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A Comparison Between Orthogonal and Non-Orthogonal Multiple Access in Cooperative Relaying Power Line Communication Systems

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ABSTRACT Most, if not all, existing studies on power line communication (PLC) systems as well as industrial PLC standards are based on orthogonal multiple access schemes, such as orthogonal frequency-division multiplexing and code-division multiple access. In this paper, we propose non-orthogonal multiple access (NOMA) for decode-and-forward cooperative relaying PLC systems to achieve higher throughput and improve user fairness. To quantitatively characterize the proposed system performance, we also study conventional cooperative relaying (CCR) PLC systems. We evaluate the performance of the two systems in terms of the average capacity. In this respect, accurate analytical expressions for the average capacity are derived and validated with Monte Carlo simulations. The impact of several system parameters, such as the branching, impulsive noise probability, cable lengths, the power allocation coefficients, and input signal-to-noise ratio, is investigated. The results reveal that the performance of the proposed NOMA-PLC scheme is superior compared with that of the CCR-PLC system. It is also shown that the NOMA-PLC system can be more effective in reducing electromagnetic compatibility associated with PLC and that increasing the network branches can considerably degrade the performance. Moreover, optimizing the power allocation coefficients is found to be of utmost importance to maximize the performance of the proposed system.

INDEX TERMS Average capacity, conventional cooperative relaying (CCR), decode-and-forward (DF) relaying, impulsive noise, log-normal fading, non-orthogonal multiple access (NOMA), power line communication (PLC).

I. INTRODUCTION

Power line communication (PLC) technology is increasingly having a great potential in future home-networking as well as in the realization of many smart grid applications. The fact that PLC is a retrofit technology, utilizing the existing power line network, can significantly reduce deployment costs. In addition, PLC can be more appealing than wireless technologies in several environments where the wireless signal suffers from high losses and attenuation such as in underground structures and in buildings with concrete and metal walls [1]–[3]. For instance, PLC can be used in multi-store buildings to extend the Wi-Fi coverage by simply using a pair of PLC modems. Another good example is advanced smart metering. It is well-acknowledged that most electricity meters tend to be in basements, which are very unfavorable environments for wireless/mobile broadband signals; consequently, PLC becomes a complementary solution. What's more, as far as security and privacy is concerned, PLC is more attractive to smart grid developers since PLC is a *through-grid* technology. To elaborate, this feature can considerably reduce reliance of utility companies on third party connectivity, which of course will reduce security and privacy issues.

On the other hand, however, this massive infrastructure was never designed to carry communication signals at higher frequencies. As a consequence, communication signals over power lines can be significantly degraded due to various impairments such as non-Gaussian noise, frequency-selectivity and high frequency/distance dependent attenuation [4]–[7]. Furthermore, what makes reliable PLC systems more challenging is the limited transmit power due to electromagnetic compatibility restrictions, resulting in very small signal-to-noise ratio (SNR) values at the receiver [8].

To reduce the severity of the non-Gaussian noise and frequency-selectivity, multi-carrier transmission systems such as orthogonal frequency-division multiplexing (OFDM) are widely adopted and used in most industrial standards [9], [10]. Along with this, relaying schemes, such as amplify-and-forward (AF) or decode-and-forward (DF), are usually implemented to alleviate the attenuation and limited transmit power constraints. In relaying PLC systems, multiple nodes along the transmission route can overhear a transmitted signal and are able to help in transmission, which is commonly referred to as cooperative relaying [11]–[13]. Taking a dual-hop relaying PLC system as an example, the destination modem can receive and combine two copies of the same source signal: one via the direct source-destination link and the other is via the relayed version. This is achieved over two transmission phases, i.e., using time-division multiplexing (TDM). Although this apparently enhances the quality of the received signal at the destination, this approach can severely limit the spectral efficiency of PLC systems because of the duplicated transmission of the same signal during the two time slots. To overcome this, we propose in this paper non-orthogonal multiple access (NOMA)[1] for PLC networks; more specifically, dual-hop cooperative DF relaying PLC is considered. It should be pointed out that NOMA-PLC proposed here is different from that in future radio access systems, proposed in [17], because of the different nature of the two interfaces. The advantages of using NOMA compared to orthogonal multiple access (OMA) systems, such as OFDM and TDM systems, include the following

- NOMA is able to provide higher system throughput. This is because a transmitting modem in NOMA can simultaneously transmit multiple data signals to different users with different power levels while each user, at the same time, occupies the entire available frequency band [18], [19].
- NOMA offers better fairness between different users, which allows serving numerous devices (users) at a single channel use. This can be particularly appealing in PLC applications where different devices communicate with different data rate requirements, simultaneously.
- NOMA in PLC, as will be shown later, can considerably reduce the severity of electromagnetic compatibility issues associated with PLCs. Hence, better coexistence with other wireless systems can be maintained.

[1]NOMA has recently been introduced in wireless communications, see e.g., [14]–[16] and the references therein.

The main contributions of this paper are as follows. First, we derive an accurate analytical expression for the average capacity of the proposed NOMA-PLC system. Second, to highlight the achievable gains attained with our proposal, we study the achievable capacity performance of the OMA system, referred to here as conventional cooperative relaying PLC (CCR-PLC). Third, we investigate the impact of various system parameters on the system performance and provide Monte Carlo simulations throughout to verify the accuracy of our analysis. We also discuss how, and by how much, different system parameters impact the achievable performance. It is worth pointing out that the PLC channel in our evaluations is assumed to have log-normal distribution, cf., e.g., [13], [20]–[23], and that the frequency/distance dependent attenuation is characterized using Zimmermann and Dostert's model reported in [24].

The rest of this paper is organized as follows. Section II elaborates the related work on relaying PLC systems. Section III describes the system model used in this study. In Section IV, we derive accurate analytical expressions for the average capacity of the proposed NOMA-PLC and CCR-PLC systems. After that, numerical examples are presented and discussed in Section V. Finally, Section VI concludes the paper and highlights the main findings.

II. RELATED WORK

Over the last decade, several wireless-based technologies, including relaying systems, have found their way into PLC [25]–[27]. Relays, or repeaters, bridge the source-to-destination distance and can diminish the impact of most PLC channel impairments offering considerable throughput improvements and range extension. Amongst many relaying protocols reported in the open literature, AF and DF are the two most extensively studied in PLC networks. For instance, in 2010, Tonello *et al.* [26] proposed cooperative opportunistic DF relaying for PLC systems and used TDMA at the medium access control (MAC) layer. The source and relay time slot durations were optimized and significant capacity enhancements were obtained in comparison to the direct-link approach. In 2011, the authors of [28] proposed beamforming in multi-hop AF and DF relaying PLC systems. In 2013, Cheng *et al.*, [29], investigated the capacity performance of a dua-hop AF PLC system from an information-theoretic perspective where they analyzed the capacity bounds and addressed the optimization problem of the source and relay transmit power. The first proposal of bi-directional relaying in PLC appeared in [30]. In 2015, Dubey and Mallik [21] studied the end-to-end bit error rate and average capacity performances of a multi-hop AF PLC network and derived approximate analytical expressions for the two performance metrics. The authors assumed in this work frequency-division multiplexing (FDM) and that PLC modems can work in full-duplex mode.

