

# Simultaneous Cancellation of Narrow Band Interference and Impulsive Noise in PLC Systems

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**Abstract**—The two major sources of disturbances for an efficient and reliable data transmission through power lines, known as power line communication (PLC), are impulsive noise (IN) and narrow band interference (NBI). In this paper, we propose an algorithm to cancel the IN and NBI simultaneously for an OFDM based PLC system. The proposed method exploits the duality of the problem, where the IN is sparse in the time domain and the NBI is sparse in the frequency domain. By virtue of this duality, we use multiple signal characterization (MUSIC) algorithm to estimate both the IN support (in time domain) and the frequency of NBI (in frequency domain). Furthermore, the minimum mean square error (MMSE) estimator is used to estimate the amplitude and phase of IN samples at the determined locations and the least square (LS) estimator is used to estimate the amplitude and phase of the NBI. Finally, the estimated IN and NBI are canceled out from the received signal, providing noise mitigated samples for demodulation. The performance of the proposed scheme is verified via numerical simulations.

**Index Terms**—Impulsive noise, LS estimator, MUSIC, MMSE, narrow band interference, power line communication.

## I. INTRODUCTION

Narrow band power line communication (NB-PLC) has been proposed as enabler of the smart grid application in the electrical grid networks [1]. The NB-PLC system, which operates in the low frequency range of 10-490 kHz, suffers from background noise, impulsive noise, narrow band interference and the frequency selective multipath behavior of the channel [2]. To overcome the frequency selectivity of the channel, orthogonal frequency division multiplexing (OFDM) has been proposed for the NB-PLC by the existing standards, namely: IEEE 1901.2, PRIME and G3-PLC [1]. In order to mitigate the effects of impulsive noise (IN) and narrow band interference (NBI), some additional efficient signal processing techniques are required [3].

On one hand, the IN, which is sparse in nature in the time domain, presents a very high amplitude as compared to the background noise. If not mitigated, these IN samples damage symbols in all the sub-carriers [4], [5]. On the other, the NBI is sparse in the frequency domain and therefore its presence typically affects only the sub-carrier that is hit by it [6]–[8]. A significant amount of research work that deals with the problem of mitigating the impulsive noise (IN) and the narrow band interference (NBI) can be found in the literature. The studies done so far do not consider the joint effect of the NBI and IN in an OFDM based PLC system. However,

for a reliable data transmission through the power lines, joint mitigation of both impairments is required. Some algorithms based on non-linear techniques, such as nulling, clipping and combination of both [5], [9], [10] as well as some recent proposals based on the compressive sensing (CS) technique [7] have been proposed in order to mitigate the IN in an OFDM system. Similarly, non-linear techniques based on frequency excision/nulling and clipping in the frequency domain along with the CS based algorithm can also be found in the literature, addressing the problem of mitigating the NBI [6], [11].

Here, we propose an algorithm to mitigate both the NBI and IN consecutively. The support of IN samples and the frequency of NBI are determined by the multiple-signal-characterizing (MUSIC) method, exploiting the duality of the problem in the time and frequency domains. The amplitude and phase of each IN sample at the determined location is estimated by the minimum mean square error (MMSE) estimator. Furthermore, the amplitude and phase of the NBI are estimated by the least square (LS) estimator. To present the algorithm in concise manner, the paper is divided into four sections. Section II describes the system model. Section III elaborates the proposed algorithm. Section IV discusses the simulation results and section V provides some conclusions.

*Notations:* The following notation will be used in this paper: A variable in bold lower case,  $\mathbf{a}$ , defines a vector in the time domain having elements as  $\{a_1, a_2, \dots, a_n\}$ . A variable in bold uppercase,  $\mathbf{A}$ , defines a matrix in the time domain. A variable in bold lowercase with a bar,  $\bar{\mathbf{a}}$ , defines a vector in the frequency domain having elements as  $\{\bar{a}_1, \bar{a}_2, \dots, \bar{a}_n\}$ . Finally, a variable in bold upper case with a bar,  $\bar{\mathbf{A}}$ , defines a matrix in the frequency domain.

