Overestimation of Electric Radiated Emissions of Power Line Communication Systems

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Abstract— **In this paper, we study the wave impedance of the radiated emissions due to an injection of power line communications (PLC) signals belonging to a frequency range beginning at $1MHz$ up to $30MHz$ on a low voltage buried cable. As most of regulation working groups consider that the value of the wave impedance for any power line network can be taken equal to the wave impedance of freespace, we will show by simulations based on the antenna theory and a modified version of the finite element method that such an approach leads to an overestimate of the real value of the electric field.**

I. Introduction

High data rates communications in the frequency of some MHz and of the order of a few $Mbits/s$ on the low voltage network have recently been added to the many fields of interest in power line communications for applications such as internet, voice over IP, video, ... The radiated emissions associated to PLC signals have to be characterised for radio broadcasting systems protection.

Most of regulation committees use, for estimating the electric radiated emissions due to PLC signals belonging to the frequency range $[1MHz-30MHz]$ on low voltage network, the wave impedance of freespace, once the magnetic field modulus has been measured (see e.g. [1][2][3]). Hence, the approximation of far field and transverse electromagnetic propagation (TEM) enables one to get an estimate of the electric field modulus $|E|$ (in V/m) knowing the magnetic one $|H|$ (in A/m) and the freespace wave impedance $Z_0 = 377\Omega$ according to equation (1).

$$|E| = Z_0 \, |H| \tag{1}$$

The reason why equation (1) is often used is related to the fact that the electric field is less easily measured and with less accuracy than the magnetic field in the frequency range of $[1MHz-30MHz]$. In fact, monopole antennas are used for the electric field measurement but give only accurate results for the vertical component of the previous one whereas loop antennas give accurate results for the three magnetic field components.

Having an accurate estimate of the electric radiated emissions associated with PLC systems is very important in a normalization context since many regulation authorities propose radiated limits based on equation (1) (see e.g. [4][5][6][7]).

In the following, we will show why we have focused on a common propagation mode and the original model we have developed to take it into account. Then, we will recall some theoretical basic points concerning the methods we used and some simulation results based on the antenna theory and a modified version of the finite element method will be presented. Some experimental data will then attest the validity of our simulation results. We have focused on an outdoor configuration as many else studies have already been done on indoor ones (see e.g. [8][9]).

II. Common propagation mode of a buried cable

Considering the low voltage buried $HN33S33$ cable depicted on figure 1 with the associated caption of the table below, we are going to present an original model we have developed to take into account a common mode propagation for characterising the radiations of this cable.

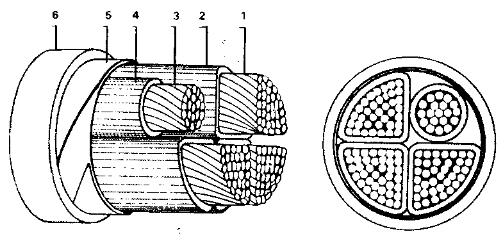

Fig. 1. Low voltage underground $HN33S33$ cable

1	Phase conductor(aluminium)
2	Chemically reticuled polyethylene(CRP)
3	Neutral conductor(aluminium)
4	Sheath of the neutral conductor
5	Shield(steely ribbon)
6	Exterior CRP sheath

In normal operational conditions, a signal is injected between the neutral conductor and a phase conductor thus creating a pure differential establishing mode. In fact, the steely ribbon in galvanic contact with the neutral conductor can be seen as a shield which will mask the majority of the radiated emissions coming from this previous propagation mode. Moreover, the neutral conductor is connected to the ground via earthing electrodes which will create a circulation of PLC signals in a secondary loop integrating the ground. It is the reason why we can consider that all the radiated emissions of this kind of cable come from a common mode propagation.

To take into account the ground, we can consider a coaxial model, the core representing the whole cable and the shield the ground. The ground will then present dielectric losses properties and the core is assumed to be a perfect conductor. The separation between the two previous materials is chemically reticuled polyethylene (CRP). The core radius corresponds to the equivalent cable radius and the shield has been taken in order of the penetration depth in the ground at $1MHz$. Figures 2 and 3 summarize our proposed model for characterising the radiated emissions due to a common mode propagation.