Furthermore, fountain-and-forward (FCF) relaying, introduced first for wireless systems [31], has been examined in the context of PLC, first in [32] and later in [33]. In 2014, Valencia *et al.*, [34], considered cooperative AF and DF PLC with three receiver combining techniques, namely, selection combining (SC), equal gain combining (EGC) and maximum ratio combining (MRC). Later, in 2015, Ezzine *et al.* [35] provided a comparative study of the capacity performance for AF, DF and FCF relaying protocols in OFDM-based dual-hop relaying PLC systems. Very recently, Facina *et al.* [36] have conducted extensive measurement campaigns and presented interesting results. For instance, they showed that AF relaying in PLC does not always enhance performance and that DF relaying can, by far, have better performance than the former approach. Also, very recently, Rabie *et al.* [23], [37], [38] have introduced the concept of energy-harvesting to AF and DF relaying PLCs, and have demonstrated that further capacity improvements are achievable.

To the best of the authors' knowledge, all the aforementioned works are based on OMA transmission, either OFDM or/and TDMA, which can be spectral inefficient as briefly discussed in the introduction. In contrast, this paper studies NOMA over PLC channels. More specifically, a dual-hop cooperative relaying PLC system is considered with MRC processing implemented at the destination.

III. SYSTEM MODEL

The following notations are used hereafter. The notation $|\cdot|$ is the magnitude operator and $\mathbb{E}[\cdot]$ is the statistical expectation. The notation $f(\cdot)$ denotes the probability density function (PDF), while $F(\cdot)$ represents the cumulative distribution function (CDF).

The system model used in this study consists of a source modem and two receiving modems. One receiving modem acts also as a relay, while the other as a destination. The source-to-destination, source-to-relay and relay-to-destination channel gains are represented by h_{sd}, h_{sr}, and h_{rd}, respectively, with corresponding distances d_{sd}, d_{sr}, and d_{rd}. The PLC channel gains are assumed to be log-normally distributed, [39], [40], the PDF of which is given by

$$f(h_i) = \frac{1}{\sqrt{2\pi}\sigma_i h_i} \exp\left[-\frac{(\ln(h_i)-\mu_i)^2}{2\sigma_i^2}\right], \tag{1}$$

where the subscript i is defined as $i \in \{\mathrm{sd, sr, rd}\}$, μ_i and σ_i are the mean and the standard deviation of $10\log_{10}(h_i)$ (both in decibels), respectively. It should be noted that the adoption of a log-normal statistical model accounts for the random effects introduced by branches and unmatched loads that are present in the ensemble of network topologies. In addition, the frequency- and distance-dependent attenuation is incorporated into our system model, and this will be denoted by $A_i(f, d_i)$ where f is the operating frequency [24], [41]–[43].

Let P_{s} denote the total source transmit power. The proposed NOMA cooperative relaying PLC system accomplishes communication over two time slots. During the first time slot, the source modem transmits the signal $\sqrt{a_1 P_{\mathrm{s}}}s_1 + \sqrt{a_2 P_{\mathrm{s}}}s_2$ to the relay and destination, where s_k is the k-th data message, $k \in \{1,2\}$, $\mathbb{E}[|s_k|^2] = 1$, and a_1 and a_2 denote the power allocation coefficients for the destination and relaying devices, respectively. It is assumed that $a_1 > a_2$ and that $a_1 + a_2 = 1$. It is also assumed that there may exist a direct link between the two end modems. With this in mind, the received signals at the relay and destination during the first time slot can be expressed as

$$y_{\mathrm{sr}} = \left(\sqrt{a_1 P_{\mathrm{s}}}s_1 + \sqrt{a_2 P_{\mathrm{s}}}s_2\right) A_{\mathrm{sr}}(f, d_{\mathrm{sr}})\, h_{\mathrm{sr}} + n_{\mathrm{r}}, \text{ and} \tag{2}$$

$$y_{\mathrm{sd}} = \left(\sqrt{a_1 P_{\mathrm{s}}}s_1 + \sqrt{a_2 P_{\mathrm{s}}}s_2\right) A_{\mathrm{sd}}(f, d_{\mathrm{sd}})\, h_{\mathrm{sd}} + n_{\mathrm{d}}, \tag{3}$$

respectively, where n_{r} and n_{d} are the noise at the relaying and destination modems, respectively, consisting of two components, namely, background and impulsive [44].

At the relaying device, the signal s_1 is decoded while treating s_2 as noise, and then canceled using successive interference cancellation (SIC) to obtain s_2. Therefore, with perfect cancellation [45]–[48], the SNRs of s_1 and s_2 at the relaying modem can be respectively written as

$$\gamma_{\mathrm{sr}}^{(1)} = \frac{a_1 P_{\mathrm{s}} A_{\mathrm{sr}}(f, d_{\mathrm{sr}})^2 h_{\mathrm{sr}}^2}{a_2 P_{\mathrm{s}} A_{\mathrm{sr}}(f, d_{\mathrm{sr}})^2 h_{\mathrm{sr}}^2 + \sigma_{\mathrm{r}}^2} \text{ and} \tag{4}$$

$$\gamma_{\mathrm{sr}}^{(2)} = \frac{a_2 P_{\mathrm{s}} A_{\mathrm{sr}}(f, d_{\mathrm{sr}})^2 h_{\mathrm{sr}}^2}{\sigma_{\mathrm{r}}^2}. \tag{5}$$

It should be noted that co-device interference in NOMA-PLC leads to some devices being served with low data rates which is perfectly acceptable in many control and monitoring PLC applications. These applications may include home automation and sensor networks such as intruder alarms, fire detection, detection of gas leaks etc [49]. Unlike OMA-PLC designs in which tens of devices and sensors within a home are expected to be interacting using tens of bandwidth channels, NOMA-PLC accomplishes this at a single channel use, resulting in a more spectral efficiency.

During the first time slot, the destination modem will treat the signal s_2 as noise, and therefore the SNR for s_1 can be given by

$$\gamma_{\mathrm{sd}} = \frac{a_1 P_{\mathrm{s}} A_{\mathrm{sd}}(f, d_{\mathrm{sd}})^2 h_{\mathrm{sd}}^2}{a_2 P_{\mathrm{s}} A_{\mathrm{sd}}(f, d_{\mathrm{sd}})^2 h_{\mathrm{sd}}^2 + \sigma_{\mathrm{d}}^2}. \tag{6}$$

In the second time slot, and assuming that the relaying node successfully decodes s_2, the relay forwards this signal to the destination with transmit power P_{r}. Hence, the received signal at the destination modem during the second time slot can be expressed as

$$y_{\mathrm{rd}} = \sqrt{P_{\mathrm{r}}}s_2 A_{\mathrm{rd}}(f, d_{\mathrm{rd}})\, h_{\mathrm{rd}} + n_{\mathrm{d}}, \tag{7}$$

and the corresponding SNR is simply given by

$$\gamma_{\mathrm{rd}} = \frac{P_{\mathrm{r}} A_{\mathrm{rd}}(f, d_{\mathrm{rd}})^2 h_{\mathrm{rd}}^2}{\sigma_{\mathrm{d}}^2}. \tag{8}$$

It is clear that the proposed NOMA-PLC system allows transmitting two symbols over two time slots. This is more spectral efficient than the CCR-PLC approach in which only one symbol is conveyed over two time slots [37].

