

A Novel OCDM-Based Sounding Sequence for a PLC Channel Measurement Methodology

Victor H. Cecatto, Evelio M. G. Fernández, Moisés V. Ribeiro, and Ândrei Camponogara

Abstract—This paper proposes a novel sounding sequence based on orthogonal chirp-division multiplexing (OCDM) for application in a power line communication (PLC) channel measurement methodology. The proposal integrates the Schmidl & Cox synchronization technique and utilizes a structure of consecutive orthogonal symbol blocks, with orthogonality achieved via circular shifting in the discrete-Fresnel domain. To evaluate its effectiveness, numerical results are presented in terms of the mean, standard deviation, and error rate of the estimated synchronization points for different numbers of symbol blocks, using a real dataset comprising PLC channel estimates collected in a measurement campaign. Compared to an existing OCDM-based sounding sequence, the results show that the proposed sounding sequence achieves a better symbol block synchronization with a lower error rate.

Keywords—Power line communication; measurement methodology; orthogonal chirp-division multiplexing; sounding; synchronization.

I. INTRODUCTION

The growing demand for reliable and cost-effective communication infrastructures has driven the adoption of power line communication (PLC) systems in applications such as the Internet of Things (IoT), smart metering, smart grids, and Industry 4.0 [1], [2]. By leveraging existing electrical networks, PLC is an attractive alternative to conventional communication technologies. However, the hostile characteristics of power line channels, including impedance mismatching and impulsive noise impose significant challenges to reliable data transmission [3], [4]. Therefore, accurate and robust channel estimation methods are essential for PLC systems.

The measurement methodology presented in [5] proposes a sounding signal based approach employing the discrete multitone (DMT) modulation and cyclic prefix (CP) correlation for symbol-block synchronization. Subsequently, [6] improved this methodology by adopting the Schmidl-Cox symbol block synchronization technique as described in [7].

More recently, orthogonal chirp-division multiplexing (OCDM) has been employed in communication systems due to its low time-frequency distortion and its ability to maintain

orthogonality in environments with severe channel effects. With the advancement of OCDM, [8] proposed an OCDM-based pilot symbol that adopts the synchronization technique introduced in [7]. Extending this further, [9] proposed a sounding sequence composed of multiple OCDM blocks, also combined with the method in [7]. This evolution enables more accurate results and a more reliable representation of PLC channel behavior, particularly in higher frequency bands or in scenarios with greater temporal variability.

In this context, this paper investigates the application of OCDM within the PLC measurement methodology presented in [5], with a specific focus on symbol block synchronization performance. To this end, a novel OCDM-based sounding sequence is proposed, employing a circular shift in the discrete-Fresnel domain to generate orthogonal symbol blocks. The performance of this proposal is compared against the OCDM-based sounding sequence introduced in [9] through numerical simulations, analyzing the mean, standard deviation, and error rate of the estimated synchronization points. The obtained results show that the proposal provides a more reliable symbol block synchronization, with lower standard deviation and, consequently, lower symbol block error rate.

The remainder of this paper is organized as follows. Section II describes the PLC channel measurement methodology. Section III introduces the OCDM-based sounding sequences, while Section IV details the Schmidl & Cox synchronization technique. Finally, Section V presents the numerical results, and Section VI provides the concluding remarks.

II. MEASUREMENT METHODOLOGY

Fig. 1 depicts the block diagram of the measurement setup employed to estimate PLC channels. The measurement methodology encompasses four stages [5]: (i) transmission and reception of a sounding sequence, (ii) synchronization, (iii) channel estimation, and (iv) channel enhancement. The sounding sequence is essential for these operations and is given by

$$s[n] = \sum_{i=0}^{K-1} \sum_{m=0}^{N_f-1} s_f[m] \delta[n - m - iN_f], \quad (1)$$

where K is the number of transmission frames of length N_f and $\delta[\cdot]$ denotes the Kronecker delta. Moreover, $s_f[n]$ corresponds to the n^{th} element of the vector representation of the discrete-time frame sequence $\mathbf{s}_f = [s_f[0] \ s_f[1] \ \cdots \ s_f[N_f - 1]]^T$. This frame is constructed by concatenating two orthogonal symbol blocks represented by the vectors $\mathbf{x}_1, \mathbf{x}_2 \in \mathbb{R}^{N \times 1}$ with their respective cyclic