It is this model we are going to simulate via the antenna theory (implemented thanks to the Numerical Electromagnetic Code $NEC4$) and a modified version of the finite element method (implemented through MicroWave Studio MWS code).

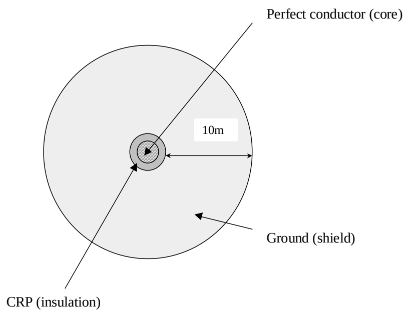

Fig. 2. Coaxial cable model into a common mode approach

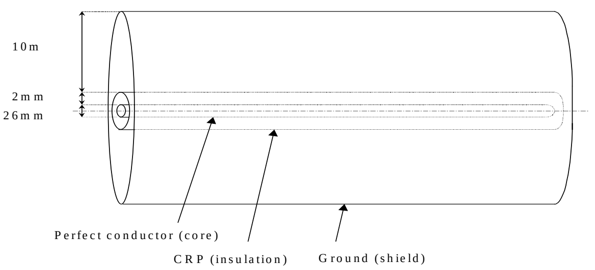

Fig. 3. Coaxial cable model into a common mode approach

III. Theoretical aspects

As it has been mentioned in the previous section, two different approaches have been followed to get some simulation results. Among these, are the antenna theory (method of moments) and a modified version of the finite element method. We are going to recall some theoretical and practical points relative to the application of these two theories.

A. The antenna theory

By using the $NEC4$ code (see [12]), we suppose that the radiating structure is embedded in ground. The ground can be seen in our frequency range as a losses dielectric characterised by a finite conductivity (or a dielectric losses angle) and a relative permittivity. The main assumption in NEC is the thin wire hypothesis which considers that the radiating structure can be modelled as a union of thin wires. A wire can be considered as thin when its radius is neglictible compared to the wavelengths of the injected signals (see [10]).

The only condition one has to check before simulating a given model with NEC, is that the spatial discretization step δ of the structure verifies condition of equation (2) where λ_0 and α are for the injected signal wavelength in vacuum and a ponderation weighting factor, respectively, which typical values can be taken between $1/20$ and $1/10$ (see [11]).

$$\delta \leq \alpha \times \lambda_0 \tag{2}$$

To take into account the presence of a non perfectly conductor ground, we can use the Fresnel coefficients approximation or the rigorous Sommerfeld integrals calculation. In the following results, we used the Sommerfeld integrals. Besides, fields calculations are processed via the method of moments for calculating first the induced currents distribution located as each spatial sampling point of the structure.

B. A modified version of the finite element method

MWS is a complete three dimentionnal Maxwell equations solver based on the finite integration method to discretize the four Maxwell equations. The radiating structure is here discretized according to a mesh composed with elementary bricks which are partially filled with one or two given materials. It is this last property which modifies the classical finite element method formulation (see [13]). This software enables the introduction of lossy materials which is very important to simulate the radiated emissions in ground.

The convergence of the simulation results is determined by the way the structure is spatially discretized. Even if we adopt an adaptive form of meshing, one has to check that a condition similar to equation (2) is verified.

IV. Simulation results

Since the physical coaxial model could not be directly simulated with the $NEC4$ and the MWS codes, we have first adapted them to the capabilities of our simulation tools. We will then present the simulation contexts concerning the studied frequency, the loads values, the injected power and the observation points for the electromagnetic fields. Once the electric field modulus $|E|$ and the magnetic field $|H|$ calculated, we put on a first graphic $|E|$ and $|H| + 51,5$ and we have represented on a second one the difference between $|E|$ and $|H|$. The $51,5dB$ comes from a logarithmic written of equation (1) since $E(dB) = H(dB) + 20\log_{10}(Z_0) = H(dB) + 51,5$. Besides, the electric field modulus has been represented according to a logarithmic scale in $dB\mu V/m$.