## II. SYSTEM MODEL

The discrete-time complex baseband equivalent model for the OFDM system under consideration can be written as

$$\mathbf{r}^{(i)} = \mathbf{H}^{(i)} \mathbf{x}^{(i)} + \mathbf{e}^{(i)} + \mathbf{i}^{(i)} + \mathbf{w}^{(i)}, (M \times 1). \quad (1)$$

The variable  $i \in \{1, 2, \dots, N_{\text{frame}}\}$  in (1) indexes the frame number, where  $N_{\text{frame}}$  is the total number of frames, and where each frame is  $M$  samples long. The vector  $\mathbf{r}^{(i)} = [r_1^{(i)}, r_2^{(i)}, \dots, r_M^{(i)}]^T$  in (1) contains the time domain received samples transmitted in the  $i^{\text{th}}$  frame. Since the proposed algorithm works taking a frame at a time, for the sake of simplicity

the super index variable  $i$  is dropped in the equations to follow. The matrix  $\mathbf{H}$  in (1) is the circular-convolutional channel matrix, which is assumed to be static for the duration of a frame transmission. The time domain transmitted signal vector  $\mathbf{x} = [x_1, \dots, x_{N_p}, x_{(N_p+1)}, \dots, x_M]^T$  contains the transmitted time domain samples in a transmission frame as shown in Fig. 1. The first  $N_p$  samples in the frame correspond to the time domain samples of the preamble symbol. The remaining  $M - N_p$  samples correspond to the time domain samples of the cyclic prefix prepended OFDM symbol, where the length of the cyclic prefix is  $N_{cp}$ . The total number of sub-carriers in the system is denoted by  $N$ . Among these, only  $N_u$  sub-carriers are used for data transmission. The remaining  $N - N_u$  sub-carriers are nulled/unused. The vector  $\mathbf{i} \in \mathbb{C}^{M \times 1}$  contains the time domain samples of the IN. The IN samples denoted by  $\mathbf{i} = \{i_1, i_2, \dots, i_M\}$  are modeled as a Bernoulli-Gaussian process. Each sample of the IN is denoted by:

$$i_n = b_n \cdot g_n, \quad (2)$$

where  $n = \{1, 2, \dots, M\}$  defines the sample index,  $b$  is a Bernoulli distributed random variable having probability of success  $P$  and  $g$  denotes an independent and identically distributed (i.i.d) Gaussian random variable with variance  $\sigma_i^2$  and zero mean. The variance  $\sigma_i^2$  denotes the IN power. The vector  $\mathbf{e} \in \mathbb{C}^{M \times 1}$  contains the time domain samples of the frequency interferer. The frequency interferer is the sum of all the NBIs occurring during a frame transmission and is modeled by collection of superimposed tones. The adopted interference model is widely used to characterize the effect of multiple NBIs in an OFDM system [6]–[8]. According to this model, each NBI is characterized by a tone in the frequency spectrum. The occurrence of multiple NBIs is therefore modeled as the sum of complex sinusoids (equivalent to the number of NBIs) in the time domain. The discrete time representation of the frequency interferer samples,  $\mathbf{e} = \{e_1, e_2, \dots, e_M\}$ , is written as

$$e_n = \sum_{k=1}^c A_k \exp^{j(\omega_k n T + \phi_k)}, \quad (3)$$

where  $A_k$  is the amplitude,  $\omega_k$  is the normalized frequency and  $\phi_k$  is the phase of the  $k^{th}$  NBI. The variable  $c$  denotes the number of NBIs and  $T$  is the sampling period.

Finally, the vector  $\mathbf{w} \in \mathbb{C}^{M \times 1}$  contains the time domain samples of the background noise, occurring during the transmission of the  $i^{th}$  frame. The background noise is modeled as AWGN, which is defined as a sequence of i.i.d complex Gaussian random variables with zero mean and variance  $\sigma_w^2$ . The interference to signal ratio (ISR) of the individual NBI is defined as  $A_k^2/\sigma_s^2$ , and the signal to noise ratio (SNR) is defined as  $\sigma_s^2/\sigma_w^2$ , where  $\sigma_s^2$  is the transmitted signal power. Likewise, the impulsive noise to background noise ratio (INR) is defined as  $\sigma_i^2/\sigma_w^2$ . For the following derivations, we define the unitary discrete Fourier transform matrix  $\mathbf{F}$  having elements as:  $f_{a,b} = \frac{1}{\sqrt{N}} \exp^{-j2\pi ab/N}$  with  $a, b \in \{0, 1, 2, \dots, N - 1\}$ , where  $a, b$  denotes the row and column indexes in the matrix.

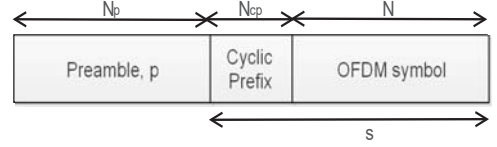


Fig. 1. Transmission frame structure.

### III. PROPOSED NBI AND IN CANCELLATION SCHEME

The frequency interferer is assumed to be deterministic for the duration of a frame transmission. Therefore the estimation of the parameters of the frequency interferer is done using the received samples corresponding to the transmitted preamble symbol. Conversely, we exploit the spectrum of the unused sub-carriers in the system, to estimate the IN. Hence, IN is estimated on symbol-by-symbol basis, where IN in the preamble symbol and transmitted symbol is estimated separately. The schematic diagram of the proposed scheme is shown in Fig. 2.

