepsilon_v = \varepsilon / \sqrt{\varepsilon_r} \).

So, using the assumptions and approximations set out above, the matrix inductance and capacitance with \((N \times N)\) dimension can be reduced into matrix \((4 \times 4)\) which coefficients exerted on and between the conductors.

**Step III.** The aim of this step is to determine the termination loads to be connected at both ends of the equivalent conductor. Here, all load impedances are considered as pure resistance and are not frequency dependent. There are two kinds of the termination loads, differential-mode loads and common-mode loads:

- **Common-mode loads.** Each load connected to the ground plane at his terminal end (Fig. 3).

  ![Fig. 3. Equivalent termination common-mode loads](image)

  The impedance equivalent \(Z_{ec}\) for \(k\) conductors in the same group \((Z_1, Z_2, \ldots, Z_k)\), is:

  \[ \frac{1}{Z_{ec}} = \frac{1}{Z_1} + \frac{1}{Z_2} + \ldots + \frac{1}{Z_k}. \]

- **Differential-mode loads.** We consider the type differential loads connected between conductors from different groups (Fig. 4). The impedance equivalent \(Z_{dis}\) for two conductors \((1 \text{ and } 2)\) in group 1 and two conductors \((N1+1 \text{ and } N1+2)\) in group 2, is:

  \[ \frac{1}{Z_{dis}} = \frac{1}{Z_{1-2}} + \frac{1}{Z_{2-N1+2}}. \]

![Fig. 4. Equivalent termination differential-mode loads](image)

The type of differential loads connected between conductors in the same group is neglected.

**Step IV.** The aim of this step is to determine the section geometry (radius, coating radius, permittivity relative of coating, height, distance between harnesses), of each reduced harness using the linear parameter matrices determined at the Step II. This part is more detailed in the original ECBM [8, 10].

For modeling the twist of pairs on complete model the matrix \(P\) [1] was used, in the reduced model this matrix is evaluated to take into account only the twisted wires of the pairs which we not apply the method [1].

For the equivalent cable for each group the current and voltage are multiplied by 1 in the matrix \(P\).

In order to determine currents and voltages in both ends of each pair, the excited pair is assumed to be a disturber and the crosstalk in an arbitrary victim pair in term of power loss (PL) is shown.

Note that the values of the power loss crosstalk \((PL_{NEXT}, PL_{FEXT})\) in decibels can be calculated as follows [15–18]:
The mathematical model established will be used for obtaining the power loss near and far (PL$_{NEXT}$ and PL$_{FEXT}$) in a bundle of twisted wire pairs. Twisted wire-pair is connected by two ends of each pair to the ground plane. Figure 5 illustrates the geometry used; it shows the first case of initial model of twenty-eight pairs denoted from 1 to 56, and the reduced cable of two pairs and two equivalent harness denoted ec1, ec2.

All wires are further assumed to have the same radius \( a = 0.523 \) mm and polyvinyl chloride coating radius of \( b = 0.549 \) mm. The length of the wire is \( L = 1 \) km. The twisted pair consists of \( N = 10000 \) loops. The height of the first wire (numeral 8 and 10) above ground conductor is \( h = 20 \) mm, the height is very close to the reference plane, the intention in doing so is to avoid the noise produced by the loop between the receptor circuit and the reference plane.

The other conductors are located just above and next to the first wire with vertical distance of 0.127 mm for the same pair conductors and 3.17 mm for other conductors, the horizontal distance between the wires is \( d = 3.17 \) mm.

In Fig. 6 the pair one is given as an example, which is connected to the voltage source \( V_0 \) at its near end through a load \( Z_{01} \) and adapted at its far end through a load \( Z_{D1} \).

The crosstalk will be studied in the frequency range from 1 kHz to 30 MHz with reference to Fig. 6 and in order to make same loads such as in XDSL systems, the loads in the pairs which we are interested are set to \( Z_{01} = Z_{D1} = Z_{02} = Z_{D2} = 120 \) \( \Omega \) under the condition, that for a twisted pair the active resistance \( R \) of the conductors is much less than the inductive resistance \( \omega L \), and the active conductivity \( G \) of the insulation is much less than the capacitive resistance \( \omega C \).

