Frequency-Shift Filtering for OFDM Recovery in Narrowband Power Line Communications

Nir Shlezinger and Ron Dabora
Department of Electrical and Computer Engineering, Ben-Gurion University, Israel
Email: nirshl@post.bgu.ac.il, ron@ee.bgu.ac.il

***Abstract*—Power line communications (PLC) has been drawing considerable interest in recent years due to the growing interest in smart grid implementation. In smart grids, network control and grid applications are allocated the frequency band of $0-500$ kHz, commonly referred to as the narrowband PLC channel. This channel is characterized by strong periodic noise and low signal to noise ratio (SNR). In this work we propose a receiver which uses frequency shift filtering to exploit the cyclostationary properties of *both* the narrowband PLC noise, as well as the information signal, digitally modulated using orthogonal frequency division multiplexing. The results show that the new receiver obtains a substantial performance gain over previously proposed receivers, without requiring any coordination with the transmitter.**

I. Introduction

In recent years the power supply network is changing its role - from a network used solely for energy distribution into a dual-purpose network, which simultaneously supports both communications as well as power distribution. In this paper we focus on power line communications (PLC) which utilizes the frequency band of $0 - 500$ kHz. This is referred to as *narrowband PLC* [1], and is used for applications of automation and control, including power management, smart homes, and automatic meter reading systems.

The statistical properties of the PLC noise are very different from the conventionally used additive white Gaussian noise (AWGN) model [1], [2]. As detailed in [3], [4], power line noise can be modeled as a superposition of several noise components. For narrowband PLC, the dominant noise components are colored background noise, narrowband periodic noise, and periodic impulsive noise synchronous with the AC frequency [5], [6]. Due to the relatively long symbol duration in narrowband PLC transmission, the periodic properties of the noise cannot be ignored, and the noise is modeled as a cyclostationary process. Two cyclostationary PLC noise models exist in literature. The first was proposed by Katayama et al. in [5], and the second was recently proposed in [7]. Both works model the noise as an additive cyclostationary Gaussian noise (ACGN) with a period of half the mains period.

Recent narrowband PLC standardization efforts [8], [9], [10] have adopted the orthogonal frequency division multiplexing (OFDM) modulation scheme. Time-domain OFDM signals are also cyclostationary [11], with a period equal to a single OFDM symbol duration. Thus, in narrowband PLC, both the information signal and the noise are cyclostationary.

In the present paper we propose a new receiver algorithm, based on the time-averaged mean squared error (TA-MSE) criterion, for recovery of OFDM signals received over the narrowband PLC channel. The receiver uses a frequency shift (FRESH) filter for exploiting the cyclostationary properties of the received signal. Specifically, we present the *first* receiver designed for PLC which takes advantage of the cyclostationary properties of *both* the *noise*, and the *information signal*. The novel idea is to utilize the cyclostationary properties of the noise to achieve noise reduction. We also show that the method is beneficial irrespective of the particular model of the cyclostationary noise, as long as cyclostationarity is maintained. The rest of this paper is organized as follows: in Section II we briefly recall the relevant aspects of cyclostationarity to be used in this work, and review the FRESH filter. In Section III, the novel receiver algorithm is developed and its theoretical performance characteristics are obtained. In Section IV simulation results are presented together with a discussion. Lastly, conclusions are provided in Section V.

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II. Preliminaries

In the following we denote vectors with lower-case boldface letters, e.g., $\mathbf{x}, \mathbf{y}$; the i-th element of a vector $\mathbf{x}$ is denoted with $(\mathbf{x})_i$. Matrices are denoted with upper-case boldface letters, e.g., $\mathbf{X}, \mathbf{Y}$; the element at the i-th row and the j-th column of a matrix $\mathbf{X}$ is denoted with $(\mathbf{X})_{i,j}$. $(\cdot)^H$ denotes the Hermitian conjugate, $(\cdot)^T$ denotes the transpose, and $(\cdot)^*$ denotes the complex conjugate. $\delta[\cdot]$ denotes the Kronecker delta function, $\mathbb{E}\{\cdot\}$ denotes the stochastic expectation, and $\langle\cdot\rangle$ denotes the time-average operator.