Under a perfect timing and frequency synchronization, the preamble observation vector is formed by selecting the first  $N_p$  samples out of the  $M$  received samples in (1) as

$$\mathbf{r}_p = \mathbf{S}_x \mathbf{r} = \mathbf{H} \mathbf{x}_p + \mathbf{e}_p + \mathbf{i}_p + \mathbf{w}_p, (N_p \times 1). \quad (4)$$

The matrix  $\mathbf{S}_x$  in (4) is the square selection matrix of size  $N_p \times M$  having a single element equal to 1 per column in the first  $N_p$  columns identifying the locations of preamble samples and zeros in rest of the columns. The vectors  $\mathbf{r}_p, \mathbf{x}_p, \mathbf{e}_p, \mathbf{i}_p$  and  $\mathbf{w}_p$  each of length  $N_p$ , in the resulting equation (4), contain the time domain received samples corresponding to the preamble symbol transmission. The quantities  $(\mathbf{H} \mathbf{x}_p), \mathbf{i}_p$  and  $\mathbf{w}_p$  in (4) are the undesired signals when it comes to the estimation of the frequency interferer parameters using (4) as the observation vector. In order to address the presence of such undesired signals, an iterative algorithm is proposed herein. First of all we get rid of the IN component ( $\mathbf{i}_p$ ) in the observation vector, followed by the subtraction of the channel ( $\mathbf{H}$ ) and preamble symbol ( $\mathbf{x}_p$ ) effect in the observation vector  $\mathbf{r}_p$ . Having done so we achieve an observation vector bearing the samples of the frequency interferer and background noise only. Hence, the precise and final estimation of the frequency interferer is then carried out.

To get rid of the IN component, we adopt an iterative implementation of the true support estimation (TSE) algorithm proposed in [4]. As the first iteration we identify the IN contaminated samples in,  $\mathbf{r}_p$ , using the TSE scheme, and null them. The resulting observation vector, denoted by  $\mathbf{r}_p^*$ , is now used to perform a preliminary estimation of the frequency interferer, denoted by  $\hat{\mathbf{e}}_{pre}^*$ . This preliminary estimate of the frequency interferer is subtracted from  $\mathbf{r}_p$  resulting in

$$\mathbf{r}_p' = \mathbf{H} \mathbf{x}_p + (\mathbf{e}_p - \hat{\mathbf{e}}_{pre}^*) + \mathbf{i}_p + \mathbf{w}_p, (N_p \times 1). \quad (5)$$

The resulting observation vector,  $\mathbf{r}_p'$ , in (5) is then used to estimate the IN occurring during the preamble symbol transmission. We again take advantage of the TSE algorithm to perform the IN estimation at this stage [4]. The estimated

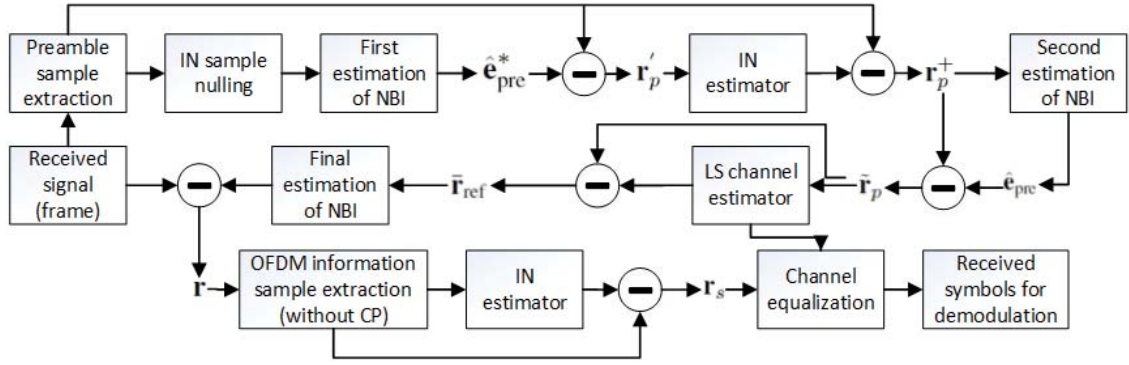


Fig. 2. Schematic block diagram of the proposed algorithm.

