

A Methodology for Estimating Passband PLC Channel Frequency Responses

Thiago F. do A. Nogueira, Mateus de L. Filomeno, Luís G. da S. Costa, and Moisés V. Ribeiro

Abstract—This paper presents an orthogonal frequency-division multiplexing (OFDM)-based methodology for estimating the passband frequency responses of power line communication (PLC) channels. The proposal combines the estimation and correction of symbol-timing, carrier-frequency, and sampling-frequency offsets with channel estimation and enhancement techniques. Its effectiveness is validated through an experimental setup in which software-defined radio (SDR) devices transmit OFDM signals over a non-energized power line cable, and the symbols received by another SDR are processed using the proposed methodology. A comparison with vector network analyzer measurements obtained from the same power line cable shows close agreement, indicating the correctness of the chosen techniques.

Keywords—power line communication, channel frequency response, passband, estimation, synchronization.

I. INTRODUCTION

Power line communication (PLC) systems over low-voltage electric power grids have been widely investigated to expand indoor data communications [1]–[3]. Given the current state of PLC systems, substantial advances require operation beyond the bandwidth allowed by existing regulations, which limit the maximum frequency to 100 MHz [4]. Extending PLC operation up to 500 MHz can increase indoor throughput beyond 4 Gbps, since spectral efficiencies of 8 bits/s/Hz or higher are readily achievable.

However, operation at higher frequencies (e.g., up to 500 MHz) requires a proper understanding of indoor electric power grids as communication media. Such characterization relies on the availability of channel state information (CSI) and the statistics of additive noise. Regarding CSI, the channel frequency response (CFR) is widely used to characterize PLC channels because it describes, in the frequency domain, how the medium affects the spectral content of the transmitted signal. Moreover, the knowledge of CFR supports the design of countermeasures against channel impairments and enables the optimization of channel resources when additive-noise statistics are available.

Under a top-down approach, CFRs of PLC channels can be estimated using two main methods: (i) frequency sweep,

in which a vector network analyzer (VNA) changes the test-signal frequency step by step over a specified range to obtain the CFR; and (ii) sounding, in which the transmitted signal is designed so that CFR estimates can be obtained from the received signal. Owing to its simplicity, the former has been widely used when the injection and extraction points are close, and the electric circuits are time-invariant. However, a VNA may yield inaccurate CFR estimates because its sweep time is typically greater than 100 ms, whereas the coherence time T_c of electric power grids, *i.e.*, the interval during which the channel impulse response remains time invariant, is lower than 2 ms [5]. In contrast, sounding methods are more suitable for estimating PLC CFRs, since the sounding signal can be designed to provide one CFR estimate within a time interval shorter than the coherence time; see [6] and references therein, making it more convenient for time-varying electric circuits. This method is also more appropriate when the distance between the injection and extraction points is large. Existing sounding methodologies rely on the discrete multitone transceiver (DMT) technique, a baseband version of orthogonal frequency-division multiplexing (OFDM) that covers frequency bands up to 100 MHz; see [6], [7] and references therein. To extend coverage up to 500 MHz, a passband OFDM-based methodology can be used to estimate CFRs.

In this context, this work details a complete OFDM sounding-based methodology for estimating the CFRs of passband PLC channels for data communication in indoor facilities. In this sense, it focuses on the channel estimation and enhancement techniques, as well as the adopted solutions for handling sampling time offset (STO), carrier frequency offset (CFO), and sampling frequency offset (SFO). Using a measurement setup based on software-defined radio (SDR) devices, the methodology is applied to measure PLC channels in the white-space bands from 54 MHz to 72 MHz, 174 MHz to 216 MHz, and 470 MHz to 500 MHz, which are regulated by ANATEL in Brazil [8]. It is also benchmarked against a VNA device. The results show that the proposed methodology and VNA exhibit negligible differences across the three frequency bands, indicating the suitability of the proposed methodology for estimating the CFR of electric power grids up to 500 MHz.