A. Cyclostationary Signals

A complex-valued discrete-time process $x[n]$ is said to be *wide sense second order cyclostationary* (referred to henceforth as cyclostationary) if there exists an integer N_0, such that $\mathbb{E}\{x[n]\} = \mathbb{E}\{x[n+N_0]\}$, and $c_{xx}(n,l) = \mathbb{E}\{x[n+l]x^*[n]\} = c_{xx}(n+N_0,l)$. As $c_{xx}(n,l)$ is periodic in the variable n, it has a Fourier series expansion, whose coefficients, referred to as *cyclic autocorrelation function*, are $c_{xx}^{\alpha_k}(l) = \frac{1}{N_0}\sum\limits_{n=0}^{N_0-1} c_{xx}(n,l)e^{-j2\pi\alpha_k n}$, where $\alpha_k = \frac{k}{N_0}, k = 0,1,...,N_0-1$, are referred to as the *cyclic frequencies*.

B. Frequency Shift Filtering

The FRESH filter consists of a linear time-invariant (LTI) filter-bank applied to frequency shifted versions of the input signal [12], [13]. Consider the received signal given by $r[n] = d[n] + w[n]$, where $d[n]$ denotes the desired (cyclostationary) signal and $w[n]$ denotes the additive noise. Let each LTI filter in the implementation of the FRESH filter consist of a finite impulse response (FIR) filter with L_{FIR} taps. Let K denote the number of cyclic frequencies used by the FRESH filter, α_k denote the k-th cyclic frequency, $h_k[i]$ denote the i-th coefficient of the k-th FIR filter, and let $\mathbf{z}[n]$ denote the frequency shifted input vector at time n, defined as:

$$\mathbf{z}[n] = \left[\mathbf{r}_0[n], \mathbf{r}_1[n], ..., \mathbf{r}_{K-1}[n]\right]^T, \tag{1}$$

with $(\mathbf{r}_k[n])_i = r[n-i]e^{-j2\pi\alpha_k(n-i)}$, $i \in \mathcal{L} \triangleq \{0,1, \ldots, L_{\text{FIR}}-1\}$. Lastly, we define $\mathbf{h} = [\mathbf{h}_0, \mathbf{h}_1, \ldots, \mathbf{h}_{K-1}]^T$, where $\mathbf{h}_k = [h_k[0], h_k[1], ..., h_k[L_{FIR}-1]]$. The input-output relationship of the FRESH filter can now be written as $y[n] = \mathbf{h}^H \mathbf{z}[n]$. As detailed in [12], [14], the *minimum TA-MSE* FRESH filter for recovering $d[n]$ is given by

$$\mathbf{h} = \bar{\mathbf{C}}_{\mathbf{zz}}^{-1}\bar{\mathbf{c}}_{\mathbf{z}d}, \tag{2}$$

where N_0 is the period of the autocorrelation function of $d[n]$, N_A is an integer multiple of N_0, $\bar{\mathbf{c}}_{\mathbf{z}d} \triangleq \langle \mathbf{c}_{\mathbf{z}d}[n] \rangle = \frac{1}{N_A}\sum_{n=0}^{N_A-1} \mathbb{E}\{\mathbf{z}[n]d^*[n]\}$, $\bar{\mathbf{C}}_{\mathbf{zz}} \triangleq \left\langle \mathbb{E}\{\mathbf{z}[n]\mathbf{z}^H[n]\}\right\rangle$, and L_{FIR} is greater than the largest value of $l \in \{0,1,\ldots,N_0-1\}$ for which there exists an index n such that $c_{dd}(n,l) \neq 0$.

As in [15], we use the independence of the desired signal and the noise, together with the fact that $\mathbb{E}\{w[n]\} = 0$, to write the i-th element of $\mathbf{c}_{\mathbf{z}d}[n]$ as $(\mathbf{c}_{\mathbf{z}d}[n])_i = c_{dd}(n,-q_i)e^{-j2\pi\alpha_{p_i}(n-q_i)}$, where $i \in \mathcal{M} \triangleq \{0,1,\ldots,KL_{\text{FIR}}-1\}$, and $q_i \in \mathcal{L}$, $p_i \in \mathcal{K} \triangleq \{0,1,\ldots,K-1\}$, are such that $i = p_i L_{\text{FIR}} + q_i$. Lastly we have $(\bar{\mathbf{c}}_{\mathbf{z}d})_i = \langle (\mathbf{c}_{\mathbf{z}d}[n])_i \rangle$. Next, consider $\mathbf{C}_{\mathbf{zz}}[n]$: Writing the indexes $u, v \in \mathcal{M}$ as $u = p_u L_{\text{FIR}} + q_u$ and $v = p_v L_{\text{FIR}} + q_v$, $p_u, p_v \in \mathcal{K}$, $q_u, q_v \in \mathcal{L}$, and defining $\gamma_{u,v}(n) \triangleq \alpha_{p_u}(n-q_u) - \alpha_{p_v}(n-q_v)$, we write