A. Simulations with the $NEC4$ code

We have the simulation model for $NEC4$ of figure 4 where the buried cable into a common mode approach has been described by a central insulated conductor and two terminal bare bars which represent the earthing electrodes used in practical conditions. As we can see on figure 4, we consider a monochromatic signals generator with a $5V$ magnitude and an internal impedance of 100Ω located at $x = -50m$ and a terminal resistive load of 100Ω too. The total power delivered by the generator is equal to $P_0 = 14,3dBm$ and we assume, by using the NEC code, that the excitation source is perfectly matched to the impedance presented by the injection port. Hence, this means that we approximate the power really transmitted to the buried cable to P_0. We have chosen a $100m$ long cable buried in a ground presented a conductivity $\sigma = 10^{-3}S/m$ and a dielectric constant $\varepsilon = 15$ which correspond to a normal soil characteristics.

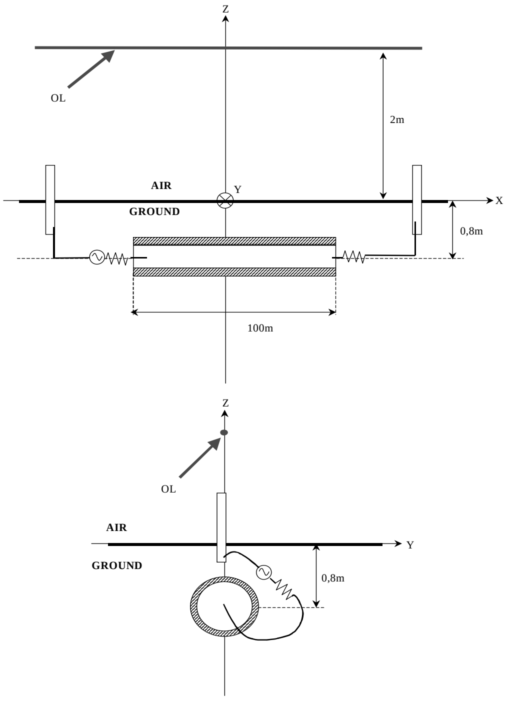

Fig. 4. $NEC4$ simulation model

We have calculated the electromagnetic fields according to the observation line defined by equation (3) and for the following studied frequencies $4MHz$, $10MHz$ and $17MHz$. We obtained the results of figures 5 and 6.

$$OL = \left\{(x, y, z) \in R^3 \mid -50m \leq x \leq 50m, y = 0m, z = 2m\right\} \quad (3)$$

B. Simulations with MWS code

We have considered the $50m$ long cable described on figure 7. Then, we have calculated the electromagnetic fields according to the observation line defined by equation (4) and at the following studied frequencies $4MHz$, $10MHz$ and $17MHz$. We obtained the results of figures 8 and 9 for an injected power of $30dBm$.

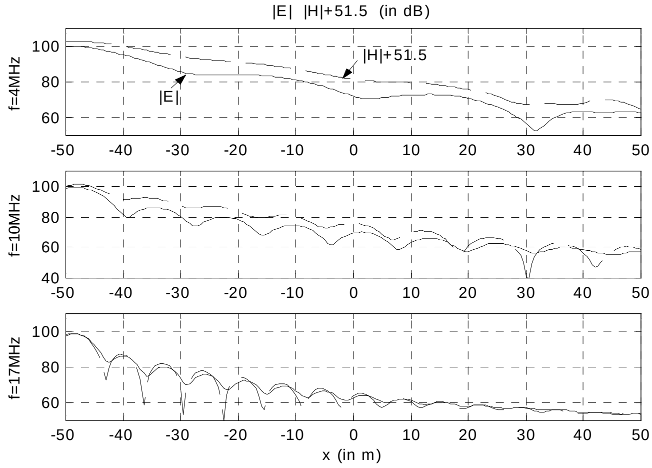

Fig. 5. $|E|$ and $|H| + 51,5$ calculations (in $dB\mu V/m$)