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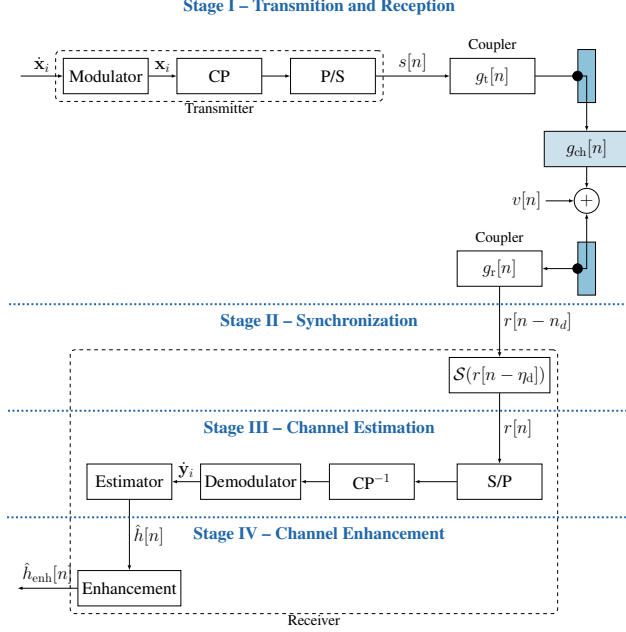


Fig. 1. Block diagram of the discrete-time equivalent representation of the PLC channel measurement methodology.

prefixes (CPs) of length N_{CP} appended; consequently, $N_f = 2(N + N_{CP})$.

The sounding sequence is converted to an analog signal and is then injected into the electric power grid by a coupler circuit. At the receiver, the signal is collected through another coupler circuit and digitized. As the transmitter and receiver have different local oscillators, a sampling frequency offset (SFO) is expected. Since this paper focuses on symbol block synchronization, a negligible SFO is assumed.

The PLC channel is assumed to be linear and time-invariant (LTI) during a symbol block duration. Accordingly, its discrete-time channel impulse response (CIR) is denoted by $\{g_{ch}[n]\}_{n=0}^{L_{ch}-1}$, where L_{ch} represents the channel length. The received sequence is given by

$$r[n - \eta_d] = \sum_{m=0}^{KN_f-1} s[m]h[n - \eta_d - m] + v[n] \quad (2)$$

in which $\{v[n]\}$ denotes a zero-mean additive noise with $\mathbb{E}\{v[n]v[m]\} = 0, \forall n \neq m$ and $\mathbb{E}\{v^2[n]\} = \sigma_v^2$. $\{h[n]\}_{n=0}^{L_h-1}$ represents the equivalent PLC CIR of length L_h such that $h[n] = g_t[n] \star g_{ch}[n] \star g_r[n]$, with $\star$ being the convolution operator. Here, $\{g_t[n]\}_{n=0}^{L_t-1}$ and $\{g_r[n]\}_{n=0}^{L_r-1}$ are the respective impulse responses of the transmit and receive coupler circuits. Consequently, $L_h = L_t + L_{ch} + L_r - 2$. Finally, η_d represents the time delay between the beginning of the sounding sequence and the receiver's local time reference.

Following signal acquisition at the receiver, stage (ii) performs symbol block synchronization to compensate for the delay η_d . Then, $r[n] = \mathcal{S}(r[n - \eta_d])$, where $\mathcal{S}(\cdot)$ is an operator used to estimate and correct η_d . In this paper, the symbol block synchronization technique proposed by Schmidl & Cox [7] is adopted to obtain $\{r[n]\}$.