IN noise,  $\hat{\mathbf{i}}_p$ , is then subtracted from the observation vector  $\mathbf{r}_p$ , resulting in

$$\mathbf{r}_p^+ = \mathbf{H}\mathbf{x}_p + \mathbf{e}_p + (\mathbf{i}_p - \hat{\mathbf{i}}_p) + \mathbf{w}_p, (N_p \times 1). \quad (6)$$

The resulting equation in (6) is now used for the estimation of the NBI.

The component  $(\mathbf{H}\mathbf{x}_p)$  in  $\mathbf{r}_p^+$  is an undesired signal affecting the precise frequency interferer estimation. Therefore, we again proceed with the iterative estimation of the frequency interferer. Next, a second preliminary estimation of the interferer is done. This noisy preliminary estimate of the interferer is subtracted from  $\mathbf{r}_p^+$ , followed by the channel estimation using the known preamble symbol at the receiver. Furthermore, with the known preamble and the approximated channel coefficients, the contribution of both the channel and preamble from the observation vector is minimized. This minimization provides a refined observation vector that contains samples only from the NBIs and the background noise. The final frequency interferer estimation is done using the refined observation vector.

**Note:** The estimation of  $\hat{\mathbf{e}}_{\text{pre}}^*$  is also done following the procedure as mentioned in the following sub-section *Frequency Interferer Parameter Estimation* using  $\mathbf{r}_p^*$ .

#### A. Frequency Interferer Parameter Estimation

Taking  $\mathbf{r}_p^+$  as an observation vector we proceed towards estimating the second preliminary estimate of the frequency interferer after the removal of the IN. The estimation of the parameters  $A_k$ ,  $\omega_k$  and  $\phi_k$  of the interferer from the equation (6) will be carried out in three steps.

- Estimation of the number of NBIs (order estimation)
- Frequency estimation of each NBI
- Amplitude and phase estimation of each NBI

1) *Order Estimation of Frequency Interferer:* The order of the frequency interferer determines the number of the NBIs occurring during a frame transmission. To estimate the number of NBIs, we start by choosing an arbitrary number  $l$ , as the sample window size, to generate  $L = N_p - l + 1$  sample

vectors from the observation vector in (6). Each sample vector of length  $l$  is denoted by  $\tilde{\mathbf{r}}_b$ ,  $b = \{1, 2, 3, \dots, L\}$ .

$$\tilde{\mathbf{r}}_b = [r_{p(b+l-1)}^+, r_{p(b+l-2)}^+, \dots, r_{p_b}^+]^T, (l \times 1), \quad (7)$$

the variables identified by  $r_p^+$  are the elements of  $\mathbf{r}_p^+$  vector, whose locations are defined by the sub-scripted variables. The window size  $l$ , to form the sample vector, should be chosen such that the condition  $l - c > c$ , is satisfied. It is also required that the size of  $l$  should not be larger than  $N_p/2$ , where  $N_p$  is the length of the observation vector. A sample window size larger than  $N_p/2$  and not satisfying  $l - c > c$ , will in fact degrade the performance of the MUSIC estimator [12]. After generating the  $L$  sample vectors, a sample covariance matrix  $\mathbf{C}$  of size  $l \times l$  is formed as

$$\mathbf{C} = \frac{1}{L} \sum_{b=1}^L \tilde{\mathbf{r}}_b \tilde{\mathbf{r}}_b^H, (l \times l). \quad (8)$$

The number of NBIs is estimated by evaluating the eigenvalues of the sample covariance matrix, given by the eigenvalue decomposition (ED) of the matrix  $\mathbf{C}$ . The ED of  $\mathbf{C}$  results into  $l$  eigenvectors and  $l$  eigenvalues. The eigenvalues arranged in decreasing order  $\lambda_1 \geq \lambda_2 \geq \lambda_3 \geq \dots, \lambda_l$ , are then evaluated according to the MDL criterion, from where the number of NBIs  $c$  is estimated [13].

2) *Frequency Estimation of Each NBI:* After estimating the number of NBIs, in this step we estimate their frequencies by using the high resolution frequency estimator MUSIC.

Based on the estimated number of interferers, the eigenvectors of  $\mathbf{C}$  are classified into two sub-sets. The first sub-set, denoted by  $\hat{\mathbf{S}} = \{\hat{\mathbf{a}}_1, \dots, \hat{\mathbf{a}}_c\}$ , contains the  $c$  eigenvectors associated to the  $c$  largest eigenvalues of  $\mathbf{C}$  also referred to as signal-subspace in MUSIC terminology, and the second sub set  $\hat{\mathbf{G}} = \{\hat{\mathbf{b}}_1, \dots, \hat{\mathbf{b}}_{l-c}\}$  contains the remaining  $l - c$  eigenvectors of  $\mathbf{C}$  namely the noise-subspace in MUSIC terminology. Furthermore, we define the vector  $\boldsymbol{\alpha}$  as a function of  $\hat{\omega}$  as:

$$\boldsymbol{\alpha}(\hat{\omega}) = [1, e^{-j\hat{\omega}}, \dots, e^{-j(l-1)\hat{\omega}}]^T, \hat{\omega} \in [0, 2\pi]. \quad (9)$$