II. PROBLEM FORMULATION

Data communication in baseband and passband configurations entails trade-offs between sampling requirements and implementation complexity. In baseband, the transmitted signal occupies 0 Hz to B and therefore requires $f_s \geq 2B$ by

This work was supported in part by Coordenação de Aperfeiçoamento de Pessoal de Nível Superior (CAPES), Conselho Nacional de Desenvolvimento Científico e Tecnológico (CNPq) under grants 312908/2025-3 and 445958/2024-3, Fundação de Amparo à Pesquisa do Estado de Minas Gerais (FAPEMIG) under grants APQ-05901-25 and APQ-04623-22.

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the Nyquist–Shannon criterion. As the operating frequency increases, this requirement becomes more restrictive. Passband transmission, by contrast, can be described by a complex envelope using in-phase and quadrature (I/Q) components. For a passband bandwidth of $2B$, this representation permits operation with $f_s \geq 2B$. The advantage is offset by greater front-end complexity, since frequency-conversion stages, such as mixers and local oscillators, are needed and may introduce carrier-frequency and phase offsets.

The relation between operating frequency and signal bandwidth is thus central to choosing the representation. When a signal of bandwidth B is concentrated near zero frequency, baseband is natural, possibly with low-frequency spectral nulls to meet regulatory constraints. This is consistent with conventional PLC systems, often described as baseband PLC systems, since transmission must remain near zero frequency to avoid interference with electric power systems and comply with regulations. Broadband PLC systems typically use 1.7 MHz to 100 MHz, whereas narrowband PLC systems operate from 3 kHz to 500 kHz [4]. However, at hundreds of MHz, baseband transmission becomes hardware-intensive. For instance, directly digitizing spectra up to 500 MHz requires sampling frequencies above 1 GHz, which is difficult with low-cost, widely available hardware. Passband transmission reduces this burden by operating over a narrower bandwidth centered at a nonzero carrier frequency. This is particularly relevant to white-space spectrum, whose available bands are centered at relatively high carrier frequencies.

Accordingly, a passband representation is more suitable for PLC systems operating at higher frequencies. Here, these channels are termed passband PLC channels to stress that their spectral content is carrier-centered and that their analysis benefits from signal-processing tools used in passband communication systems. The term does not define a new class of PLC systems; it denotes a modeling perspective especially relevant to transmitting information-carrying signals through electric power circuits at frequencies well above 100 MHz.

Conventional PLC measurement methodologies must therefore be adapted, as they were primarily developed for frequencies up to approximately 100 MHz, where baseband communications apply. When white-space spectrum is considered for PLC systems, baseband transmission may no longer be appropriate. Frequency-conversion stages and their impairments require methodologies that accurately capture the channel response under passband communications. This motivates methodologies designed to characterize passband PLC channels, as detailed in Section III.

III. METHODOLOGY FOR PASSBAND PLC CHANNELS

This section presents the proposed methodology for estimating the CFR of passband PLC channels. It is an offline methodology that uses a top-down approach, thus working with previously acquired data. Fig. 1 shows a block diagram of the proposed methodology to perform all necessary tasks to provide a reliable estimate of CFR, namely: signal generation (OFDM transmitter block); STO estimation and correction (STO block); CFO estimation and correction (CFO block);

channel estimation (CE Block); channel enhancement (CEN block); SFO estimation and correction (SFO block).

A. Signal Generation

Signal generation is the first step of the proposed methodology. OFDM is adopted as the modulation scheme to generate the transmitted signal $x(t)$. Figure 1 illustrates the OFDM transmitter, where the signal generation occurs and which begins with two distinct modulated pseudo-noise (PN) sequences¹, namely A and B. These sequences are grouped into different symbols, both of length $N \in \mathbb{N}$, and then submitted to a serial-to-parallel conversion stage, producing the vectors $\mathbf{X}_i \in \mathbb{C}^{N \times 1}$, with $i \in \{A, B\}$.