After serial-to-parallel conversion, CP removal and the

demodulation process, the vector representation of the i^{th} received symbol block in the discrete-Fresnel domain, $\hat{\mathbf{y}}_i \in \mathbb{C}^{N \times 1}$, is obtained. Then, channel estimation (iii) is performed. Note that this process strongly depends on the digital modulation adopted and sounding-sequence design [5], [9]. Finally, channel enhancement (iv) is carried out to mitigate the effect of additive noise [5]. Considering a CIR estimate $\{\hat{h}[n]\}_{n=0}^{N-1}$, stage (iv) consists of truncating this sequence to $\hat{L}_h$ samples, which correspond to 99% of its energy, and then zero-padding it with $N - \hat{L}_h$ zeros, resulting in

$$\hat{h}_{enh}[n] = \begin{cases} \hat{h}[n], & n = 0, 1, \dots, \hat{L}_h - 1 \\ 0, & n = \hat{L}_h, \hat{L}_h + 1, \dots, N - 1 \end{cases} \quad (3)$$

III. OCDM-BASED SOUNDING SEQUENCE

Let $\hat{\mathbf{x}}_i = [\hat{x}_i[0] \ \hat{x}_i[1] \ \dots \ \hat{x}_i[N-1]]^T$, with $i \in \{1, 2\}$, denote the vector representation of the i^{th} symbol block in the discrete-Fresnel domain. The discrete-time version of the i^{th} symbol block is given by $\mathbf{x}_i = \Phi^\dagger \hat{\mathbf{x}}_i$, where $\Phi = \mathbf{F}^\dagger \mathbf{\Gamma} \mathbf{F}$ is the discrete Fresnel Transform (DFnT) matrix, $\mathbf{F}$ is the normalized N -size discrete Fourier transform (DFT) matrix, and $\mathbf{\Gamma}$ is a diagonal matrix containing the eigenvalues of Φ . The structure of $\mathbf{\Gamma}$ depends on whether the system operates in passband or baseband. Assuming $N \equiv 0 \pmod{2^p}$ and $p > 2$, in the former case, the $(k, k)^{\text{th}}$ element of $\mathbf{\Gamma}$ is given by [10]

$$\Gamma[k, k] \triangleq e^{-j \frac{\pi}{N} k^2}, \quad k = 0, \dots, N-1, \quad (4)$$

whereas, in the latter case, the $(k, k)^{\text{th}}$ element can be expressed as [9], [11]

$$\Gamma[k, k] \triangleq \begin{cases} e^{-j \frac{2\pi}{N} k^2}, & k = 0, \dots, \frac{N}{2} - 1, \\ e^{j \frac{2\pi}{N} k^2}, & k = \frac{N}{2}, \dots, N-1. \end{cases} \quad (5)$$

The proposed sounding sequence is designed to comply with the symbol block synchronization technique proposed by Schmidl & Cox, which requires symbol blocks to exhibit identical halves in the discrete-time domain [7]. To generate an OCDM-based symbol block with this property, as in [9], we adopt the approach presented in [8], in which the first symbol block in the discrete-Fresnel domain is expressed as

$$\hat{x}_1[k] = \delta[k] + \delta[k - N/2], \quad k = 0, 1, \dots, N-1. \quad (6)$$

As aforementioned, effective symbol block synchronization in the measurement methodology requires the symbol blocks represented by the vectors $\mathbf{x}_1, \mathbf{x}_2 \in \mathbb{R}^{N \times 1}$ to be orthogonal [5]. In this regard, Pimenta *et al.* [9] generated the second symbol block by shifting the first symbol block by one sample in the discrete-frequency domain, as follows:

$$\hat{x}_2[k] = \hat{x}_1[k] \cdot e^{j \frac{2\pi}{N} k} = \delta[k] - \delta\left[k - \frac{N}{2}\right]. \quad (7)$$

In this paper, an alternative approach is proposed to generate the second symbol block. It consists of applying a circular shift of ℓ samples to the first symbol block in the discrete-Fresnel domain, i.e.,

$$\hat{x}_2[k] = \hat{x}_1[\langle k - \ell \rangle_N], \quad k = 0, 1, \dots, N-1, \quad (8)$$

where $\langle a \rangle_N \triangleq a \bmod N$, and $\ell \in \mathbb{Z} \mid 0 < \ell < N/2$.