Based on the estimated value of  $c$ , the values of  $\hat{\omega}$  corresponding to the  $c$  largest peaks of the pseudo-periodogram function  $\hat{f}(\hat{\omega})$ , as defined in (10), are the estimated frequencies of the interferers [12],

$$\hat{f}(\hat{\omega}) = \frac{1}{\boldsymbol{\alpha}^H(\hat{\omega})\hat{\mathbf{G}}\hat{\mathbf{G}}^H\boldsymbol{\alpha}(\hat{\omega})}. \quad (10)$$

3) *Amplitude and Phase Estimation of Each NBI*: After estimating the number and the corresponding frequencies of the NBIs, in this step we perform amplitude and phase estimation of each NBI using the LS estimator.

Without loss of generality, let

$$z_k = A_k e^{j\phi_k}, k \in \{1, 2, \dots, c\}, \quad (11)$$

be the variable containing the values of amplitude and phase of the  $k^{\text{th}}$  NBI. Consider the column vector

$$\mathbf{z} = [z_1, \dots, z_c]^T, (c \times 1). \quad (12)$$

where the variables  $z_k$  are defined in (11).

Let,

$$\mathbf{Q} = \begin{bmatrix} e^{j\hat{\omega}_1} & \dots & e^{j\hat{\omega}_c} \\ \vdots & & \vdots \\ e^{jN_p\hat{\omega}_1} & \dots & e^{jN_p\hat{\omega}_c} \end{bmatrix}, (N_p \times c) \quad (13)$$

be the matrix of size  $N_p \times c$  constructed using the estimated frequencies of NBIs, where  $\{\hat{\omega}_1, \dots, \hat{\omega}_c\}$  are the estimated frequencies.

Using (12) and (13), (6) can now be expressed as:

$$\mathbf{r}_p^+ = \mathbf{Q}\mathbf{z} + \text{noise}, (N_p \times 1), \quad (14)$$

where the term ‘‘noise’’ corresponds to the received samples of the preamble and background noise in (6). From (14), the vector  $\mathbf{z}$  is estimated by using the LS estimator. The LS estimated vector  $\hat{\mathbf{z}}$  of  $\mathbf{z}$ , is thus given by

$$\hat{\mathbf{z}} = (\mathbf{Q}^H\mathbf{Q})^{-1}\mathbf{Q}^H\mathbf{r}_p^+, (c \times 1), \quad (15)$$

where  $[\cdot]^H$  defines the Hermitian operation. With estimated order, amplitude, frequency and phase, a preliminary estimate of the frequency interferer,  $\hat{\mathbf{e}}_{\text{pre}} = \{\hat{e}_{\text{pre}_1}, \hat{e}_{\text{pre}_2}, \dots, \hat{e}_{\text{pre}_{N_p}}\}$  is

$$\hat{e}_{\text{pre}_{(\tilde{n})}} = \sum_{k=1}^c \hat{z}_k e^{j(\hat{\omega}_k \tilde{n} T)}, \quad (16)$$

where  $\tilde{n} = 1, 2, \dots, N_p$  and  $\hat{z}_k$  is the estimated value of  $z_k$ .

### B. Observation Vector Refinement

The preliminary estimate of the frequency interferer,  $\hat{\mathbf{e}}_{\text{pre}}$ , is subtracted from the observation vector in (6),

$$\tilde{\mathbf{r}}_p = \mathbf{r}_p^+ - \hat{\mathbf{e}}_{\text{pre}} = \mathbf{H}\mathbf{x}_p + (\mathbf{e}_p - \hat{\mathbf{e}}_{\text{pre}}) + \mathbf{w}_p, (N_p \times 1). \quad (17)$$

The resulting samples in (17) are now used for the channel estimation. The known preamble symbol at the receiver is used to approximate the least square estimate of the channel coefficients. The main motivation behind performing this step is to estimate the channel coefficients such that the contribution

from preamble symbol and the channel in the observation vector can be subtracted.