The discrete-frequency domain representations of the vector $\mathbf{X}_i$ are then converted to the discrete-time domain vector representation using

$$\mathbf{x}_i = \mathbf{F}^\dagger \mathbf{X}_i, \quad \forall i \in \{A, B\}, \quad (1)$$

where $\mathbf{F} = (1/\sqrt{N})\mathbf{W}$ is the normalized $N \times N$ -size discrete Fourier transform (DFT) matrix and $[\cdot]^\dagger$ is the Hermitian operator.

Two copies of the vector $\mathbf{x}_i$ are concatenated and then cyclically extended. Symbol duplication enables timing synchronization, whereas cyclic extension helps deal more easily with intersymbol interference (ISI). Consequently, the vector representation of one OFDM symbol block is given by

$$\mathbf{x}_{cp,i} = [x_i[N - L_{cp}] \ x_i[N - L_{cp} + 1] \ \dots \ x_i[N - 1] \ \mathbf{x}_i^T \ \mathbf{x}_i^T]^T, \quad (2)$$

in which $L_{cp} \in \mathbb{N}_+$ is the cyclic prefix length, $[\cdot]^T$ indicates the transpose operator. Subsequently, the vectors $\mathbf{x}_{cp,A}$ and $\mathbf{x}_{cp,B}$ are concatenated to form the transmitted vector frame

$$\mathbf{x} = [\mathbf{x}_{cp,A}^T \ \mathbf{x}_{cp,B}^T]^T. \quad (3)$$

This frame is continuously repeated to generate the complex envelope of the discrete-time sequence $\{x[n]\}$. Finally, the digital-to-analog converter (DAC) converts the envelope complex sequence $\{x[n]\}$ into the continuous-time envelope complex signal $x(t)$. In the sequel, it applies an up-converter, denoted by the Up block, to modulate and transmit $x(t)$ through the passband PLC channel.

B. Symbol Timing Offset

At the receiver, after the down-converter, represented by the Down block in Fig. 1, the complex envelope of the received signal in continuous time is given by $y_R(t - t_0) = \tilde{y}_R(t - t_0) + W(t)$, where $\tilde{y}_R(t - t_0) = x'(t) \star h(t)$ is the received signal free of noise, $h(t)$ is the channel impulse response (CIR), $\star$ is the convolution operation, and $W(t)$ is the additive noise. An analog-to-digital converter (ADC) converts $y_R(t - t_0)$ to its discrete-time complex envelope representation, $y_R[n - n_0]$. Subsequently, the next stages of the proposed methodology are applied, beginning with the estimation and correction of STO, the temporal misalignment

¹The PN sequences are adopted due to their ease of implementation and low peak-to-average power ratio, reducing nonlinear distortion introduced by the transmitter amplifier.