$$\begin{aligned}(\mathbf{C}_{\mathbf{zz}}[n])_{u,v} &= \mathbb{E}\left\{r[n-q_u]r^*[n-q_v]e^{-j2\pi\gamma_{u,v}(n)}\right\} \\ &= c_{dd}(n-q_v, q_v-q_u)e^{-j2\pi\gamma_{u,v}(n)} \\ &\quad + c_{ww}(n-q_v, q_v-q_u)e^{-j2\pi\gamma_{u,v}(n)}, \end{aligned} \tag{3}$$

and $\left(\bar{\mathbf{C}}_{\mathbf{zz}}\right)_{u,v} = \langle (\mathbf{C}_{\mathbf{zz}}[n])_{u,v} \rangle$. Since the output signal produced by the FRESH filter is orthogonal to the error [16], the TA-MSE between the output and the desired signal is (see [16, Pg. 431]) $\langle \mathbb{E}\{|y[n]-d[n]|^2\}\rangle = P_d - \bar{\mathbf{c}}_{\mathbf{z}d}^H \bar{\mathbf{C}}_{\mathbf{zz}}^{-1}\bar{\mathbf{c}}_{\mathbf{z}d}$, where $P_d = \langle \mathbb{E}\{d[n]d^*[n]\}\rangle = \langle c_{dd}(n,0)\rangle$.

III. Minimum TA-MSE Signal Recovery

In this section we present a new receiver scheme for the recovery of an OFDM signal received over an additive cyclostationary noise channel. The scheme exploits the spectral correlation of the OFDM signal $d[n]$ *as well as the spectral correlation of the noise*. The received signal is given by $r[n] = d[n] + w[n]$, where $w[n]$ is the noise, and $d[n]$ and $w[n]$ are mutually independent. N_{sym} and N_{noise} denote the periods of $d[n]$ and $w[n]$, respectively. All time averages are over the least common multiple of N_{sym} and N_{noise}. While we derive the optimal receiver for channels without inter-symbol interference (ISI), the same structure can be applied also to ISI channels: Since LTI systems preserve cyclostationarity [17], the receiver can be designed to recover the signal component with ISI from the received signal. The ISI is afterwards inherently handled by the OFDM signal detection process. We note that the same performance gain reported for the no-ISI channel is obtained for channels with ISI.

A. A New Receiver Algorithm: Signal Recovery with Noise Estimation and Cancellation

Our new receiver algorithm applies noise estimation and cancellation prior to signal extraction. The algorithm processing, depicted in Fig. 1, consists of two FRESH filters in series: the first filter, $h_1[n]$, is a noise estimation FRESH filter tuned to extracting the cyclostationary noise. The estimated noise is then subtracted from the received signal, and then a signal extraction FRESH filter, $h_2[n]$, tuned to recovering the OFDM signal is applied.

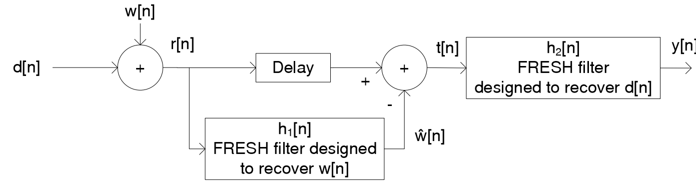

Fig. 1. FRESH filtering for signal recovery with noise extraction and cancellation.