The least square (LS) estimation of the channel is performed after transforming the samples in (17) to frequency domain. The frequency domain transformation is carried out by multiplying (17) with the DFT matrix  $\mathbf{F}$ .

$$\bar{\mathbf{r}}_p = \mathbf{F}\tilde{\mathbf{r}}_p = \bar{\mathbf{H}}\bar{\mathbf{x}}_p + \text{NBI}_{\text{residue}} + \bar{\mathbf{w}}_p, (N_p \times 1), \quad (18)$$

where  $\bar{\mathbf{r}}_p$  is the vector containing the received symbols corresponding to the transmitted preamble symbol,  $\bar{\mathbf{H}}$  is the diagonal matrix containing the coefficients of the channel frequency response,  $\bar{\mathbf{x}}_p$  is the vector of the received symbols corresponding to the transmitted symbols in the preamble,  $\text{NBI}_{\text{residue}}$  is the residue of interference and  $\bar{\mathbf{w}}_p$  contains the frequency domain samples of the complex Gaussian background noise.

The known preamble symbol  $\bar{\mathbf{x}}_{\text{preamble}}$  (transmitted in used sub-carriers) at the receiver, is used to determine the least square (LS) estimate of the channel coefficients from (18) as

$$\bar{\mathbf{h}}_{LS} = \bar{\mathbf{r}}_{p_{\text{used}}} ./ \bar{\mathbf{x}}_{\text{preamble}}, (N_u \times 1), \quad (19)$$

where  $./$  defines the element wise division operation, the vector  $\bar{\mathbf{r}}_{p_{\text{used}}}$  contain the received symbols in used sub-carriers and the resulting vector  $\bar{\mathbf{h}}_{LS}$  in (19), contains the estimated channel coefficients.

With the estimated channel coefficients and known preamble, the effect of the corresponding quantities from (6) is eliminated, providing a refined observation vector for the final frequency interferer estimation. To achieve the refined observation vector, the samples in (6) are transformed to the frequency domain first and then the subsequent quantities are subtracted as:

$$\bar{\mathbf{r}}_{\text{ref}} = \mathbf{F}\mathbf{r}_p^+ - \bar{\mathbf{H}}_{LS}\bar{\mathbf{x}}_p = (\bar{\mathbf{H}} - \bar{\mathbf{H}}_{LS})\bar{\mathbf{x}}_p + \bar{\mathbf{e}}_p + \bar{\mathbf{w}}_p, (N_p \times 1). \quad (20)$$

where  $\bar{\mathbf{H}}_{LS}$  is the diagonal matrix with elements of the vector  $\bar{\mathbf{h}}_{LS}$  at the locations corresponding to the used sub-carriers and zeros at unused sub-carriers location. Hence  $\bar{\mathbf{r}}_{\text{ref}}$  is the vector essentially containing frequency domain samples of NBI ( $\bar{\mathbf{e}}_p$ ) and the background complex Gaussian noise ( $\bar{\mathbf{w}}_p$ ).

The IDFT of (20) is now the required refined time domain observation vector for the final frequency interferer estimation. The IDFT of (20) is done by multiplying it with  $\mathbf{F}^H$ ,

$$\mathbf{r}_{\text{ref}} = \mathbf{F}^H\bar{\mathbf{r}}_{\text{ref}} = \mathbf{e}_p + \mathbf{w}_p + \text{residue}, (N_p \times 1). \quad (21)$$

The resulting vector  $\mathbf{r}_{\text{ref}}$  in (21) contains the time domain samples of the frequency interferer and the background noise, with some residue from the channel estimation done in the presence of  $\text{NBI}_{\text{residue}}$  in (18).

### C. Final Frequency Interferer Estimation and Cancellation

The final estimation of the frequency interferer is carried out by following the steps exactly as mentioned in the subsection *Frequency Interferer Parameter Estimation*, using the new observation vector  $\mathbf{r}_{\text{ref}}$  from (21).

The purpose of performing such an iterative estimation operation, is to enhance the precision in the NBI parameter estimation. The first ( $\hat{\mathbf{e}}_{\text{pre}}^*$ ) and second ( $\hat{\mathbf{e}}_{\text{pre}}$ ) preliminary estimated frequency interferer are very noisy, and hence a refined frequency interferer estimation needs to be achieved. The significance of iterative estimation becomes more evident at high SNR values. At high SNR, the preamble symbol with high power acts as strong noise for the interferer estimation. Hence, removal of the preamble along with the channel effect considerably improves a precise interference estimation.