A. Orthogonality Analysis

To verify the orthogonality between the proposed symbol blocks in the discrete-time domain, we use their complex baseband representations for simplicity. In this regard, the discrete-time version of the first symbol block is given by [10]

$$\begin{aligned}
 x_1[n] &= e^{-j\frac{\pi}{4}} \sum_{k=0}^{N-1} \dot{x}_1[k] e^{j\frac{\pi}{N}(n-k)^2} \\
 &= e^{-j\frac{\pi}{4}} \left[e^{j\frac{\pi}{N}n^2} + e^{j\frac{\pi}{N}(n-N/2)^2} \right] \\
 &= e^{-j\frac{\pi}{4}} e^{j\frac{\pi}{N}n^2} \left[1 + e^{-j\frac{\pi}{N}nN} e^{j\frac{\pi}{N}\frac{N^2}{4}} \right] \\
 &\stackrel{(a)}{=} e^{-j\frac{\pi}{4}} e^{j\frac{\pi}{N}n^2} \left[1 + (-1)^n e^{j\frac{\pi}{4}N} \right] \\
 &= e^{-j\frac{\pi}{4}} e^{j\frac{\pi}{N}n^2} \left[1 + (-1)^n \right], \tag{9}
 \end{aligned}$$

where in (a) the term $e^{j\frac{\pi}{4}N} = 1$ for the considered values of N . The discrete-time domain version of the second symbol block is expressed as

$$\begin{aligned}
 x_2[n] &= e^{-j\frac{\pi}{4}} \sum_{k=0}^{N-1} \dot{x}_1[(k-\ell)_N] e^{j\frac{\pi}{N}(n-k)^2} \\
 &= e^{-j\frac{\pi}{4}} \left[e^{j\frac{\pi}{N}(n-\ell)^2} + e^{j\frac{\pi}{N}(n-N/2-\ell)^2} \right] \\
 &= e^{-j\frac{\pi}{4}} e^{j\frac{\pi}{N}(n-\ell)^2} \left[1 + e^{-j\frac{\pi}{N}(n-\ell)N} e^{j\frac{\pi}{N}\frac{N^2}{4}} \right] \\
 &= e^{-j\frac{\pi}{4}} e^{j\frac{\pi}{N}(n-\ell)^2} \left[1 + (-1)^{(n-\ell)} \right]. \tag{10}
 \end{aligned}$$

Notice that a circular shift of N samples by ℓ in the discrete-Fresnel domain results in a circular shift by ℓ in the discrete-time domain, i.e., $x_2[n] = x_1[(n-\ell)_N]$.

To demonstrate the orthogonality conditions of the proposed symbol blocks, we compute their cyclic cross-correlation $\rho[n]$ as follows:

$$\begin{aligned}
 \rho[n] &= \sum_{m=0}^{N-1} x_1[m] x_2^*[(n+m)_N] \\
 &= \sum_{m=0}^{N-1} e^{j\frac{\pi}{N}m^2} e^{-j\frac{\pi}{N}(n+m-\ell)^2} \left[1 + (-1)^m \right] \times \\
 &\quad \left[1 + (-1)^{n+m-\ell} \right] \\
 &= e^{-j\frac{\pi}{N}(n-\ell)^2} \sum_{m=0}^{N-1} e^{-j\frac{2\pi}{N}m(n-\ell)} \times \\
 &\quad \left[1 + (-1)^{n-\ell}(-1)^m + (-1)^m + (-1)^{n-\ell} \right] \\
 &= e^{-j\frac{\pi}{N}(n-\ell)^2} \left[1 + (-1)^{n-\ell} \right] \times \\
 &\quad \left[\sum_{m=0}^{N-1} e^{-j\frac{2\pi}{N}m(n-\ell)} + \sum_{m=0}^{N-1} (-1)^m e^{-j\frac{2\pi}{N}m(n-\ell)} \right] \\
 &= e^{-j\frac{\pi}{N}(n-\ell)^2} \left[1 + (-1)^{n-\ell} \right] \times \\
 &\quad \left[\sum_{m=0}^{N-1} \left(e^{-j\frac{2\pi}{N}(n-\ell)} \right)^m + \sum_{m=0}^{N-1} \left(-e^{-j\frac{2\pi}{N}(n-\ell)} \right)^m \right]. \tag{11}
 \end{aligned}$$