Different from conventional communication systems, PLC channels are subject to a mixture of both background and impulsive noise. Therefore, the capacity is calculated depending on the coding/decoding strategy implemented, namely, i) erasure decoding in which the samples affected by impulsive noise are disregarded and ii) non-erasure decoding, i.e., full decoding; both are given respectively as [50]

$$C_{\text{easure}} = (1-p)\,C_{\text{b}} \tag{9}$$

and

$$C_{\text{non-easure}} = (1-p)\,C_{\text{b}} + pC_{\text{im}}, \tag{10}$$

where p is the probability occurrence of impulsive noise whereas C_{b} and C_{im} represent the capacity of a channel affected by Gaussian noise and impulsive noise, respectively. For the sake of simplifying our analysis, we consider below the erasure coding capacity only.

IV. PERFORMANCE ANALYSIS

In this section, we analyze the average sum capacity of both the proposed NOMA-PLC system and the CCR-PLC approach. Without loss of generality, we assumed in this analysis that $\sigma = \sigma_{\text{r}} = \sigma_{\text{d}}$, $P = P_{\text{s}} = P_{\text{r}}$ and that $\rho = P/\sigma^2$, where ρ basically represents the transmit SNR.

A. AVERAGE SUM CAPACITY OF THE PROPOSED NOMA-PLC SYSTEM

The sum rate of the proposed system, C_{p}, consists of the rates associated with both s_1, denoted as C_1, and s_2, denoted as C_2, i.e.,

$$C_{\text{p}} = C_1 + C_2. \tag{11}$$

Because s_1 needs to be decoded at the relaying modem (for SIC) as well as at the destination, the achievable rate associated with this signal can be given as

$$C_1 = \frac{1}{2}\min \times \left\{ (1-p)\log_2\left(1+\gamma_{\text{sr}}^{(1)}\right), (1-p)\log_2\left(1+\gamma_{\text{sd}}\right) \right\} \tag{12}$$

On the other hand, since the end-to-end capacity of a DF-based system is determined by the weakest link capacity [51], [52], the instantaneous capacity associated with s_1 can be determined as

$$C_2 = \frac{1}{2}\min \times \left\{ (1-p)\log_2\left(1+\gamma_{\text{sr}}^{(2)}\right), (1-p)\log_2\left(1+\gamma_{\text{rd}}\right) \right\}. \tag{13}$$

Note that the factor of $\frac{1}{2}$ in (12) and (13) is due to the fact that the transmission occurs over two time slots.

To begin with, using (4) and (6), we can rewrite (12) as in (14), shown at the bottom of this page. This is obtained with the help of the substitutions $a_1 = 1 - a_2$ and $\rho = P/\sigma^2$.

To derive the average capacity $C_{1,1}$ in (14), we let

$$\mathcal{X} = \min\left\{ \underbrace{\rho A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2}_{X_{\text{sr}}}, \underbrace{\rho A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2}_{X_{\text{sd}}} \right\}, \tag{15}$$

and by implementing order statistics, the corresponding CDF of $\mathcal{X}$ can be calculated using [52]

$$F_{\mathcal{X}}(u) = 1 - \int_u^\infty f_{X_{\text{sr}}}(x)\,\mathrm{d}x \int_u^\infty f_{X_{\text{sd}}}(y)\,\mathrm{d}y \tag{16}$$

where $f_{X_{\text{sr}}}(\cdot)$ and $f_{X_{\text{sd}}}(\cdot)$ are the PDFs of the random variables (RVs) $X_{\text{sr}} = \rho A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2$ and $X_{\text{sd}} = \rho A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2$, respectively, both are given in (15). Since h_{sr}^2 and h_{sd}^2 are log-normally distributed,

$$\begin{aligned}
C_1 &= \frac{1}{2}(1-p)\log_2\left(1 + \frac{\min\left\{A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2, A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2\right\} a_1 P}{\min\left\{A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2, A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2\right\} a_2 P + \sigma^2}\right) \\
&= \frac{1}{2}(1-p)\log_2\left(\frac{\min\left\{A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2, A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2\right\} P + \sigma^2}{\min\left\{A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2, A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2\right\} a_2 P + \sigma^2}\right) \\
&= \underbrace{\frac{1}{2}(1-p)\log_2\left(1 + \min\left\{A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2, A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2\right\}\rho\right)}_{C_{1,1}} \\
&\quad - \underbrace{\frac{1}{2}(1-p)\log_2\left(1 + \min\left\{A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2, A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2\right\} a_2\rho\right)}_{C_{1,2}}
\end{aligned} \tag{14}$$

i.e., $h_{\rm sr}^2 \sim \ln\mathcal{N}\left(2\mu_{\rm sr}, 4\sigma_{\rm sr}^2\right)$ and $h_{\rm sd}^2 \sim \ln\mathcal{N}\left(2\mu_{\rm sd}, 4\sigma_{\rm sd}^2\right)$, respectively, then one gets

$$\int_u^\infty f_{X_{\rm sr}}(x)\,{\rm d}x = \frac{1}{2}{\rm erfc}\left(\frac{\ln(u) - (2\mu_{\rm sr} + \Lambda_{\rm sr})}{\sqrt{8}\sigma_{\rm sr}}\right), \quad (17)$$

$$\int_u^\infty f_{X_{\rm sd}}(y)\,{\rm d}y = \frac{1}{2}{\rm erfc}\left(\frac{\ln(u) - (2\mu_{\rm sd} + \Lambda_{\rm sd})}{\sqrt{8}\sigma_{\rm sd}}\right), \quad (18)$$

while μ_i and σ_i, $i \in \{\rm sr, sd\}$, $\Lambda_{\rm sr} = \ln\left(\rho A_{\rm sr}(f, d_{\rm sr})^2\right)$, $\Lambda_{\rm sd} = \ln\left(\rho A_{\rm sd}(f, d_{\rm sd})^2\right)$ and erfc $(\cdot)$ is the complementary error function, which is usually defined as

$${\rm erfc}(x) \triangleq \frac{2}{\sqrt{\pi}}\int_x^\infty \exp\left(-t^2\right){\rm d}t. \quad (19)$$

It is important to point out at this stage that the PLC channel variance depends on the power line network and its value increases as the number of branches, i.e. the number of nodes, between a transmitting and receiving nodes is increased [21], [53]. This could eventually lead to more severe fluctuations in the received signal amplitude. Now, substituting (17) and (18) into (16) yields

$$F_{\mathcal{X}}(u) = 1 - \frac{1}{4}\prod_{i\in\{\rm sr,sd\}} {\rm erfc}\left(\frac{\ln(u) - (2\mu_i + \Lambda_i)}{\sqrt{8}\sigma_i}\right). \quad (20)$$

and then the corresponding PDF can be obtained as

$$f_{\mathcal{X}}(u) = \frac{1}{\sqrt{32\pi}u}\sum_{\substack{i,q\in\{\rm sr,sd\}\\ i\neq q}}\left\{{\rm erfc}\left(\frac{\ln(u) - 2\mu_i - \Lambda_i}{\sqrt{8}\sigma_i}\right)\right. \\ \left.\times\frac{1}{\sigma_q}\exp\left(-\frac{\left(\ln(u) - 2\mu_q - \Lambda_q\right)^2}{8\sigma_q^2}\right)\right\}. \quad (21)$$