Since the NBI is assumed to be deterministic for the duration of the frame transmission, the individual element of the final estimate vector,  $\hat{\mathbf{e}} = \{\hat{e}_1, \hat{e}_2, \dots, \hat{e}_M\}$ , is now reconstructed as

$$\hat{e}(n) = \sum_{k=1}^c \hat{A}_k \cdot e^{j(\hat{\omega}_k n T + \hat{\phi}_k)}, \quad (22)$$

where  $\hat{A}$ ,  $\hat{\omega}$ , and  $\hat{\phi}$  are the estimated values of  $A$ ,  $\omega$ , and  $\phi$ . This final estimate of the frequency interferer is canceled out from the received signal in (1).

$$\mathbf{r} = \mathbf{H}\mathbf{x} + (\mathbf{e} - \hat{\mathbf{e}}) + \mathbf{i} + \mathbf{w}, \quad (M \times 1). \quad (23)$$

#### D. IN Estimation and Cancellation for the Transmitted Symbol

First, the time domain received samples corresponding to the transmitted OFDM information symbol is extracted. The selection of these samples is done by multiplying (23) with a selection matrix  $\mathbf{S}_{\bar{x}}$ , i.e.

$$\mathbf{r}_s = \mathbf{S}_{\bar{x}}\mathbf{r} = \mathbf{H}\mathbf{x}_s + \mathbf{i}_s + \mathbf{w}_s, \quad (N \times 1), \quad (24)$$

The matrix  $\mathbf{S}_{\bar{x}}$  in (24) is the selection matrix of size  $(N \times M)$  having single element equal to 1 per column in the last  $N$  columns identifying the locations of the transmitted symbol after the cyclic prefix. The vectors  $\mathbf{r}_s$ ,  $\mathbf{x}_s$ ,  $\mathbf{i}_s$  and  $\mathbf{w}_s$  in (24) contain the time domain samples of received symbol, transmitted symbol and the samples of IN and background noise occurring during the transmission of the OFDM information symbol.

Furthermore, the vector  $\mathbf{r}_s$  is transformed to the frequency domain and the spectrum of the unused sub-carriers is extracted. To do so, the vector  $\mathbf{r}_s$  is multiplied by the DFT matrix  $\mathbf{F}$ , followed by extraction of the spectrum corresponding to the unused sub-carriers by multiplying it with a selection matrix  $\mathbf{S}_u$  of dimension  $((N - N_u) \times N)$  having a single element equal to 1 per column at the positions that identifies the locations of unused sub-carriers in the system.

$$\bar{\mathbf{r}}_s = \mathbf{F}\mathbf{r}_s = \bar{\mathbf{H}}\bar{\mathbf{x}}_s + \bar{\mathbf{i}}_s + \bar{\mathbf{w}}_s, \quad (N \times 1). \quad (25)$$

$$\bar{\bar{\mathbf{r}}}_s = \mathbf{S}_u\bar{\mathbf{r}}_s, \quad (N - N_u \times 1). \quad (26)$$

The resulting vector  $\bar{\bar{\mathbf{r}}}_s$  is the observation vector for the TSE algorithm to estimate the IN samples occurring during the transmission of the OFDM symbol in the frame [4]. The resulting estimate of the IN,  $\hat{\mathbf{i}}_s$  is subtracted from the received samples  $\mathbf{r}_s$ . The final received samples corresponding to the

transmitted information at the receiver, after the cancellation of NBI and IN, can now be expressed as:

$$\mathbf{r}_s = \mathbf{H}\mathbf{x}_s + (\mathbf{i}_s - \hat{\mathbf{i}}_s) + \mathbf{w}_s, \quad (27)$$

where  $\mathbf{r}_s$  is only affected by the background noise samples and the multipath effect of the channel. The equalization of the multipath effect of the channel is done by using the channel coefficients estimated in (19).

## IV. SIMULATIONS AND RESULTS

In this section, we evaluate the BER performance of the proposed algorithm. The system parameters, preamble symbol, and the channel model are derived from the NB-PLC standard, IEEE 1901.2. The system parameters for the CENELEC-A band OFDM system in consideration are shown in the table below. The channel follows statistical multi-path fading

TABLE I  
SIMULATION PARAMETERS

IEEE 1901.2 OFDM system parameters	
Parameter	Value
Sampling frequency, $F_s$	400 kHz
Total sub carriers, $N$	256
Length of preamble symbol, $N_p$	256
Cyclic prefix length, $N_{cp}$	30
Inter carrier spacing	1.5625 kHz
First used sub-carrier	35.9375 kHz
Last used sub-carrier	90.625 kHz
Modulation	QPSK (uncoded)

channel model, which is characterized by:

$$H(f) = \sum_{t=1}^{N_{\text{path}}} g_t \cdot e^{-(a_0 + a_1 f)d_t} \cdot e^{-j2\pi f \frac{d_t}{v_o}} \quad (28)$$

where  $N_{\text{path}}$  is the total number of propagation paths between the transmitter and the receiver,  $g_t$  is the path gain summarizing the reflection and the transmission along  $t^{\text{th}}$  propagation path,  $a_0$  and  $a_1$  are the attenuation parameters that depend on the transmission line impedance characteristics,  $f$  is the frequency in Hertz,  $d_t$  is the length of  $t^{\text{th}}$  propagation path and  $v_o$  is the wave propagation speed. The above channel model is implemented using the realistic parameter values, as given in the standard IEEE 1901.2, where  $a_0 = 1 \times 10^{-3}$ ,  $a_1 = 2.5 \times 10^{-3}$ ,  $d$  is the Gaussian random variable having mean 1000 and standard deviation 400,  $g$  is also a Gaussian random variable with zero mean and variance 1 that is scaled by 1000,  $N_{\text{path}} = 5$  and  $v_o = 3/4 \times 10^8$  [1].