Since $N \equiv 0 \pmod{2^p}$, the geometric series above can be simplified. Specifically, the first summation satisfies the

standard DFT orthogonality property:

$$\sum_{m=0}^{N-1} \left(e^{-j\frac{2\pi}{N}(n-\ell)} \right)^m = \begin{cases} N, & n - \ell \equiv 0 \pmod{N} \\ 0, & \text{otherwise} \end{cases} \tag{12}$$

while the second summation, due to the $(-1)^m$ term, results in:

$$\begin{aligned}
 \sum_{m=0}^{N-1} \left(-e^{-j\frac{2\pi}{N}(n-\ell)} \right)^m &= \frac{1 - (-1)^N e^{-j\frac{2\pi}{N}(n-\ell)N}}{1 + e^{-j\frac{2\pi}{N}(n-\ell)}} \\
 &= \begin{cases} N, & n - \ell \equiv N/2 \pmod{N} \\ 0, & \text{otherwise.} \end{cases} \tag{13}
 \end{aligned}$$

Consequently, the cross-correlation function simplifies to

$$\rho[n] = \begin{cases} 2N, & n = \ell \\ 2N, & n = \ell \pm N/2 \\ 0, & \text{otherwise.} \end{cases} \tag{14}$$

The result indicates that the symbol blocks are not orthogonal for all n , which must be accounted for in the sounding sequence design. It is important to note that this property was derived in the complex baseband to simplify the mathematical manipulation; however, the same correlation characteristics are preserved in the equivalent baseband representation.

IV. SCHMIDL & COX TECHNIQUE

To estimate the beginning of each symbol block in the sounding sequence, the Schmidl & Cox scheme first computes the normalized correlation between two adjacent windows of length $L = N/2$ that slide along the received sequence $\{r[n]\}$. The normalized correlation is given by

$$M[n] = \frac{|\mathbf{r}_n^\dagger \mathbf{r}_{n+L}|^2}{(\mathbf{r}_{n+L}^\dagger \mathbf{r}_{n+L})^2}, \tag{15}$$

where $\mathbf{r}_n = [r[n] \ r[n+1] \ \cdots \ r[n+L-1]]^T$ and $\mathbf{r}_{n+L} = [r[n+L] \ r[n+L+1] \ \cdots \ r[n+2L-1]]^T$. They represent the windows sliding along the received sounding sequence.

Finally, the beginning of the transmitted sounding sequence, $\hat{\eta}$, is estimated by averaging the midpoints of the plateaus, $\{\hat{\eta}[q]\}_{q=0}^{N_{\text{plateau}}-1}$, found in $\{M[n]\}$ [6]. Algorithm 1 details the procedure used to estimate $\hat{\eta}$. In the algorithm, $\lfloor x \rfloor = \max \{n \in \mathbb{Z} \mid n \leq x\}$. Observe that Algorithm 1 holds only if the periodicity of $\{M[n]\}$ is guaranteed. This sequence has a fundamental period equal to $N_p = N_{\text{CP}} + N$, i.e., $M[n] = M[n+mN_p], \forall m \in \mathbb{Z}$. This periodicity is maintained only if $\ell \neq N_{\text{CP}}$. This is a straightforward consequence of the correlation property of the proposed OCDM-based sounding sequence demonstrated in (14).