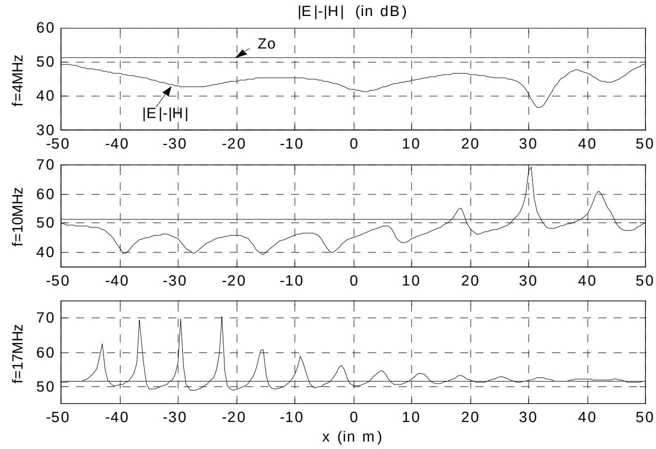

Fig. 6. Calculation of $Z = |E| - |H|$, comparaison with Z_0

$$OL = \left\{(x, y, z) \in R^3 \mid 0m \leq x \leq 50m, y = 2m, z = 2m\right\} \quad (4)$$

On the contrary to the $NEC4$ code, MWS does not suppose that the voltage source located at the injection port is perfectly matched. MWS is also used in microwaves area to characterize the scattering parameters of structures such as waveguides, cavities or electric lines. It is the reason why the value of the reflexion coefficient, due to the impedance mismatch of the internal 50Ω generator and the injection port impedance, is evaluated in MWS. Hence, the knowledge of this reflexion coefficient and the total delivered power ($30dBm$) enables one to calculate the injected power.

C. Wave impedance of the simulated radiated emissions

We can first remark, according to figures 5 and 8, that we have $|E| \leq |H| + 51,5$ for the frequencies beginning at

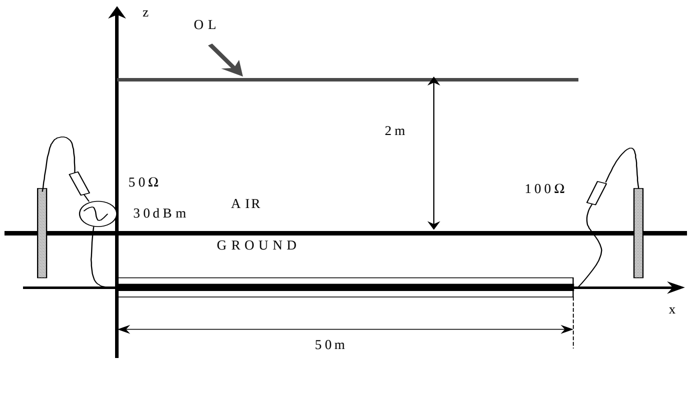

Fig. 7. MWS simulation model

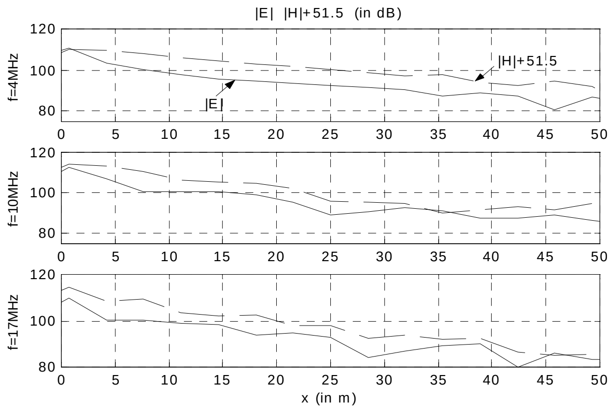

Fig. 8. $|E|$ and $|H| + 51,5$ calculations (in $dB\mu V/m$)