Similarly, to find the second capacity, $C_{1,2}$, in (14), let $\mathcal{Y}$ be defined as

$$\mathcal{Y} = \min\left\{\underbrace{a_2\rho A_{\rm sr}(f, d_{\rm sr})^2 h_{\rm sr}^2}_{Y_{\rm sr}}, \underbrace{a_2\rho A_{\rm sd}(f, d_{\rm sd})^2 h_{\rm sd}^2}_{Y_{\rm sd}}\right\}, \quad (22)$$

and following the same procedure used to analyze (15), it is straightforward to show that the PDF of $\mathcal{Y}$ can be expressed as

$$f_{\mathcal{Y}}(u) = \frac{1}{\sqrt{32\pi}u}\sum_{\substack{i,q\in\{\rm sr,sd\}\\ i\neq q}}\left\{{\rm erfc}\left(\frac{\ln(u) - 2\mu_q - \Xi_q}{\sqrt{8}\sigma_q}\right)\right. \\ \left.\times\frac{1}{\sigma_q}\exp\left(-\frac{\left(\ln(u) - 2\mu_q - \Xi_q\right)^2}{8\sigma_q^2}\right)\right\}, \quad (23)$$

where $\Xi_{\rm sr}$ and $\Xi_{\rm sd}$ are given by $\Xi_{\rm sr} = \ln\left(a_2\rho A_{\rm sr}(f, d_{\rm sr})^2\right)$ and $\Xi_{\rm sd} = \ln\left(a_2\rho A_{\rm sd}(f, d_{\rm sd})^2\right)$, respectively.

Let $\mathcal{C}(\cdot)$ be the function that is given by (24), shown at the bottom of this page, and then using (21) and (23), the average capacities of $C_{1,1}$ and $C_{1,2}$ can be calculated as

$$\mathbb{E}\left\{C_{1,1}\right\} = \mathcal{C}(\Lambda), \quad (25a)$$

$$\mathbb{E}\left\{C_{1,2}\right\} = \mathcal{C}(\Xi). \quad (25b)$$

We now derive the average capacity associated with s_2. Using (5) and (8), we can rewrite (13) as

$$C_2 = \frac{1-p}{2}\log_2\left(1 + \min\left\{a_2\rho A_{\rm sr}(f, d_{\rm sr})^2 h_{\rm sr}^2,\right.\right. \\ \left.\left.\rho A_{\rm rd}(f, d_{\rm rd})^2 h_{\rm rd}^2\right\}\right). \quad (26)$$

Next, let us define $\mathcal{Z}$ as

$$\mathcal{Z} = \min\left\{\underbrace{a_2\rho A_{\rm sr}(f, d_{\rm sr})^2 h_{\rm sr}^2}_{Z_{\rm sr}}, \underbrace{\rho A_{\rm rd}(f, d_{\rm rd})^2 h_{\rm rd}^2}_{Z_{\rm rd}}\right\}, \quad (27)$$

and following the same steps used to analyze $\mathcal{X}$ and $\mathcal{Y}$, it is easy to show that the CDF of $\mathcal{Z}$ is given by

$$F_{\mathcal{Z}}(u) = 1 - \frac{1}{4}\prod_{i\in\{\rm sr,rd\}} {\rm erfc}\left(\frac{\ln(u) - 2\mu_i - \Gamma_i}{\sqrt{8}\sigma_i}\right), \quad (28)$$

where Γ_i, $i \in \{\rm sr, rd\}$, is defined as $\Gamma_{\rm sr} = \ln\left(a_2\rho A_{\rm sr}(f, d_{\rm sr})^2\right)$ and $\Gamma_{\rm rd} = \ln\left(\rho A_{\rm rd}(f, d_{\rm rd})^2\right)$. Hence, the corresponding PDF can be expressed as

$$f_{\mathcal{Z}}(u) = \frac{1}{\sqrt{32\pi}u}\sum_{\substack{i,q\in\{\rm sr,sd\}\\ i\neq q}}\left\{{\rm erfc}\left(\frac{\ln(u) - 2\mu_i - \Gamma_i}{\sqrt{8}\sigma_i}\right)\right. \\ \left.\times\frac{1}{\sigma_q}\exp\left(-\frac{\left(\ln(u) - 2\mu_q - \Gamma_q\right)^2}{8\sigma_q^2}\right)\right\}. \quad (29)$$

By making use of the previous result, the average capacity associated with the second signal can be calculated as

$$\mathbb{E}\{C_2\} = \mathcal{C}(\Gamma). \quad (30)$$

Finally, using (25a), (25b) and (30), the average sum capacity of the proposed cooperative relaying NOMA-PLC system can be obtained as

$$\mathbb{E}\left\{C_p\right\} = \mathbb{E}\left\{C_{1,1}\right\} - \mathbb{E}\left\{C_{1,2}\right\} + \mathbb{E}\left\{C_2\right\}. \quad (31)$$

Now, to quantify the achievable rate gains using the proposed NOMA-PLC system, we next study the average capacity performance of the CCR-PLC scheme.

$$\mathcal{C}(\xi) = \frac{1-p}{\sqrt{128\pi}}\int_0^\infty \frac{1}{u}\log_2(1+u)\sum_{\substack{i,q\in\{\rm sr,sd\}\\ i\neq q}}\frac{1}{\sigma_q}\exp\left(-\frac{\left(\ln(u) - 2\mu_q - \xi_q\right)^2}{8\sigma_q^2}\right){\rm erfc}\left(\frac{\ln(u) - 2\mu_i - \xi_i}{\sqrt{8}\sigma_i}\right)du \quad (24)$$

B. AVERAGE SUM CAPACITY OF THE CCR-PLC SYSTEM

In this section, we derive an analytical expression for the average capacity of the CCR-PLC system with DF relaying. In this system, the instantaneous rate is given by [51], [54], [55]

$$C_{\text{con}} = \frac{1-p}{2} \min \times \Big\{ \log_2\left(1 + \rho A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2\right), \log_2\left(1+ \rho A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2 + \rho A_{\text{rd}}(f, d_{\text{rd}})^2 h_{\text{rd}}^2\right)\Big\} \quad (32)$$

which can also more conveniently be rewritten as

$$C_{\text{con}} = \frac{1-p}{2}\log_2 \times \left(1 + \min\Big\{\rho A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2, \rho A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2 + \rho A_{\text{rd}}(f, d_{\text{rd}})^2 h_{\text{rd}}^2\Big\}\right). \quad (33)$$

The first rate in (33) represents the maximum rate at which the source signal is reliably decoded at the relaying modem, whereas the second term indicates the maximum rate at which the destination can reliably decode the combined source and relayed signals.