Repeating the derivation in Section II-B with $L_{\text{FIR}1}$, K_1, $\mathcal{L}_1$, $\mathcal{K}_1$ and $\mathcal{M}_1$ instead of L_{FIR}, K, $\mathcal{L}$, $\mathcal{K}$, and $\mathcal{M}$, respectively, it follows from (2) that the FRESH filter $h_1[n]$, designed to recover the noise, is obtained as $\mathbf{h}_1 = \bar{\mathbf{C}}_{\mathbf{zz}}^{-1}\bar{\mathbf{c}}_{\mathbf{z}w}$, where $\mathbf{z}[n]$ is defined in (1) with $\alpha_k = \frac{k}{N_{noise}}$, $\bar{\mathbf{C}}_{\mathbf{zz}} = \langle \mathbf{C}_{\mathbf{zz}}[n]\rangle$, $\mathbf{C}_{\mathbf{zz}}[n]$ is given by (3), and $\bar{\mathbf{c}}_{\mathbf{z}w} = \langle \mathbf{c}_{\mathbf{z}w}[n]\rangle$, with $\mathbf{c}_{\mathbf{z}w}[n] = \mathbb{E}\{\mathbf{z}[n]w^*[n]\}$. By writing the index $i \in \mathcal{M}_1$ as $i = p_i L_{\text{FIR}1} + q_i$, $q_i \in \mathcal{L}_1$, $p_i \in \mathcal{K}_1$, we have

$$(\mathbf{c}_{\mathbf{z}w}[n])_i = c_{ww}(n,-q_i)\, e^{-j2\pi\alpha_{p_i}(n-q_i)}, \tag{4}$$

where $c_{ww}(n,l)$ is specified by the noise model, e.g., [5] or [7]. The estimated noise is $\hat{w}[n] = \mathbf{h}_1^H \mathbf{z}[n]$.

Consider next the FRESH filter $h_2[n]$, designed to recover the OFDM signal $d[n]$. Define for $h_2[n]$ $L_{\text{FIR}2}$, K_2, $\mathcal{L}_2$, $\mathcal{K}_2$ and $\mathcal{M}_2$ in a parallel manner to $L_{\text{FIR}1}$, K_1, $\mathcal{L}_1$, $\mathcal{K}_1$ and $\mathcal{M}_1$ defined for $h_1[n]$. The input signal to $h_2[n]$ is $t[n] = r[n] - \hat{w}[n] = d[n] + w[n] - \hat{w}[n]$. From (2), $h_2[n]$ is obtained as $\mathbf{h}_2 = \bar{\mathbf{C}}_{\mathbf{tt}}^{-1}\bar{\mathbf{c}}_{\mathbf{t}d}$, where $\bar{\mathbf{c}}_{\mathbf{t}d} = \langle \mathbf{c}_{\mathbf{t}d}[n]\rangle$, $\bar{\mathbf{C}}_{\mathbf{tt}} = \langle \mathbf{C}_{\mathbf{tt}}[n]\rangle$, $\mathbf{t}[n] = [\mathbf{t}_0[n], \mathbf{t}_1[n], ..., \mathbf{t}_{K_2-1}[n]]^T$. $(\mathbf{t}_k[n])_i = t[n-i]e^{-j2\pi\beta_k(n-i)}$, $i \in \mathcal{L}_2$, $\beta_k = \frac{k}{N_{sym}}$ denotes the k-th cyclic frequency used in $h_2[n]$. Let the indexes $u, v \in \mathcal{M}_2$ be written as $u = p_u L_{\text{FIR}2} + q_u$ and $v = p_v L_{\text{FIR}2} + q_v$, $q_u, q_v \in \mathcal{L}_2$, $p_u, p_v \in \mathcal{K}_2$, and define $\gamma_{u,v}^{(\beta)}(n) \triangleq \beta_{p_u}(n-q_u) - \beta_{p_v}(n-q_v)$. Then, we have

$$\begin{aligned}(\mathbf{C}_{\mathbf{tt}}[n])_{u,v} &= \mathbb{E}\left\{t[n-q_u]t^*[n-q_v]e^{-j2\pi\gamma_{u,v}^{(\beta)}(n)}\right\} \\ &= c_{tt}(n-q_v, q_v-q_u)\, e^{j2\pi\gamma_{u,v}^{(\beta)}(n)}. \end{aligned} \tag{5}$$