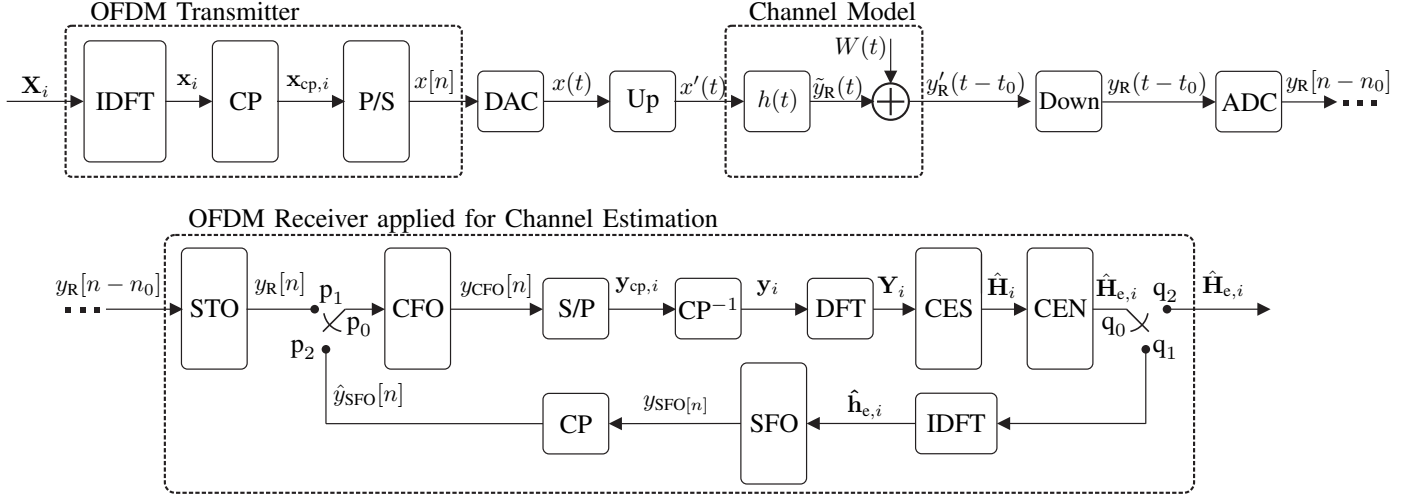


Fig. 1: The block diagram of the methodology for estimating passband CFR.

between the actual start of the transmitted symbol and the sampling instant at the receiver, which can cause ISI and performance degradation. The Schmidl and Cox algorithm [9] is employed to estimate and correct the STO. This technique requires a training symbol composed of two identical halves in the time domain, motivating the concatenation of two copies of the same vector to form the transmitted symbol, as detailed in Subsection III-A. To perform it, a sliding correlation between two successive halves of length $L = N$ samples is performed on the received signal, resulting in a correlation metric given by

$$P[d] = \sum_{n=0}^{L-1} y_R[d+n-n_0] y_R^*[d+n-n_0+L], \quad (4)$$

with $-\infty < d < \infty$, and the corresponding energy metric expressed as

$$R[d] = \sum_{n=0}^{L-1} |y_R[d+n-n_0+L]|^2. \quad (5)$$

Thus, a normalized metric is constructed from the correlation and energy metrics:

$$M[d] = \frac{|P[d]|^2}{(R[d])^2}. \quad (6)$$

The normalized metric exhibits a pronounced plateau when the observation window is aligned with the beginning of the OFDM training symbol due to the correlation between the two identical halves of the transmitted symbol. Finally, the STO estimate $\hat{d}$ is obtained by locating one value of d that maximizes $M[d]$, resulting in $\hat{d} = n_0$. After the correction, the sequence output by the STO block is denoted by $\{y_R[n]\}$.

C. Carrier Frequency Offset

The CFO occurs due to mismatches between the transmitter and receiver local carrier frequencies. In OFDM-based systems, the CFO introduces a linear phase rotation across the time-domain samples and destroys the orthogonality among

subcarriers, causing intercarrier interference (ICI) and, consequently, degrading performance. Similar to the STO, the Schmidl and Cox algorithm can be used to estimate and correct CFO [9].

The phase of the correlation metric $P[d]$, presented in (4), is directly related to the CFO. Therefore, once a perfect timing alignment is achieved (*i.e.*, after using the STO correction), a coarse estimate of the normalized frequency offset with respect to the subcarrier spacing can be obtained using

$$\hat{a} = \frac{1}{2\pi L} \angle P[\hat{d}], \quad (7)$$

In the sequel, the CFO correction can be performed by multiplying the samples of the received signal by a complex exponential with an opposite phase rotation, yielding

$$y_{\text{CFO}}[n] = y_R[n] e^{-j2\pi\hat{a}n}. \quad (8)$$

This compensation removes the linear-phase rotation introduced by the frequency mismatch between the transmitter and receiver oscillators, thereby restoring subcarrier orthogonality and reducing ICI.