To illustrate the influence of the choice of ℓ on the periodicity of $\{M[n]\}$, Fig. 2 shows $\{M[n]\}$ for a sounding sequence with $N = 1024$, $N_{\text{CP}} = 128$, and $N_p = 1152$, comparing cases where $\ell \neq N_{\text{CP}}$ and $\ell = N_{\text{CP}}$. Note that the symbol blocks of the sounding sequence are designed with identical

Algorithm 1 Symbol block synchronization

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1: Input: Normalized correlation sequence  $\{M[n]\}$  of length
    $L_M$ , symbol block length  $N$ , cyclic prefix length  $N_{CP}$ ,
   and threshold  $\gamma$ .
2: Output: Estimated synchronization point  $\hat{\eta}$ .
3:  $N_p = N + N_{CP}$ 
4:  $n_\gamma = \min\{n \mid M[n] > \gamma\}$ 
5:  $n_{\text{first}} = n_\gamma + \frac{N}{2} + N_{CP}$ 
6:  $N_{\text{win}} = \lfloor (L_M - n_{\text{first}})/N_p \rfloor$ 
7: for  $q = 0$  to  $N_{\text{win}} - 1$  do
8:    $\mathbf{m} = [M[n_{\text{first}} + qN_p], \dots, M[n_{\text{first}} + (q+1)N_p - 1]]^T$ 
9:    $\mathcal{P} = \{p \in \{0, \dots, N_p - 1\} \mid \mathbf{m}[p] > \gamma\}$ 
10:  if  $\mathcal{P} \neq \emptyset$  then
11:     $\hat{\eta}[q] = \frac{\min(\mathcal{P}) + \max(\mathcal{P})}{2} + n_{\text{first}}$ 
12:  end if
13: end for
14:  $N_{\text{plateau}} = \text{length}(\hat{\eta})$ 
15:  $\hat{\eta} = \frac{1}{N_{\text{plateau}}} \sum_{q=0}^{N_{\text{plateau}}-1} [\hat{\eta}[q] \pmod{N_p}]$ 

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halves, i.e., $\mathbf{x}_i = [\mathbf{x}'_i \ \mathbf{x}'_i^T]^T$. As shown in Fig. 2, a properly designed sounding sequence yields

$$M[n] = 0, \quad n = N/2 + N_{CP} + mN_p, \quad \forall m \in \mathbb{Z}. \quad (16)$$

In this case, we have $\mathbf{r}_n = \mathbf{x}'_1$ and $\mathbf{r}_{n+L} = [x'_2[\langle -N_{CP} \rangle_L] \ x'_2[\langle 1 - N_{CP} \rangle_L] \ \cdots \ x'_2[\langle L - 1 - N_{CP} \rangle_L]]^T$. Since $x'_2[n] = x'_1[\langle n - \ell \rangle_L]$, with $\ell \neq N_{CP}$, it follows that $\mathbf{r}_n \neq \mathbf{r}_{n+L}$ and, consequently, $M[n] = 0$. However, if $\ell = N_{CP}$, then

$$M[n] = \begin{cases} 0, & n = L + N_{CP} + 2mN_p \\ 1, & n = L + N_{CP} + (2m+1)N_p \end{cases}, \quad \forall m \in \mathbb{Z}. \quad (17)$$

Specifically, when $n = L + N_{CP} + (2m+1)N_p$, we have $\mathbf{r}_n = \mathbf{x}'_2$ and $\mathbf{r}_{n+L} = [x'_1[\langle -N_{CP} \rangle_L] \ x'_1[\langle 1 - N_{CP} \rangle_L] \ \cdots \ x'_1[\langle L - 1 - N_{CP} \rangle_L]]^T$. Under the condition $\ell = N_{CP}$, it holds that $x'_2[n] = x'_1[\langle n - N_{CP} \rangle_L]$, which leads to $\mathbf{r}_n = \mathbf{r}_{n+L}$ and, therefore, $M[n] = 1$.