Next, we define $\mathbf{d}[n] \triangleq [\mathbf{d}_0[n], \mathbf{d}_1[n], \ldots, \mathbf{d}_{K_1-1}[n]]^T$, where $(\mathbf{d}_k[n])_i = d[n-i]e^{-j2\pi\alpha_k(n-i)}$, $i \in \mathcal{L}_1$, and $\mathbf{w}[n] \triangleq [\mathbf{w}_0[n], \mathbf{w}_1[n], \ldots, \mathbf{w}_{K_1-1}[n]]^T$, where $(\mathbf{w}_k[n])_i = w[n-i]e^{-j2\pi\alpha_k(n-i)}$, $i \in \mathcal{L}_1$. $\mathbf{z}[n]$ can now be written as $\mathbf{z}[n] = \mathbf{d}[n] + \mathbf{w}[n]$. We denote by $d_1[n] = \mathbf{h}_1^H\mathbf{d}[n]$ the desired signal component at the output of $h_1[n]$, and by $w_1[n] = \mathbf{h}_1^H\mathbf{w}[n]$ the noise component at the output of $h_1[n]$. $t[n]$ may therefore be expressed as $t[n] = d[n] + w[n] - d_1[n] - w_1[n] = d_2[n] + w_2[n]$, where $d_2[n] = d[n] - d_1[n]$ and $w_2[n] = w[n] - w_1[n]$. Since $d[n]$ and $w[n]$ are mutually independent, then $d_2[n]$ and $w_2[n]$ are mutually independent. $c_{tt}(n,l)$ may be therefore obtained by:

$$c_{tt}(n,l) = c_{d_2d_2}(n,l) + c_{w_2w_2}(n,l). \tag{6}$$

As the noise models of both [5] and [7] include a stationary component, $c_{ww}^0(l) \neq 0$, it follows from [14], that $h_1[n]$ must include the cyclic frequency $\alpha_{k_0} = 0$, $k_0 \in \mathcal{K}_1$. Let

$\mathbf{i}_{L_{\mathrm{FIR1}}\cdot k_0}$ denote a column vector such that its i-th coordinate is obtained by $\left(\mathbf{i}_{L_{\mathrm{FIR1}}\cdot k_0}\right)_i = \delta[i - L_{\mathrm{FIR1}} \cdot k_0]$. We may now write $d[n] = \mathbf{i}^H_{L_{\mathrm{FIR1}}\cdot k_0}\mathbf{d}[n]$ and $w[n] = \mathbf{i}^H_{L_{\mathrm{FIR1}}\cdot k_0}\mathbf{w}[n]$. Next, we write $d_2[n] = d[n] - \mathbf{h}_1^H\mathbf{d}[n] = \check{\mathbf{i}}_1^H\mathbf{d}[n]$, where $\check{\mathbf{i}}_1 \triangleq \mathbf{i}_{L_{\mathrm{FIR1}}\cdot k_0} - \mathbf{h}_1$. Letting $\mathbf{C}_{\mathbf{dd}}(n,l) = \mathbb{E}\left\{\mathbf{d}[n+l]\mathbf{d}^H[n]\right\}$, the autocorrelation of $d_2[n]$ may be therefore expressed as

$$c_{d_2d_2}(n,l) = \check{\mathbf{i}}_1^H\mathbf{C}_{\mathbf{dd}}(n,l)\check{\mathbf{i}}_1. \quad (7)$$

By writing the indexes $u_1, v_1 \in \mathcal{M}_1$ as $u_1 = p_{u_1}L_{\mathrm{FIR1}} + q_{u_1}$ and $v_1 = p_{v_1}L_{\mathrm{FIR1}} + q_{v_1}$, where $p_{u_1}, p_{v_1} \in \mathcal{K}_1$, and $q_{u_1}, q_{v_1} \in \mathcal{L}_1$, and defining $\gamma^{(\alpha)}_{u_1,v_1}(n,l) \triangleq \alpha_{p_{u_1}}(n+l-q_{u_1}) - \alpha_{p_{v_1}}(n - q_{v_1})$, we write $\left(\mathbf{C}_{\mathbf{dd}}(n,l)\right)_{u_1,v_1} = \mathbb{E}\{d[n+l-q_{u_1}]d^*[n - q_{v_1}]e^{-j2\pi\gamma^{(\alpha)}_{u_1,v_1}(n,l)}\}$. Therefore $\left(\mathbf{C}_{\mathbf{dd}}(n,l)\right)_{u_1,v_1} = c_{dd}(n - q_{v_1}, q_{v_1} + l - q_{u_1})e^{-j2\pi\gamma^{(\alpha)}_{u_1,v_1}(n,l)}$. Applying the steps used in the derivation of $c_{d_2d_2}(n,l)$ to the derivation of $c_{w_2w_2}(n,l)$, we obtain

$$c_{w_2w_2}(n,l) = \check{\mathbf{i}}_1^H\mathbf{C}_{\mathbf{ww}}(n,l)\check{\mathbf{i}}_1, \quad (8)$$

where $\left(\mathbf{C}_{\mathbf{ww}}(n,l)\right)_{u_1,v_1} = c_{ww}(n - q_{v_1}, q_{v_1} + l - q_{u_1}) \times e^{-j2\pi\gamma^{(\alpha)}_{u_1,v_1}(n,l)}$. The correlation (5) is obtained by plugging (7) and (8) into (6), and plugging (6) into (5).