Letting $\mathcal{W} = \min\{W_{\text{sr}}, W_{\text{sd}} + W_{\text{rd}}\}$ where $W_{\text{sr}} = \rho A_{\text{sr}}(f, d_{\text{sr}})^2 h_{\text{sr}}^2$, $W_{\text{sd}} = \rho A_{\text{sd}}(f, d_{\text{sd}})^2 h_{\text{sd}}^2$ and $W_{\text{rd}} = \rho A_{\text{rd}}(f, d_{\text{rd}})^2 h_{\text{rd}}^2$, the CDF of $\mathcal{W}$ can be expressed as

$$F_{\mathcal{W}}(u) = 1 - \underbrace{\int_u^\infty f_{W_{\text{sr}}}(x)\,\mathrm{d}x}_{\overline{F}_{W_{\text{sr}}}(u)} \underbrace{\int_u^\infty f_{W_{\text{sd}}+W_{\text{rd}}}(y)\,\mathrm{d}y}_{\overline{F}_{W_{\text{sd}}+W_{\text{rd}}}(u)} \quad (34)$$

where $f_{W_{\text{sr}}}(\cdot)$ and $f_{W_{\text{sd}}+W_{\text{rd}}}(\cdot)$ are the PDFs of the RVs W_{sr} and $W_{\text{sd}} + W_{\text{rd}}$, respectively; whereas $\overline{F}_{\text{W}}(\cdot)$ denotes the complementary CDF (CCDF) of W. Since h_{sr}^2 is log-normally distributed, the first CCDF in (34) can be written as

$$\overline{F}_{W_{\text{sr}}}(u) = \frac{1}{2}\text{erfc}\left(\frac{\ln(u) - 2\mu_{\text{sr}} - \Lambda_{\text{sr}}}{\sqrt{8}\sigma_{\text{sr}}}\right), \quad (35)$$

where $\Lambda_{\text{sr}} = \ln\left(\rho A_{\text{sr}}(f, d_{\text{sr}})^2\right)$.

On the other hand, the second CCDF in (34) represents the CCDF of the sum of two log-normal RVs. In this respect, several methods with different degrees of accuracy and complexity have been proposed in the literature to approximate the PDF and CDF of the sum of log-normal RVs. These methods are broadly classified into those that approximate the log-normal sum by a single log-normal RV such as Fenton-Wilkinson and Beaulieu-Xie methods [56]–[58]; and those that compute a compound distribution based on the properties of the log-normal RV such as Farley and Scheher methods [59]–[61]. Amongst those methods, Fenton-Wilkinson's method is the only one that offers a closed-form solution of the underlying parameters of the approximated log-normal RV [61]; therefore, we adopt this method in our analysis below to find the CCDF of the sum log-normal RV $W_{\text{sd}} + W_{\text{rd}}$.

The Fenton-Wilkinson method is based on computing the values of $\mu_{W_{\text{sd+rd}}}$ and $\sigma^2_{W_{\text{sd+rd}}}$ by matching the first and second central moments of $W_{\text{sd+rd}}$ with those of $\sum_{l\in\{\text{sd,rd}\}} W_l$, such that

$$\int_0^\infty u f_{W_{\text{sd+rd}}}(u)\,\mathrm{d}u = \sum_{l\in\{\text{sd,rd}\}} \int_0^\infty u f_{W_l}(u)\,\mathrm{d}u \quad (36)$$

and

$$\int_0^\infty \left(u - \mu_{W_{\text{sd+rd}}}\right)^2 f_{W_{\text{sd+rd}}}(u)\,\mathrm{d}u = \sum_{l\in\{\text{sd,rd}\}} \int_0^\infty \left(u - \mu_{W_l}\right)^2 f_{W_l}(u)\,\mathrm{d}u \quad (37)$$

where $\mu_{W_{\text{sd+rd}}}$ and μ_{W_l} denote the means of the RVs $W_{\text{sd}} + W_{\text{rd}}$ and W_l with $l \in \{\text{sd, rd}\}$, respectively, and

$$f_{W_{\text{sd+rd}}}(u) = \frac{1}{\sqrt{8\pi}\sigma_{W_{\text{sd+rd}}} u}\exp\left(-\frac{\left(\ln(u) - 2\mu_{W_{\text{sd+rd}}}\right)^2}{8\sigma^2_{W_{\text{sd+rd}}}}\right), \quad (38)$$

$$f_{W_l}(z) = \frac{1}{\sqrt{8\pi}\sigma_l z}\exp\left(-\frac{(\ln(z) - 2\mu_l - \Upsilon_l)^2}{8\sigma_l^2}\right), \quad (39)$$

where Υ_l have the values $\Upsilon_{\text{sd}} = \ln\left(\rho A_{\text{sd}}(f, d_{\text{sd}})^2\right)$ and $\Upsilon_{\text{rd}} = \ln\left(\rho A_{\text{rd}}(f, d_{\text{rd}})^2\right)$, respectively. Using (38) and (39), the integrals in (36) can be found to be

$$\int_0^\infty u f_{W_{\text{sd+rd}}}(u)\,\mathrm{d}u = \exp\left(2\mu_{W_{\text{sd+rd}}} + 2\sigma^2_{W_{\text{sd+rd}}}\right) \quad (40)$$

and

$$\sum_{l\in\{\text{sd,rd}\}} \int_0^\infty u f_{Z_l}(u)\,\mathrm{d}u = \sum_{l\in\{\text{sd,rd}\}} \exp\left(2\mu_l + \ln\left(\rho A_l(f, d_l)^2\right) + 2\sigma_l^2\right), \quad (41)$$

where the proof is included in Appendix A. Now, substituting (40) and (41) into (36), and solving for $\mu_{W_{\text{sd+rd}}}$, we can obtain (42), shown at the bottom of the next page. Similarly, the integrals in (37) can be found as

$$\int_0^\infty \left(u - \mu_{W_{\text{sd+rd}}}\right)^2 f_{W_{\text{sd+rd}}}(u)\,\mathrm{d}u = \left(\exp\left(4\sigma^2_{W_{\text{sd+rd}}}\right) - 1\right)\exp\left(4\mu_{W_{\text{sd+rd}}} + 4\sigma^2_{W_{\text{sd+rd}}}\right) \quad (43)$$

and

$$\sum_{l\in\{\text{sd,rd}\}} \int_0^\infty \left(u - \mu_{W_l}\right)^2 f_{W_l}(u)\,\mathrm{d}u = \sum_{l\in\{\text{sd,rd}\}} \exp\left(4\mu_l + 2\ln\left(\rho A_l(f, d_l)^2\right) + 8\sigma_l^2\right) - \sum_{l\in\{\text{sd,rd}\}} \exp\left(4\mu_l + 2\ln\left(\rho A_l(f, d_l)^2\right) + 4\sigma_l^2\right), \quad (44)$$