V. NUMERICAL RESULTS

This section presents a numerical analysis of the proposed OCDM-based sounding sequence in terms of the mean, standard deviation, and error rate of the estimated synchronization points. Additionally, a comparison with the OCDM-based sounding sequence introduced in [9] is provided. For this analysis, one assumes $N = 2048$, $N_{CP} = 512$, and a circular shift of $\ell = N/8$. Furthermore, to estimate the synchronization points, a threshold $\gamma = 0.8$ is applied.

The numerical simulations are conducted using a real dataset composed of broadband in-home PLC channel frequency response (CFR) estimates collected during the measurement campaign presented in [12]. Specifically, based on (3), only CIRs of length $\hat{L}_h = 256$ are selected to satisfy the condition $\hat{L}_h \leq N_{CP}/2$. Consequently, 53 CFR estimates covering the frequency band from 1.7 to 100 MHz are considered.

Figs. 3, 4, and 5 compare the proposed sounding sequence with the one presented in [9] in terms of the mean, standard

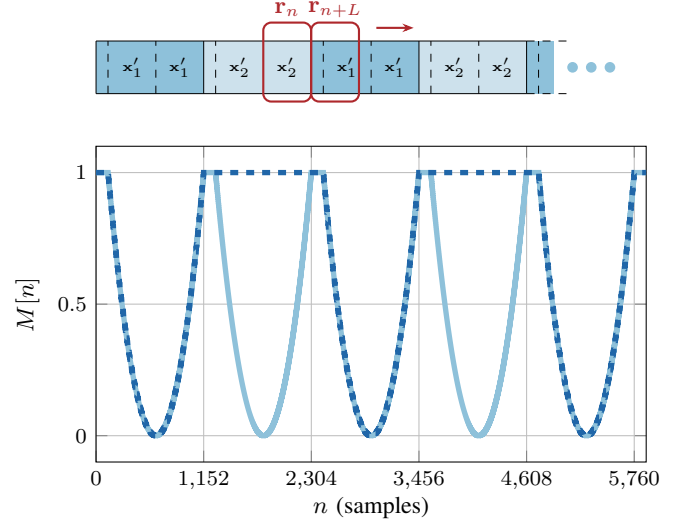


Fig. 2. Normalized correlation, $\{M[n]\}$, computed over the proposed OCDM-based sounding sequence design with two different values of circular shift: $\ell \neq N_{CP}$ (solid line) and $\ell = N_{CP}$ (dashed line).

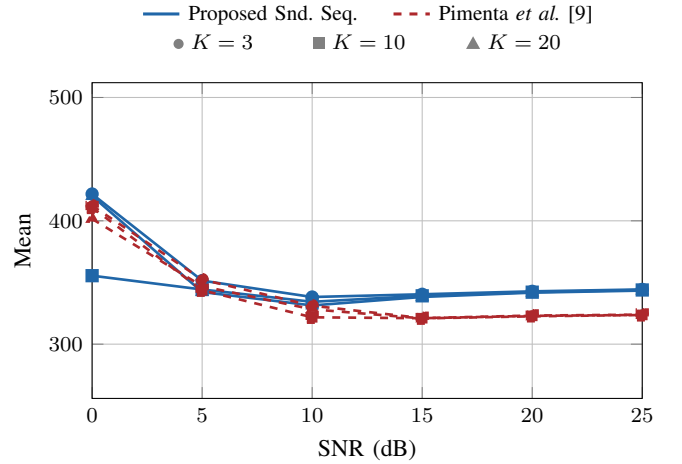


Fig. 3. Mean of the estimated synchronization points, $\hat{\eta}$, as functions of SNR for different values of K .

