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Bottom-up single-wire power line communication channel modeling considering dispersive soil characteristics

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ABSTRACT

This paper investigates the influence of frequency-dependent soil conductivity and permittivity on narrowband and broadband power line communication (PLC) channel modeling using transmission line theory. For narrowband PLC channels, the variation of the transfer function *H*(*f*) of a single overhead wire over a finitely conducting ground due to frequency-dependent soil parameters is minimal for low-resistivity soils, but more relevant for high-resistivity soils in the upper frequency range. For broadband PLC channels, *H*(*f*) strongly depends on the soil model for frequencies below 10 MHz or so, but is relatively insensitive to this parameter at higher frequencies. The average channel attenuation is more affected by soil model and soil parameters than the achievable data rate. Also, the voltage definition assumed in the derivation of the per-unit-length parameters of the line has little influence on the results. A comparison between different formulations traditionally used in PLC channel modeling indicates that the admittance associated with a finitely conducting ground should not be neglected in the modeling of broadband PLC channels.

1. Introduction

With the advent of the concept of smart grids and smart cities, interest in power line communication (PLC) has grown in applications such as remote load control, remote metering, and internet of things [1–3]. Along with other technologies, PLC is one of the solutions that can be exploited in a cooperative scenario that is expected to provide greater flexibility, reliability and coverage of communication over electric power grids [3].

One of the possibilities of modeling a PLC channel consists in the so-called bottom-up approach, in which the channel model is obtained from the application of transmission line theory [4–12]. For accurately determining signal attenuation levels and achievable data rates in a PLC channel using the bottom-up approach, it is first necessary to calculate the per-unit-length parameters of the line including the influence of a finitely conducting ground, for which many expressions are available [13–24].

For narrowband applications, which involve frequencies up to 500 kHz [2], it is expected that Carson's theory [14] be sufficiently accurate for the calculation of the ground return impedance as long as displacement currents in the ground are negligible. Carson's equations are used, for example, in [4] for studying PLC channels in medium-voltage distribution lines. Other possibilities were also considered in narrowband PLC channel modeling (e.g., [5,11]), such as Sunde's equations [15], which other than Carson's equations allow the selection of an arbitrary value of relative ground permittivity [23], and Pollaczek's equations [13], which can be considered equivalent to Carson's equations for overhead line modeling [23].

For broadband applications, which involve frequencies in the range 1.7–100 MHz [2], Carson's and Pollaczek's equations no longer hold and other formulations should be used. For example, the logarithmic approximation of D'Amore and Sarto [20] is used in [9,12] for calculating the per-unit-length parameters of medium-voltage distribution lines used as PLC channels. This formulation allows considering both the impedance and admittance associated with a finitely conducting ground, which are necessary for accurate transmission line modeling in the high-frequency range [17–24]. In spite of that, the influence of the admittance associated with a finitely conducting ground was neglected in [7] when investigating the efficacy of using ground return in broadband power line communications. In this particular reference, only the ground return impedance was calculated using Sunde's equations [15] for frequencies up to 100 MHz.

As seen from the discussion above, there seems to be no consensus about how to include the ground influence on PLC channel modeling. In addition, even though the conductivity and permittivity of the soil are known to vary with frequency [25–30], all referred papers assumed

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constant soil electrical parameters in the estimation of the impedance and admittance associated with a finitely conducting ground. Although a recent experimental study was presented to give support to the modeling of a narrowband PLC channel considering the dispersive nature of the soil [31], the analysis was restricted to low-resistivity soils of 10 Ωm and 100 Ωm for frequencies up to few hundreds of kHz, in which case the variation of the ground conductivity and permittivity with frequency is known to have a minimal effect on the line propagation characteristics [32]. This means that it is still necessary to investigate to what extent considering dispersive soil parameters will affect the attenuation level and the achievable data rate in a PLC channel for a large range of soil characteristics in both narrowband and wideband applications.

In this paper, a theoretical study is presented to investigate the impact of dispersive soil parameters on the modeling of both narrowband and broadband PLC channels in a medium-voltage distribution line. In the analysis, a single phase conductor is assumed and different formulations are considered for the calculation of the impedance and admittance associated with a finitely conducting ground. Besides emphasizing the impact of ground losses on the simulated PLC channel including frequency-dependent soil parameters, assuming a single conductor with earth return is able to provide information about the ground mode in the modeling of multiconductor PLC channels. It also gives information for the modeling of PLC channels in single wire earth return (SWER) systems, which are used in many countries to supply electric power to remote areas at low cost [33].

This paper is organized as follows. Section 2 presents the models and solution method assumed in this study. It also discusses the formulation used to estimate the average attenuation and the achievable data rate in the simulated PLC channel. Results and analysis are presented in Section 3, followed by conclusions in Section 4.