### Application of the equivalent cable harness method

The classification of groups is made according to the comparison with the common-mode loads determined by the modal analysis.

The pairs of the complete cable bundle sorted into four groups as follows (Fig. 5):

- group 1: pairs 1 (conductors 1-2);
- group 2: pairs 2 (conductors 3-4);
- group 3: harness 1 (conductors 5-30);
- group 4: harness 2 (conductors 31-56).

The p.u. parameter matrices \( [L_{red}] \) and \( [C_{red}] \) of the reduced harnesses cables are calculated in nH/m and pF/m for a 1 km long line respectively:

\[
[L_{red}] = \begin{bmatrix}
1.02 & 0.88 & 0.25 & 0.25 & 0.25 & 0.19 \\
1.02 & 0.25 & 0.24 & 0.24 & 0.19 \\
1.02 & 0.88 & 0.13 & 0.17 & 0.13 & 0.17 \\
0.36 & 0.16 & 0.34 & \\
43.74 & 36.59 & -0.53 & -0.47 & -4.13 & -1.21 \\
43.74 & -0.47 & -0.52 & -4.64 & -0.71 & 43.70 & -36.59 & -0.64 & -1.04 & 43.72 & -0.65 & -0.64 & 46.10 & -15.20 & 46.17
\end{bmatrix};
\]

\[
[C_{red}] = \begin{bmatrix}
43.74 & -36.59 & -0.53 & -0.47 & -4.13 & -1.21 \\
43.74 & -0.47 & -0.52 & -4.64 & -0.71 \\
43.70 & -36.59 & -0.64 & -1.04 \\
43.72 & -0.65 & -0.64 \\
46.10 & -15.20 & \\
46.17 &
\end{bmatrix}
\]

Next before and after applying our equivalent cable harness method the results are compared. The culprit pair is the pair one numerated (1, 2) in Fig. 5, we are interested in the voltage and current of the second pair numerated (3, 4) in Fig. 5.

The near end of conductor one (first pair (culprit pair)) is excited with a constant voltage source of 1 V.

The first pair is activated and we calculate the PL in both cases with comparison with the common-mode loads determined by the modal analysis. The culprit pair is excited with a constant voltage source of 1 V.

Figures 7, 8 show the power loss in the second pair (conductor 3) for NEXT and FEXT successively for the initial model and the complete model.
parameter matrices. Configurations studied which affect the terminal loads and mode is considered; electrically long and the transverse electric and magnetic domain from 1 kHz to 30 MHz where the line is.

Results for this case confirm that equivalent cable harness differences are observed which are possibly due to the two models is a few decibels. In the high frequency some telecommunications networks the length of the wire is.


Network theory for cable of 28 pairs (56 conductors), which have been performed on a 2.5 GHz processor and a 4 GB RAM memory computer. Numerical simulations presented in this paper validate the efficiency and the advantages of the proposed method.

Conflict of interest. The authors declare that they have no conflicts of interest.

REFERENCES

Received 19.11.2021
Accepted 20.01.2022
Published 20.04.2022

Samir Bensiammar ¹, M.Sc. Student, Moussa Lefouili ², PhD, Professor, Soufiane Belkhelfa ³, M.Sc. Student, ¹ Mechatronic Laboratory, University of Jijel, Algeria, e-mail: bensiammar.samir@umc.edu.dz (Corresponding Author), lefouili_moussa@yahoo.fr, sofiankhalil@gmail.com

How to cite this article:
Bensiammar S., Lefouili M., Belkhelfa S. Equivalent cable harness method generalized for predicting the electromagnetic emission of twisted-wire pairs. Electrical Engineering & Electromechanics, 2022, no. 2, pp. 29-34. doi: https://doi.org/10.20998/2074-272X.2022.2.05

34 Electrical Engineering & Electromechanics, 2022, no. 2