D. Channel Frequency Response

To associate the received samples with their corresponding transmitted symbols, the sequence $y_{\text{CFO}}[n]$ is segmented to constitute the vectors $\mathbf{y}_{\text{cp},i}$ of length $N + L_{\text{cp}}$. In the sequel, the block cyclic prefix (CP)⁻¹ performs the CP removal and outputs the vector $\mathbf{y}_i = [y_{\text{cp},\ell_i} \ y_{\text{cp},\ell_i+1} \ \dots \ y_{\text{cp},\ell_i+N-1}]^T$, where $\ell_i = iN + (\lfloor i/2 \rfloor + 1)L_{\text{cp}}$ denotes the i^{th} received vector after CP removal. The vector $\mathbf{y}_{\text{cp},i}$ is associated with the vector $\mathbf{x}_A$ if $i \bmod 4 \in \{0, 1\}$ and with the vector $\mathbf{x}_B$ if $i \bmod 4 \in \{2, 3\}$.

Applying the DFT to $\mathbf{y}_i$ yields the discrete-frequency domain vector $\mathbf{Y}_i \in \mathbb{C}^{N \times 1}$. The transmitted frequency-domain vector associated with the i^{th} vector is given by

$$\mathbf{X}_i = \begin{cases} \mathbf{X}_A, & i \bmod 4 \in \{0, 1\}, \\ \mathbf{X}_B, & i \bmod 4 \in \{2, 3\}. \end{cases} \quad (9)$$

Then, the k^{th} element of the estimated CFR associated with the i^{th} received vector can be obtained as

$$\hat{H}_{i,k} = \frac{Y_{i,k}}{X_{i,k}} \in \mathbb{C}, \quad \forall k \in \{1, 2, \dots, N\}, \quad (10)$$

whereas $\hat{H}_{i,k}$, $Y_{i,k}$, and $X_{i,k}$ are the k^{th} elements of vectors $\hat{\mathbf{H}}$, $\mathbf{Y}_i$, and $\mathbf{X}_i$, respectively.

E. Channel Enhancement

Following [6], the enhancement of the estimated CFR is performed as follows: First, obtain $\hat{\mathbf{h}}_i = (1/\sqrt{N})\mathbf{F}^\dagger \hat{\mathbf{H}}_i$. In the sequel, the ℓ^{th} element of the enhanced version of the estimated CIR vector, $\hat{\mathbf{h}}_{e,i}$, is obtained using the following expression:

$$\hat{h}_{e,i,\ell} = \begin{cases} \hat{h}_{i,\ell}, & \ell = 0, 1, \dots, \hat{L} - 1, \\ 0, & \ell = \hat{L}, \hat{L} + 1, \dots, N - 1. \end{cases} \quad (11)$$

where $\hat{h}_{i,\ell}$ is the ℓ^{th} element of the vector $\hat{\mathbf{h}}_i$, and $\hat{L}$ is chosen to ensure that $\|\hat{\mathbf{h}}_{e,i}\|_2^2 < \alpha \|\hat{\mathbf{h}}_i\|_2^2$ and $\alpha \in \mathbb{R} | 0 \ll \alpha < 1$. Also, we can consider $\hat{L} = L_{\text{cp}}$ because a significant portion of the CIR is expected to be confined into the first L_{cp} elements of the vector $\hat{\mathbf{h}}_i$. Finally, the noise-enhanced version of the estimated CFR is given by $\hat{\mathbf{H}}_{e,i} = \sqrt{N}\mathbf{F}\hat{\mathbf{h}}_{e,i}$.

F. Switches

The switches, p_0 and q_0 , shown in Fig. 1, are used to facilitate the understanding of the methodology for passband PLC channels. As both work synchronously, in state 1, $p_1 \rightarrow p_0$ implicates that the i^{th} OFDM symbol is used to estimate the CFR, and $q_0 \rightarrow q_1$ forwards the enhanced PLC CFR to the input of the SFO estimation and correction block. After this operation, the switches change to state 2. Then, $p_2 \rightarrow p_0$ allows the output of the SFO block to be forwarded to the input of the CFO estimation and correction block, followed by the CFR estimation and channel enhancement stages.