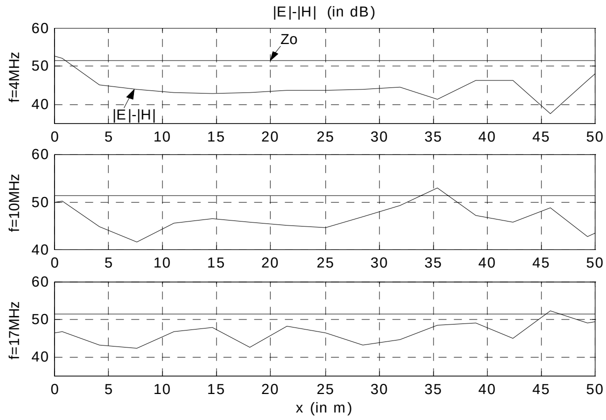

Fig. 9. Calculation of $Z = |E| - |H|$, comparaison with Z_0

$1MHz$ up to $17MHz$. Hence, we have $Z = |E| - |H| \leq Z_0 = 51,5$ (see figures 6 and 9) which shows that the wave impedance of the radiated emissions Z is smaller than the value of the freespace impedance Z_0. This result means that by using Z_0 and $|H|$ for giving an estimate of $|E|$ leads to an overestimation of the real value of $|E|$.

A second important point is relative to the evolution of Z versus frequency. We note that the difference $Z_0 - Z$ is a decreasing function of frequency and that we have $Z \leq Z_0$ for frequencies belonging to $[1MHz - 17MHz]$ and $Z \approx Z_0$ for frequencies belonging to $[17MHz - 30MHz]$. This means that we consider that we are in far fields conditions for frequencies higher than $17MHz$. The freespace wavelength λ_0 is equal to $17m$ at $17MHz$. We suppose that we are in far fields conditions as soon as the distance d between the radiating source and the observation point is bigger than $\frac{\lambda_0}{2\pi}$. In our $NEC4$ simulations, $\frac{\lambda_0}{2\pi} \approx 1,6m$ and $2m \leq d \leq 140m$. For the MWS simulations, we still have $\frac{\lambda_0}{2\pi} \approx 2m$ at $17MHz$ but $2m \leq d \leq 50m$. In both cases, we have $d \geq \frac{\lambda_0}{2\pi}$ which means that we are in far fields conditions and it implies that $Z \approx Z_0$.

The previous simulation results show us that approximating the wave impedance of the radiated emissions of PLC systems, using buried cables as a communication medium, to the value of the freespace impedance can be considered as true for frequencies higher than the transition frequency $17MHz$ which is related to the chosen calculation line for the electromagnetic fields in our simulations and does not correspond to a well-known physical phenomenon. But for the frequency range $[1MHz - 17MHz]$, we have $Z \leq Z_0$.

V. Measurements on a $HN33S33$ cable

A. Experimental conditions

We have taken a $100m$ long buried $HN33S33$ cable at a $80cm$ depth in a chosen site in Paris suburb such that there were no other cables or lines in its near neighbourhood, hence avoiding parasitic couplings with them. We added two earthing electrodes at each extremity of the cable to practise a signal injection between one electrode and the neutral conductor of the cable of a high frequency signal belonging to $[1MHz - 30MHz]$, a resistive load has been placed at the terminal connexion port for the same common mode.

An important practical point is relative to the way the injection is done as we have a quite important impedance mismatch between the internal impedance of 50Ω of the Rhode&Schwarz signal generator we used and the injection port impedance. We had to cope with this issue since, for a total power delivered by the generator, we had a transmitted proportion and a reflected proportion of the total delivered power. All the measurements done for electromagnetic radiations on PLC networks have been classically realized by other laboratories by inserting a network analyzer in the measurement setup to quantify the reflexion coefficient at the injection port. Therefore, an addition of a wideband amplifier was necessary to produce a sufficient injected power for significant electromagnetic radiations. We decided to not follow this usual approach as we did not use a amplifier whereas we added a coupling device composed with variable inductances and capacitances selectable by the user and providing an adaptation at the injection port. This adaptive coupling was adjusted for one given frequency.