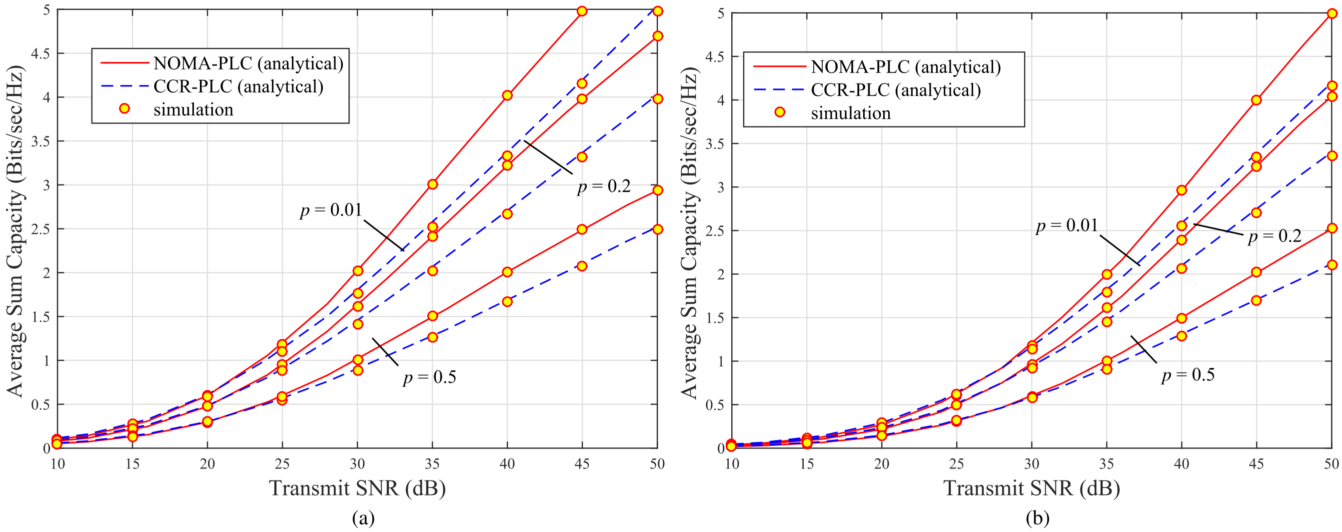

FIGURE 1. Average capacity performance versus transmit SNR for the proposed NOMA-PLC and CCR-PLC systems with different source-to-destination distances. Note that $d_{rd} = 3d_{sr}$, $a_1 = 0.8$ and $a_2 = 0.2$. (a) $d_{sd} = 400$m. (b) $d_{sd} = 600$m.

where the proof is included in Appendix B. Using (36), (37), (40)−(44), and solving for $\sigma^2_{W_{\text{sd+rd}}}$, we obtain (45), shown at the bottom of this page. Next, $\overline{F}_{W_{\text{sd+rd}}}(u)$ can be expressed as

$$\overline{F}_{W_{\text{sd}}+W_{\text{rd}}}(u) = \frac{1}{2}\text{erfc}\left(\frac{\ln(u) - 2\mu_{W_{\text{sd+rd}}}}{\sqrt{8}\sigma_{W_{\text{sd+rd}}}}\right) \tag{46}$$

where $\mu_{W_{\text{sd+rd}}}$ and $\sigma_{W_{\text{sd+rd}}}$ are given by (42) and (45), respectively. Substituting (35) and (46) into (34) yields the CDF of $\mathcal{W}$, which can be expressed as

$$F_{\mathcal{W}}(u) = 1 - \frac{1}{4}\,\text{erfc}\left(\frac{\ln(u) - 2\mu_{W_{\text{sd+rd}}}}{\sqrt{8}\sigma_{W_{\text{sd+rd}}}}\right) \times \text{erfc}\left(\frac{\ln(u) - 2\mu_{\text{sr}} - \Lambda_{\text{sr}}}{\sqrt{8}\sigma_{\text{sr}}}\right). \tag{47}$$

Now, by making use of this result, the average capacity of the CCR-PLC system can finally be calculated using (48), shown at the bottom of this page. It is worthwhile mentioning that the Fenton-Wilkinson method suffers from inaccuracy for large values of σ^2 [57], [62].

Comparing the analyses in Sections IV-A and IV-B, one can easily observe that deriving and computing the average capacity for the CCR-PLC system is far more complex than that of the proposed NOMA-PLC approach. This is basically because of the complexity involved in analyzing the sum of two long-normal RVs which necessitated using the Fenton-Wilkinson method.

V. RESULTS AND DISCUSSIONS

This section presents some numerical examples of the derived expressions above along with Monte Carlo simulations.

$$\mu_{W_{\text{sd+rd}}} = \frac{1}{2}\left(\ln\left[\sum_{l\in\{\text{sd},\text{rd}\}} \exp\left(2\mu_l + \ln\left(\rho A_l(f, d_l)^2\right) + 2\sigma_l^2\right)\right] - 2\sigma^2_{W_{\text{sd+rd}}}\right) \tag{42}$$

$$\sigma^2_{W_{\text{sd+rd}}} = \frac{1}{4}\ln\left[\frac{\sum_{l\in\{\text{sd},\text{rd}\}} \exp\left(4\mu_l + 2\ln\left(\rho A_l(f,d_l)^2\right) + 8\sigma_l^2\right) - \sum_{l\in\{\text{sd},\text{rd}\}} \exp\left(4\mu_l + 2\ln\left(\rho A_l(f,d_l)^2\right) + 4\sigma_l^2\right)}{\sum_{l\in\{\text{sd},\text{rd}\}} \exp\left(4\mu_l + 2\ln\left(\rho A_l(f,d_l)^2\right) + 4\sigma_l^2\right) + 2\exp\left(\sum_{l\in\{\text{sd},\text{rd}\}} 2\mu_l + \ln\left(\rho A_l(f,d_l)^2\right) + 2\sigma_l^2\right)} + 1\right] \tag{45}$$

$$\mathbb{E}\{C_{\text{con}}\} = \frac{1-p}{\sqrt{128\pi}}\int_0^\infty \frac{1}{u}\log_2(1+u)\left[\frac{1}{\sigma_{\text{sr}}}\,\text{erfc}\left(\frac{\ln(u) - 2\mu_{W_{\text{sd+rd}}}}{\sqrt{8}\sigma_{W_{\text{sd+rd}}}}\right)\exp\left(-\frac{(\ln(u) - 2\mu_{\text{sr}} + \Lambda_{\text{sr}})^2}{8\sigma_{\text{sr}}^2}\right) + \frac{1}{\sigma_{W_{\text{sd+rd}}}}\,\text{erfc}\left(\frac{\ln(u) - 2\mu_{\text{sr}} + \Lambda_{\text{sr}}}{\sqrt{8}\sigma_{\text{sr}}}\right)\exp\left(-\frac{\left(\ln(u) - 2\mu_{W_{\text{sd+rd}}}\right)^2}{8\sigma^2_{W_{\text{sd+rd}}}}\right)\right]du \tag{48}$$

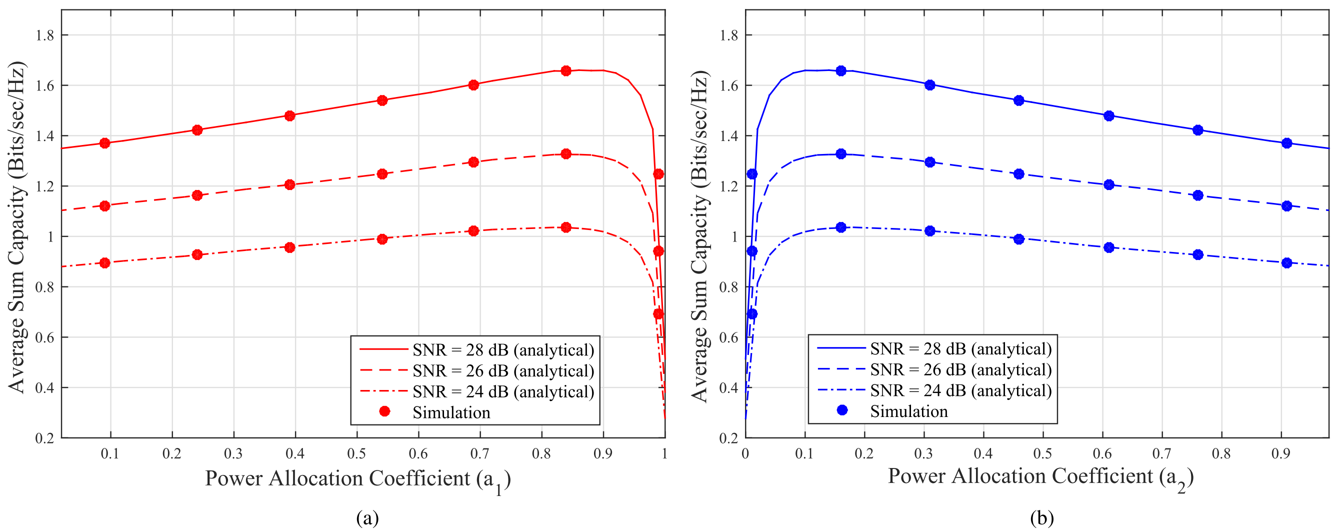

FIGURE 2. **Average capacity performance with respect to the power allocation coefficients for various values of the transmit SNR. Note that $d_{sr} = 100$ m and $d_{rd} = 300$ m. (a) Power allocation coefficient a_1. (b) Power allocation coefficient a_2.**