Next, we compute $\mathbf{c}_{\mathbf{t}d}[n] = \mathbb{E}\left\{\mathbf{t}[n]d^*[n]\right\}$. We note that $\mathbb{E}\{t[n+l]d^*[n]\} = \mathbb{E}\{(d_2[n+l] + w_2[n+l])\, d^*[n]\} = c_{d_2d}(n,l)$. Therefore, by writing the index $i \in \mathcal{M}_2$ as $i = p_iL_{\mathrm{FIR2}} + q_i$, $p_i \in \mathcal{K}_2$, $q_i \in \mathcal{L}_2$, we obtain

$$\left(\mathbf{c}_{\mathbf{t}d}[n]\right)_i = c_{d_2d}(n, -q_i)\, e^{-j2\pi\beta_{p_i}(n-q_i)}. \quad (9)$$

Note that $c_{d_2d}(n,l) = \check{\mathbf{i}}_1^H\mathbb{E}\left\{\mathbf{d}[n+l]d^*[n]\right\}$. By writing the index $u \in \mathcal{M}_1$ as $u = p_uL_{\mathrm{FIR1}} + q_u$, $p_u \in \mathcal{K}_1$, $q_u \in \mathcal{L}_1$, we have $\left(\mathbb{E}\left\{\mathbf{d}[n+l]d^*[n]\right\}\right)_u = \mathbb{E}\left\{d[n+l-q_u]e^{-j2\pi\alpha_{p_u}(n+l-q_u)}d^*[n]\right\} = c_{dd}(n, l - q_u)e^{-j2\pi\alpha_{p_u}(n+l-q_u)}$. Equations (3) and (4) provide a closed form expression for $h_1[n]$, and equations (5) and (9) provide a closed form expression for $h_2[n]$. The TA-MSE of this receiver is given by: TA-MSE $= P_d - \bar{\mathbf{c}}^H_{\mathbf{t}d}\bar{\mathbf{C}}^{-1}_{\mathbf{tt}}\bar{\mathbf{c}}_{\mathbf{t}d}$.

B. Best of Previous Work: A FRESH Filter Designed in [15] for Direct Signal Recovery

The best previously proposed scheme for this model is a FRESH filter tuned to extracting the OFDM signal based on the minimum TA-MSE criterion, proposed in [15]. The filter $h[n]$ is derived as detailed in Subsection II-B.

IV. Simulations

In this section, the performance of the receiver developed in Section III is evaluated by simulations. The information bits are encoded in accordance with the IEEE P1901.2 standard [8]: an outer Reed Solomon $(255, 239)$ code is followed by an inner rate $\frac{1}{2}$ convolutional code with generator polynomials 171_{octal} and 155_{octal} and an interleaver. The information signal is a passband OFDM signal with 32 subcarriers over the frequency band $3 - 148.5$ kHz, each modulated with QPSK constellation. This simulated frequency range is in accordance with the European CENELEC regulations [19]. We use a cyclic prefix consisting of 16 samples, hence the total number of samples at each OFDM symbol is 80.

Three types of noise were simulated - (1) AWGN (in order to show robustness to the noise model); (2) ACGN based on the Katayama model [5], with two sets of the typical parameters taken from [5, Sec VI.] and [18, Res. 1], referred to herein as KATA1 and KATA2, respectively; (3) ACGN based on the model proposed in [7] and adopted by the IEEE P1901.2 standard [8], with two sets of the typical parameters taken from [8, LV8 site] and [8, LV14 site], referred to herein as IEEE1 and IEEE2, respectively. The cyclic period of the cyclostationary noise is set to $N_{noise} = 1000$ samples. Note that the noise period, N_{noise}, and the length of the OFDM signal symbol, N_{sym}, are both scaled by a factor of $\frac{1}{2.5}$ compared to their practical values to reduce simulation time. However, as $\frac{N_{sym}}{N_{noise}}$ is the same as in practical systems, the results correspond to the performance of practical systems.