deviation, and error rate of the estimated synchronization points $\hat{\eta}$, respectively, as functions of signal-to-noise ratio (SNR), for $K \in \{3, 10, 20\}$. The results indicate that at higher SNR regimes, $\hat{\eta}$ tends to stabilize near the midpoint between intersymbol interference (ISI)-free region, $[\hat{L}_h, N_{CP}]$. This reduces the standard deviation and, consequently, drives the error rate toward zero. In contrast, at lower SNRs, the additive noise significantly degrades the estimates computed from both sounding sequences, resulting in a higher standard deviation, even though the mean remains within the ISI-free region. Moreover, one sees that the number of consecutive symbol blocks K reduces the additive noise degradation on the estimates. Finally, the proposed sounding sequence provides an average $\hat{\eta}$ closer to the ISI-free region midpoint along with a lower standard deviation, which indicates a more stable synchronization point. As a result, the proposed sounding sequence exhibits a significantly lower error rate, converging

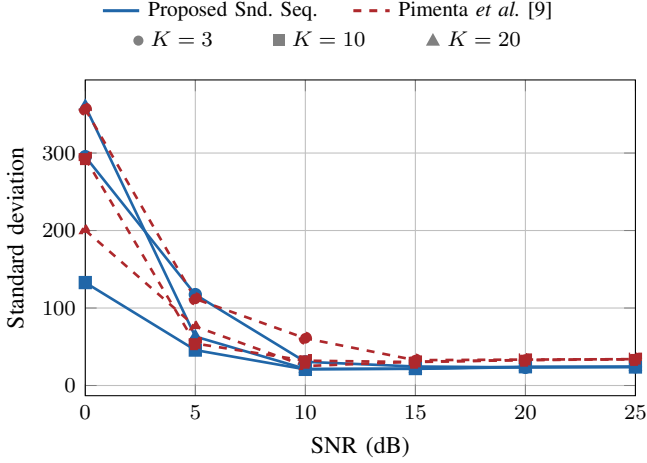


Fig. 4. Standard deviation of the estimated synchronization points, $\hat{\eta}$, as functions of SNR for different values of K .

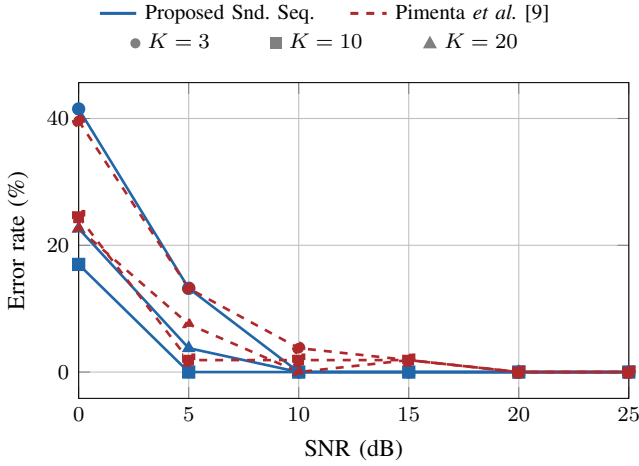


Fig. 5. Error rate of the estimated synchronization points, $\hat{\eta}$, as functions of SNR for different values of K .

to 0 at approximately 10 dB, whereas the sounding sequence proposed in [9] requires nearly 20 dB to achieve similar convergence.

VI. CONCLUSIONS

This paper investigates the suitability of OCDM to a PLC channel measurement methodology and proposes a novel OCDM-based sounding sequence, which employs a circular shift in the Fresnel domain to generate orthogonal symbol blocks. The proposal is evaluated through numerical simulations in terms of the mean, standard deviation, and error rate of the estimated synchronization points. Furthermore, it is compared with an existing OCDM-based sounding sequence from the literature, which relies on a different symbol block orthogonalization approach.

Numerical results show that the proposed OCDM-based sounding sequence provides a more effective orthogonalization between consecutive symbol blocks, leading to improved

accuracy in the symbol block synchronization. Compared to the literature baseline, the proposal achieves lower error rates across a wide range of SNR regimes, consistently maintaining synchronization points within the ISI-free region. These findings highlight the OCDM scheme as a promising and robust solution for PLC channel measurement applications.

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