2. Modeling assumptions

2.1. Transmission line modeling

The full-wave approach is a technique that can be used for determining transmission line parameters in a wide frequency range [16,18–22,24]. Assuming a wire of radius r and infinite length parallel to coordinate x at height $y = h$ above an imperfectly conducting ground, and considering a harmonic current in the form $I = I_m e^{-\gamma x}$, where I_m is the current amplitude and γ is an unknown propagation constant, it is possible to write the so-called modal equation at the angular frequency ω from the application of Maxwell's equations and consideration of the boundary conditions at the air–ground and air-–wire interfaces [16,18–22,24]. The numerical solution of the modal equation yields values for γ that can be used to determine the per-unit-length parameters of the line. However, these parameters are sensitive to the voltage definition considered in their derivation. Possible voltage definitions are the line integral of the vertical electric field from the ground surface to the wire [19,21], the potential difference between the wire and the ground surface [21], the wire scalar potential with the reference at infinity [21], or the line integral of the vertical electric field from a reference plane at an infinite distance above or below the ground surface to the wire [19]. Here, two different voltage definitions are considered. One is given by (1), in which the voltage $U_i(x)$ is assumed to correspond to the wire scalar potential $\varphi(x, y = h - r)$ with the reference at infinity. The other is given by (2), in which the voltage $U_{ii}(x)$ is obtained from the line integral of the vertical electric field from the ground surface to the wire. In this expression, A_y is the magnetic vector potential at direction y, $\varphi(x, y = h - r)$ is the wire potential, and $\varphi(x, y = 0)$ is the ground surface potential (see [21,24] for details). Eq. (2) is considered more rigorous than (1) at high frequencies because it includes the contribution of A_y [24].

$$U_i(x) = \varphi(x, h - r) \tag{1}$$

$$U_{ii}(x) = \varphi(x, h - r) - \varphi(x, 0) + j\omega \int_0^{h-r} A_y(x, y')dy' \tag{2}$$

Calculating the propagation constant from the numerical solution of the modal equation resulting from the full-wave model is cumbersome and computationally inefficient [24]. To circumvent this difficulty, the so-called quasi-TEM approximation can be used. It consists in assuming that the unknown propagation constant γ is equal to the intrinsic propagation constant of the medium in which the wire is immersed [21,24]. Using voltage definition $U_i(x)$, Pettersson [21] arrived at (3) and (4) for calculating the per-unit-length impedance and admittance of an overhead wire. On the other hand, considering voltage definition $U_{ii}(x)$, Pettersson obtained (5) and (6) [21].

$$Z_{U_i} = Z_{int} + \frac{j\omega\mu_0}{2\pi}(M + S_1) \tag{3}$$

$$Y_{U_i} = j\omega 2\pi\varepsilon_0 (M + S_2)^{-1} \tag{4}$$

$$Z_{U_{ii}} = Z_{int} + \frac{j\omega\mu_0}{2\pi}[M + S_1 - (T + S_2)] \tag{5}$$

$$Y_{U_{ii}} = j\omega 2\pi\varepsilon_0 (M - T)^{-1} \tag{6}$$

If $r \ll h$ and a non-magnetic wire is assumed, the parameters in (3)–(6) read [21,24]

$$M = ln\left(\frac{2h}{r}\right) \tag{7}$$

$$Z_{int} = \frac{1}{2\pi r}\sqrt{\frac{j\omega\mu_0}{\sigma}}\frac{I_0(r\sqrt{j\omega\mu_0\sigma})}{I_1(r\sqrt{j\omega\mu_0\sigma})} \tag{8}$$

$$S_1 = 2\int_0^\infty \frac{e^{-2h\lambda}}{\lambda + u_2}\cos(r\lambda)d\lambda \tag{9}$$

$$T = 2\int_0^\infty \left(\frac{u_2}{\lambda}\right)\left(\frac{e^{-h\lambda} - e^{-2h\lambda}}{n^2\lambda + u_2}\right)\cos(r\lambda)d\lambda \tag{10}$$

$$S_2 = 2\int_0^\infty \frac{e^{-2h\lambda}}{n^2\lambda + u_2}\cos(r\lambda)d\lambda \tag{11}$$

$$n = \frac{\gamma_2}{\gamma_1} = \frac{\sqrt{j\omega\mu_0(\sigma_g + j\omega\varepsilon_r\varepsilon_0)}}{j\omega\sqrt{\mu_0\varepsilon_0}} \tag{12}$$

$$u_2 = \sqrt{\lambda^2 + \gamma_2^2 - \gamma_1^2} \tag{13}$$

where μ_0 and ε_0 are respectively the permeability and permittivity of the vacuum, σ_g is the ground conductivity, ε_r is the relative permittivity of the ground, σ is the wire conductivity, $\gamma_1 = j\omega\sqrt{\mu_0\varepsilon_0}$, $\gamma_2 = \sqrt{j\omega\mu_0(\sigma_g + j\omega\varepsilon_r\varepsilon_0)}$, and I_0 and I_1 are modified Bessel functions of first kind and orders zero and one, respectively.

In Pettersson's Eqs. (5) and (6), which are based on the rigorous voltage definition given by (2), the correction of the per-unit-length impedance and admittance due to a finitely conducting ground is written in terms of S_1, S_2, and T. As discussed in [21], these terms extend the validity of the transmission line equations to frequencies in the range of tens of MHz, which are suitable to the analysis of broadband PLC channels. If $A_y = 0$ and $\varphi(x, y = 0) = 0$ are assumed in (2), that is, if the voltage definition given by (1) is considered, equations (5) and (6) reduce to (3) and (4), which can be shown to be equivalent to the integral equations proposed by Nakagawa [17] if the magnetic permeability of media 1 and 2 are equal. This means that Nakagawa's equations consider the impedance (S_1) and admittance (S_2) corrections due to a finitely conducting ground, but implicitly assume voltage as equal to the wire potential. If it is further assumed that $S_2 = 0$, then the correction due to a finitely conducting ground is restricted to S_1 in (3). It can be shown that this term reduces to Carson's integral equation