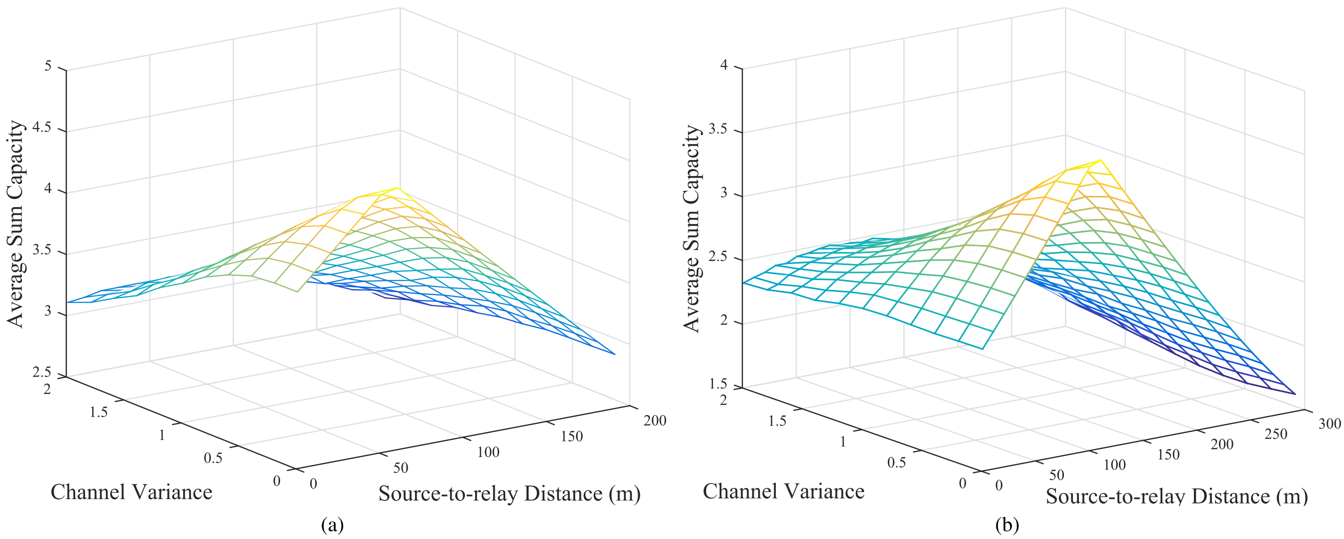

FIGURE 3. **3D surface plots for the average capacity as a function of the PLC channel variance and the source-to-relay distance when SNR = 30dB and $a_1 = 0.8$. (a) $d_{sd} = 200$m. (b) $d_{sd} = 300$m.**

The cable attenuation model used here is given by $A_l(f, d_l) = \exp(-\alpha d_l)$, where $\alpha = b_o + b_1 f^k$ is the attenuation factor, f is the operating frequency in MHz, k is the exponent of the attenuation factor, d_l is the distance, b_o and b_1 are constants determined from measurements and $l \in \{\text{sr}, \text{rd}, \text{sd}\}$. In all our investigations in this section, unless specified otherwise, we used the following parameters: $b_0 = 9.4 \times 10^{-3}$, $b_1 = 4.2 \times 10^{-7}$, $f = 30$ MHz, $k = 0.7$, $p = 0.01$, $a_1 = 0.8$, $\mu_{sr} = \mu_{rd} = \mu_{sd} = 0.4$ and $\sigma_{sr} = \sigma_{rd} = \sigma_{sd} = 0.6$.

To begin with, we plot in Fig. 1 the average capacity of the proposed NOMA-PLC system as a function of the transmit SNR for different values of the impulsive noise probability and source-to-destination distance. Note that the relay here is placed at the midpoint between the end modems. Results for CCR-PLC are also included on this plot for the sake of comparison. The good agreement between the analytical results, obtained from (31) and (48), and the simulated ones clearly indicates the accuracy of our analysis. It is evident that the proposed NOMA-PLC system always outperforms the CCR approach for all the considered system configurations. In particular, for a given capacity requirement and same system parameters, NOMA-PLC has lower transmit power relative to the conventional approach. For example, for $p = 0.01$ and $d_{sd} = 400$ m, NOMA-PLC achieves a capacity of 4 bits/s/Hz with input SNR of 40dB whereas CCR-PLC achieves the same rate with input SNR of 44dB. This can consequently reduce electromagnetic emissions from the power cables, without compromising performance, and hence better coexistence with other wireless systems can be attained. Comparing Figs. 1a and 1b, one can see that increasing the source-to-destination distance, for a given transmit SNR, will always degrade the average capacity for both systems. In addition, it is noticeable that the attained performance gains with the proposed system become more pronounced as the impulsive noise probability becomes lower.

The power allocation coefficients in the proposed NOMA-PLC system are important parameters and to illustrate their impact on the system performance, we show in Figs. 2a and 2b the average capacity as a function of a_1 and a_2, respectively, for several values of the transmit SNR. It can be observed from Fig. 2a that when a_1 is very small, the average capacity is poor irrespective of the SNR value. Similarly, at very large values of a_1, the capacity performance degrades severely. Looking closely at Fig. 2b, the exact opposite trends can be noticed. However, in both figures, it is interesting to notice that for each SNR value there exists an optimal value for the power allocation coefficient that maximizes the average capacity. This implies that optimizing the power allocation coefficient is a key factor to achieve the best performance in NOMA-PLC systems.

Furthermore, Fig. 3 depicts 3D surface plots of the average capacity for the NOMA-PLC scheme as a function of the PLC channel variance and the source-to-relay distance when SNR $=$ 30dB. The total source-to-destination distance is kept fixed at 200m and 300m and the relay is moved from the source towards the destination. It is clear that as the channel variance increases, the performance deteriorates regardless of the relay position and the total end-to-end distance. It is also apparent that the optimal average capacity is influenced by the position of the relay.