Subsequently, the enhanced PLC CFR is routed once again through the switch q_0 , which now operates in state 2. In this state, $q_0 \rightarrow q_2$ indicates that its input corresponds to a valid estimate of the desired CFR. After the valid CFR estimate is obtained, the switches return to state 1, restarting the processing cycle.

G. Sampling Frequency Offset

The SFO results from a mismatch between the transmitter and receiver sampling clocks, leading to progressive time drift of the received symbols. In OFDM-based systems, SFO introduces a phase distortion that varies linearly with both the subcarrier and OFDM symbol indexes, thereby degrading the subcarrier orthogonality and generating ICI.

After STO and CFO corrections and channel estimation and enhancement, the SFO error is estimated, and a new sequence $\{y_{\text{SFO}}[n]\}$ is obtained through an interpolation technique [6]. The SFO can be estimated from two consecutive received OFDM symbols, using

$$\text{SFO}_{\text{error}} = \frac{1}{N + L_{\text{cp}}} \left(\arg \max_{\ell} \{|\hat{\mathbf{h}}_{e,i+1}[\ell]|\} - \arg \max_{\ell} \{|\hat{\mathbf{h}}_{e,i}[\ell]|\} \right) \quad (12)$$

where $\ell = 0, 1, \dots, N - 1$.

Then, an enhanced version of $\text{SFO}_{\text{error}}$, defined as $\overline{\text{SFO}}_{\text{error}}$, can be calculated by averaging the error values from each pair of consecutive improved CIRs. Based on the $\overline{\text{SFO}}_{\text{error}}$ value, the n^{th} sample of the sequence $\{y_{\text{SFO}}[n]\}$ is obtained using [10]:

$$y_{\text{SFO}}[n] = \sum_{\alpha=0}^{N_p} \mu_n^\alpha \sum_{\lambda=0}^{L_1-1} c_\alpha[\lambda] y_R[m_n - \lambda], \quad (13)$$

where L_1 is the number of taps of the finite impulse response (FIR) interpolating filter, N_p is the polynomial's degree, $c_\alpha[\lambda]$ are independent coefficients, $\mu_n = (1 - \overline{\text{SFO}}_{\text{error}}) - m_n$, and $m_n = \lfloor n(1 - \overline{\text{SFO}}_{\text{error}}) \rfloor$, in which $\lfloor x \rfloor = \max\{m \in \mathbb{Z} | m \leq x\}$ denotes the largest integer not exceeding x . The values of $c_\alpha[\lambda]$, N_p , and L_1 can be obtained in [11]. Thus, the output of block SFO is cyclically extended, generating $\hat{y}_{\text{SFO}}[n]$, and redirected to the entrance of the CFO block.