Another important point concerns the used sensors to get some valuable measurements of the electric and magnetic fields components. We used a $HFH2Z2$ (Rohde&Schwarz) loop antenna connected to a $HP4395A$ (Hewlett-Packard) spectrum analyzer to measure the three magnetic components and a dipole and a rod $HFH2Z1$ (Rohde&Schwarz) antenna for the electric field.

The electric field has been measured according to the cable axis, equation (5) gives a representation of this measurement line.

$$ML = \left\{(x, y, z) \in R^3 \mid 0m \leq x \leq 100m, y = 0m, z = 2m\right\} \tag{5}$$

For example, we have obtained the results of figures 10 and 11 for the simulated and measured, respectively, vertical component of the electric field. These simulated and measured results give the radiations levels for an injected power of $15dBm$ and a frequency of $17MHz$.

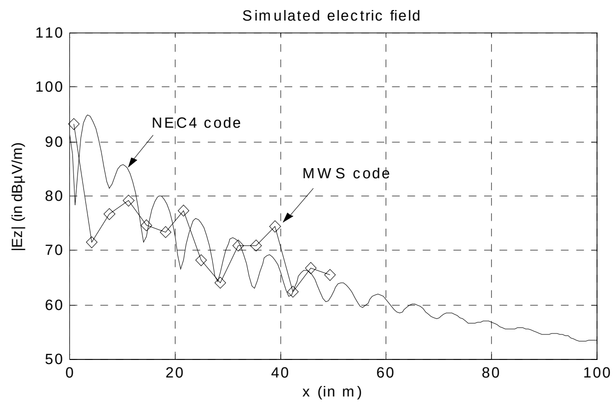

Fig. 10. Simulated field at $17MHz$

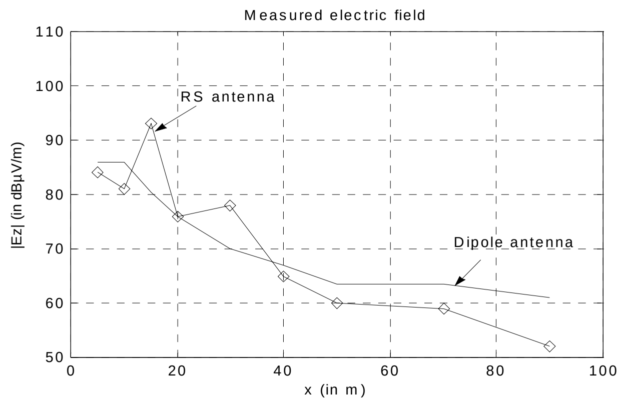

Fig. 11. Measured field at $17MHz$

VI. Conclusions

In this paper, we showed that the classical freespace wave impedance used by regulation authorities leads to an overestimation ($8dB$, $5dB$ and $1,5dB$ at $4MHz$, $10MHz$ and $17MHz$, respectively) of the real value of the electric radiations. The studied radiations were caused by a low voltage underground cable used to transmit data thanks to power line communications signals belonging to a frequency range beginning at $1MHz$ up to $30MHz$.

Some numerical simulations have been done with two different methods. A first one is based on the antenna theory ($NEC4$ code implementation) whereas the second one uses a modified version of the finite element method (MWS code implementation). Both of them enables us to conclude that the wave impedance of the radiated emissions Z is less important than the freespace wave impedance Z_0 for frequencies beginning at $1MHz$ up to $17MHz$. Between $17MHz$ and $30MHz$, approximating Z to Z_0 does not produce a big estimation error.

Some experimental data validated our simulation results with a good accuracy.

VII. Acknowledgement

We would like to gratefully thank Jerry Burke from the Lawrence Livermore Laboratory for having simulated our model, for his helpful advices and his kindness.

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