VI. CONCLUSION

This paper studied NOMA over PLC channels with DF cooperative relaying. The performance of the proposed system was evaluated in terms of the average capacity and accurate analytical expressions for this performance metric were derived and verified with computer simulations. In addition, we have also assessed the performance of the CCR-PLC approach. Results have clearly demonstrated the superiority of proposed NOMA-PLC system performance over that of the CCR-PLC approach in terms of enhanced average capacity. More specifically, the proposed system was found to achieve same capacity as the CCR-PLC approach with 4dB lower in the transmit power. It was also shown that increasing the number of branches over the PLC link can deteriorate the capacity performance and that good selection of the power allocation coefficient is important to achieve best performance. NOMA-PLC can have the potential to more efficiently reduce the electromagnetic comparability issues associated with the PLC technology compared to OMA-PLC systems. Finally, it is worth mentioning that non-ideal signal cancellation in NOMA-PLC systems will indeed be a subject of future research.

APPENDIX A

This Appendix proves (40) and (41). Using (38), the integral in (40) can be written as

$$\int_0^\infty u f_{W_{\mathrm{sd+rd}}}(u)\,\mathrm{d}u = \frac{1}{\sqrt{8\pi}\sigma_{W_{\mathrm{sd+rd}}}}\int_0^\infty \exp\left(-\frac{\left(\ln(u)-2\mu_{W_{\mathrm{sd+rd}}}\right)^2}{8\sigma^2_{W_{\mathrm{sd+rd}}}}\right)\mathrm{d}u. \tag{49}$$

Next, (37) can be rewritten as

$$\frac{1}{\sqrt{8\pi}\sigma_{W_{\mathrm{sd+rd}}}}\int_0^\infty \exp\left(-\frac{\left(y-2\mu_{W_{\mathrm{sd+rd}}}\right)^2}{8\sigma^2_{W_{\mathrm{sd+rd}}}}+y\right)\mathrm{d}y \tag{50}$$

Let $z = y - 2\mu_{Z_{\mathrm{sd+rd}}}$, then one may rewrite the integral as

$$\begin{aligned}
&= \frac{\exp\left(2\mu_{W_{\mathrm{sd+rd}}}\right)}{\sqrt{8\pi}\sigma_{W_{\mathrm{sd+rd}}}}\int_0^\infty \exp\left(-\frac{z^2}{8\sigma^2_{W_{\mathrm{sd+rd}}}}+z\right)\mathrm{d}z\\
&= \frac{\exp\left(2\mu_{W_{\mathrm{sd+rd}}}\right)}{\sqrt{8\pi}\sigma_{W_{\mathrm{sd+rd}}}}\\
&\quad\times\int_0^\infty \exp\left(\frac{-\left(z-4\sigma^2_{W_{\mathrm{sd+rd}}}\right)^2+16\sigma^4_{W_{\mathrm{sd+rd}}}}{8\sigma^2_{W_{\mathrm{sd+rd}}}}\right)\mathrm{d}z\\
&= \exp\left(2\mu_{W_{\mathrm{sd+rd}}}+2\sigma^2_{W_{\mathrm{sd+rd}}}\right)\\
&\quad\times\underbrace{\int_0^\infty \frac{1}{\sqrt{8\pi}\sigma_{W_{\mathrm{sd+rd}}}}\exp\left(-\frac{\left(z-4\sigma^2_{W_{\mathrm{sd+rd}}}\right)^2}{8\sigma^2_{W_{\mathrm{sd+rd}}}}\right)\mathrm{d}z}_{=1}\\
&= \exp\left(2\mu_{W_{\mathrm{sd+rd}}}+2\sigma^2_{W_{\mathrm{sd+rd}}}\right).
\end{aligned} \tag{51}$$

This proves (40). The proof for (41) is omitted for the sake of brevity since it can be easily proven following the same procedure as above.

APPENDIX B

In this Appendix, we prove (43) and (44). To begin with, the integral in (43) basically represents the second moment and can be rewritten as

$$\int_0^\infty \left(u-\mu_{W_{\mathrm{sd+rd}}}\right)^2 f_{W_{\mathrm{sd+rd}}}(u)\,\mathrm{d}u = \mathbb{E}\left\{U^2\right\} - \left(\mathbb{E}\left\{U\right\}\right)^2 \tag{52}$$

$\left(\mathbb{E}\{U\}\right)^2$ can be obtained by making use of (51) and $\mathbb{E}\left\{U^2\right\}$ is calculated as $\mathbb{E}\left\{U^2\right\} = \int_0^\infty u^2 f_{W_{\mathrm{sd+rd}}}(u)\,\mathrm{d}u$, which yields

$$\mathbb{E}\left\{U^2\right\} = \frac{1}{\sqrt{8\pi}\sigma_{W_{\mathrm{sd+rd}}}}\int_0^\infty u\exp\left(-\frac{\left(\ln(u)-2\mu_{W_{\mathrm{sd+rd}}}\right)^2}{8\sigma^2_{W_{\mathrm{sd+rd}}}}\right)\mathrm{d}u. \tag{53}$$

Using the substitution $y = \ln(u)$ and $\mathrm{d}u = e^y\mathrm{d}y$, we can rewrite (37) as

$$= \frac{1}{\sqrt{8\pi}\sigma_{W_{\mathrm{sd+rd}}}}\int_0^\infty \exp\left(-\frac{\left(y-2\mu_{W_{\mathrm{sd+rd}}}\right)^2}{8\sigma^2_{W_{\mathrm{sd+rd}}}}+2y\right)\mathrm{d}y \tag{54}$$

Now, using $z = y - 2\mu_{W_{\text{sd+rd}}}$, $2y = 2z + 4\mu_{W_{\text{sd+rd}}}$ and $dz = dy$, we obtain

$$\begin{aligned}
&= \frac{\exp\left(4\mu_{W_{\text{sd+rd}}}\right)}{\sqrt{8\pi}\sigma_{W_{\text{sd+rd}}}} \int_0^\infty \exp\left(-\frac{z^2}{8\sigma^2_{W_{\text{sd+rd}}}} + 2z\right) dz \\
&= \frac{\exp\left(4\mu_{W_{\text{sd+rd}}}\right)}{\sqrt{8\pi}\sigma_{W_{\text{sd+rd}}}} \int_0^\infty \exp\left(\frac{-z^2 + 16z\sigma^2_{W_{\text{sd+rd}}}}{8\sigma^2_{W_{\text{sd+rd}}}}\right) dz \\
&= \frac{\exp\left(4\mu_{W_{\text{sd+rd}}}\right)}{\sqrt{8\pi}\sigma_{W_{\text{sd+rd}}}} \\
&\quad \times \int_0^\infty \exp\left(\frac{-\left(z - 8\sigma^2_{W_{\text{sd+rd}}}\right)^2 + 64\sigma^4_{W_{\text{sd+rd}}}}{8\sigma^2_{W_{\text{sd+rd}}}}\right) dz \\
&= \exp\left(4\mu_{W_{\text{sd+rd}}} + 8\sigma^2_{W_{\text{sd+rd}}}\right) \\
&\quad \times \underbrace{\int_0^\infty \frac{1}{\sqrt{8\pi}\sigma_{W_{\text{sd+rd}}}} \exp\left(-\frac{\left(z - 8\sigma^2_{W_{\text{sd+rd}}}\right)^2}{8\sigma^2_{W_{\text{sd+rd}}}}\right) dz}_{=1} \\
&= \exp\left(4\mu_{W_{\text{sd+rd}}} + 8\sigma^2_{W_{\text{sd+rd}}}\right). \qquad (55)
\end{aligned}$$

Now, substituting (51) and (55) into (52) with some basic algebraic manipulations, we obtain (43). Similarly, we can obtain (44).

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