IV. NUMERICAL RESULTS

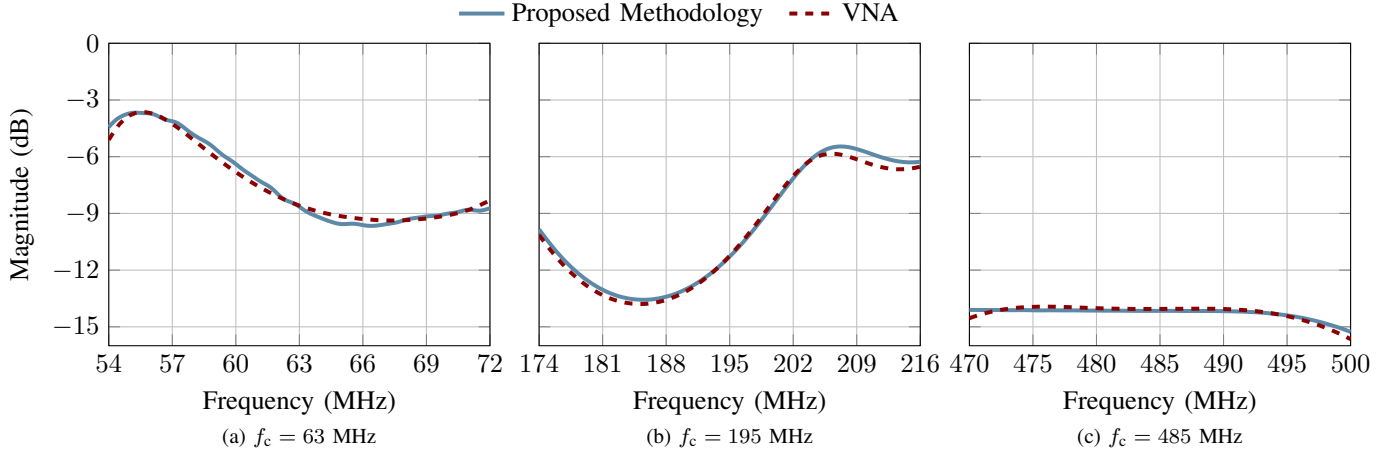
The numerical results in this section are based on raw data obtained from a measurement campaign conducted in the Laboratory of Communication (LCOM) at Federal University of Juiz de Fora (UFJF). The experimental setup is shown in Figure 2. It is constituted by a transmitter that is composed of a personal computer (PC) for generating $\{x[n]\}$, a USRP X310 SDR [12] for producing the transmitted signal $x(t)$, and a capacitive PLC coupling device for injecting $x(t)$ into the power line cable. The power line cable is not connected to the electric power grid to ensure that the measured PLC channel is time-invariant, thereby allowing validation of the proposed methodology against a VNA. The receiver uses a capacitive PLC coupling device to extract the signal from the electric power cable, a USRP X310 SDR to obtain $\{y[n]\}$ from $y(t)$, and a PC to store the samples of $\{y[n]\}$. Table I lists the parameters used in the CFR estimation methodology.



Fig. 2: The experimental setup.

Three different central frequencies were considered: 63 MHz, 195 MHz, and 485 MHz. Therefore, the frequency

Fig. 3: Comparison between PLC CFR estimates achieved by the presented methodology and a VNA.



bands used for transmission were 54 to 72 MHz, 174 to 216 MHz, and 470 to 500 MHz because they constitute the white space spectrum according to ANATEL [8]. The estimated CFRs were averaged for each individual frequency band. Moreover, each measurement of the PLC channel produced 4880 successive estimates of CFRs, and altogether 14,640 CFRs were collected across the three frequency bands, with 4880 CFR estimates acquired per band. This large number of estimates was used to obtain an average CFR with a small additive-noise influence. Moreover, to validate the proposed methodology, the obtained average CFRs were compared with measurements from an Agilent E5061B VNA.

TABLE I: Parameters employed in the proposed methodology.

Number of sub-carriers	$N = 2048$
CP length	$L_{cp} = 256$
Frequency band	$B = 50$ MHz
Carrier frequencies	$f_c \in \{63, 195, 485\}$ MHz

Figure 3 shows the magnitude spectrum of the CFRs obtained with the proposed methodology and the VNA for the three frequency bands. One can note that the proposed methodology yields results that closely match those obtained with the VNA for the analyzed frequency bands. Differences below 0.5 dB are observed only in specific frequency ranges: 63 to 67 MHz, 205 to 215 MHz, and near 500 MHz. These localized discrepancies may be associated with residual synchronization inaccuracies, variations in front-end frequency response, and noise inherent to the measurement setup. Nevertheless, the overall agreement across the analyzed bands indicates that the proposed methodology can be applied to estimate passband PLC CFRs.

V. CONCLUSIONS

This study presented a methodology for estimating passband PLC CFRs that combines channel estimation and enhancement with STO, CFO, and SFO corrections. Its effectiveness was validated experimentally by comparing the PLC CFR estimates obtained using the proposed methodology with those from a VNA. The attained results show that the proposal can reliably

estimate passband PLC CFRs. Therefore, it can support a better understanding of passband PLC channels and foster research on the use of white-space spectrum by PLC systems.

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