To demonstrate the performance of the proposed algorithm in different scenarios, we define two scenarios characterized as:

- **scenario I:** One NBI, whose frequency is uniformly distributed between  $[0, F_s]$ , occurs during a frame transmission and IN samples occur with the probability of success,  $P = 1.2 \times 10^{-2}$ .
- **scenario II:** Two NBIs, whose frequencies are uniformly distributed between  $[0, F_s]$ , occur during a frame transmission and IN samples occur with the probability of success,  $P = 2 \times 10^{-2}$ .

## V. CONCLUSION

A novel noise mitigation technique for PLC systems has been proposed in this paper. The proposed scheme is robust against both narrow band interference and impulsive noise. Due to the high precision in parameter estimation, precise estimation and hence the cancellation of both types of noise can be done. The performance of the presented scheme is close to the AWGN bound and is consistent over different scenarios having distinct level of NBI and IN power, as verified by simulation results. The use of the same algorithm for estimating both the IN samples support and the NBI frequency, also facilitates an efficient PLC receiver architecture.

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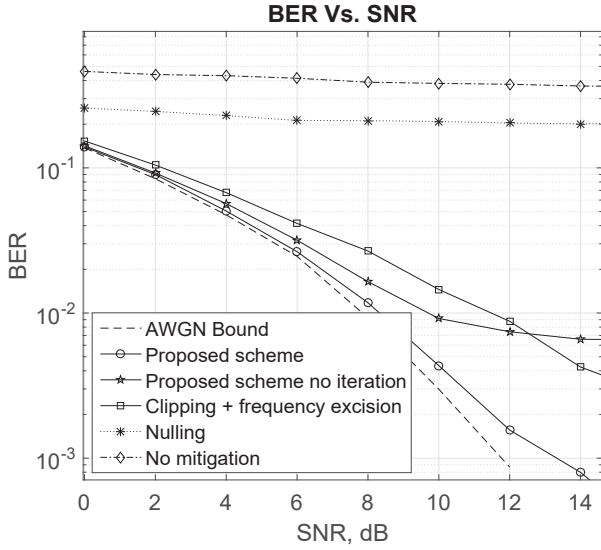


Fig. 3. BER performance in scenario I, INR = 40 dB and SIR = -40 dB.

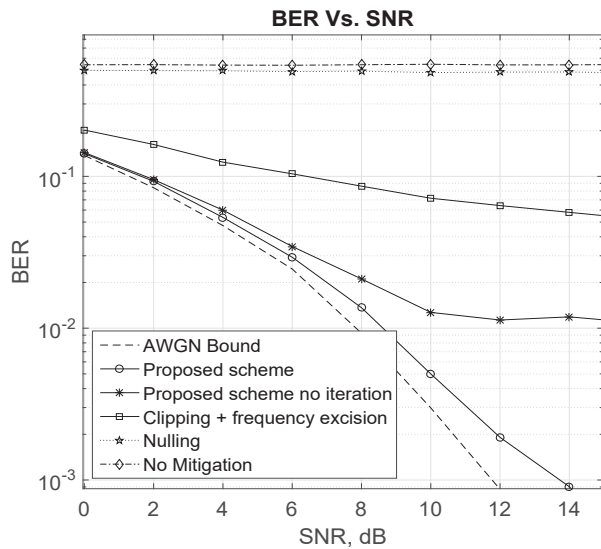


Fig. 4. BER performance in scenario II, INR = 40 dB and SIR = -40 dB.

The BER performance of the algorithm in scenario I and scenario II are shown in Fig. 3 and Fig. 4 respectively. The AWGN bound in the simulation results, defines the scenario when there is no occurrence of both the NBI and IN. As shown, the proposed scheme has performance close to the AWGN bound and is superior than nulling and clipping with frequency excision schemes. Apart from this, the performance of the non-iterative implementation of the algorithm is also shown. As anticipated, the non-iterative approach reaches the saturation after SNR of 10 dB in both cases. Conversely, the iterative approach converges towards a reasonable SNR value in both scenarios.