Four receivers were simulated - (1) Rx_1: A receiver with *no filtering* applied to the input signal prior to decoding; (2) Rx_2: A receiver which implements a *stationary FIR Wiener filter* [20, Ch. 12.7] with $N_{sym} + \frac{N_{noise}}{2}$ taps applied to the input signal $r[n]$; (3) Rx_3: *Best of previous work* is represented by a receiver with a FRESH filter tuned to extract the desired OFDM signal [15]. The filter utilizes 5 frequency shifts in the range $\frac{-2}{N_{sym}}, \ldots, \frac{2}{N_{sym}}$ such that each FIR has $N_{sym} + \frac{N_{noise}}{2}$ taps; (4) Rx_4: *Our newly proposed algorithm* is demonstrated by a receiver with the FRESH filter $h_1[n]$ utilizing 5 frequency shifts in the range $\frac{-2}{N_{noise}}, \ldots, \frac{2}{N_{noise}}$, and at each branch the FIR has $\frac{N_{noise}}{2}$ taps, and the FRESH filter $h_2[n]$ utilizing 5 frequency shifts in the range $\frac{-2}{N_{sym}}, \ldots, \frac{2}{N_{sym}}$, and at each branch the FIR has N_{sym} taps.

Note that all four receivers have the same delay, and that the new receiver (Rx_4) and the scheme of [15] (Rx_3) have the same number of taps. We also note that the stationary Wiener filter (Rx_2) has less coefficients than our filter but it has the same delay. Increasing the number of taps in Rx_2 increases the delay but does not improve the performance of Rx_2 in the simulations. The results are plotted for various values of input SNR defined as $\mathrm{SNR}_{in} \triangleq \frac{P_d}{\langle\mathbb{E}\{w[n]w^*[n]\}\rangle}$. For evaluating the bit error rate (BER) performance, a per-subcarrier maximum likelihood (ML) decoder is used.

In the following the TA-MSE and robustness to the noise model are studied, as well as coded BER performance and the corresponding SNR gain.

1) Evaluating TA-MSE Performance and Verifying Robustness of the New Algorithm to the Noise Model: First we verified that the new receiver operates well also in AWGN: The simulation results are presented in Fig. 2. Observe that when the new receiver Rx_4 is applied to AWGN it achieves the same TA-MSE as the FRESH filter without noise cancellation of [15] (Rx_3), which is tuned to recover the OFDM signal. This is because in AWGN there is no cyclic redundancy that can be used for noise cancellation. Both Rx_4 and Rx_3 achieve 0.8 dB gain over the stationary Wiener filter (Rx_2) for $\mathrm{SNR}_{in} \leq 2$ dB, the gain decreases to 0.55 dB at $\mathrm{SNR}_{in} = 6$ dB.

Next, the TA-MSE was evaluated for four sets of ACGN models: KATA1 and KATA2 for the Katayama model [5], depicted in Fig. 3, and IEEE1 and IEEE2 for the model of [7], depicted in Fig. 4. Observe that the performance improvement depends on the cyclostationary characteristics of the noise: When the impulsive noise is of typical width of $300 - 400$ microseconds as in KATA2, IEEE1 and IEEE2, the noise has a

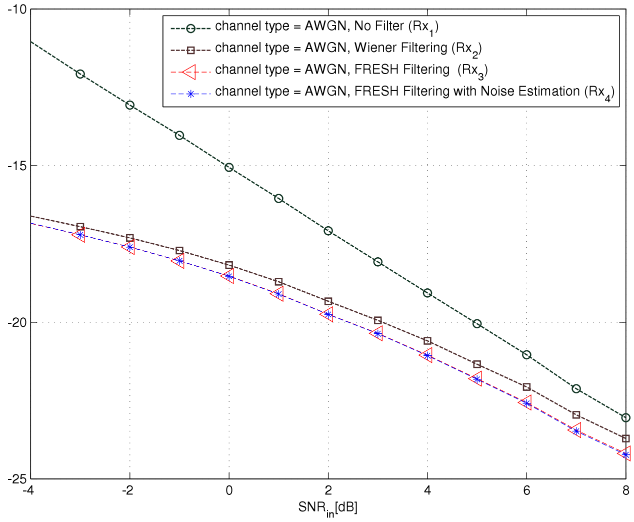

Fig. 2. TA-MSE comparison for an AWGN channel.

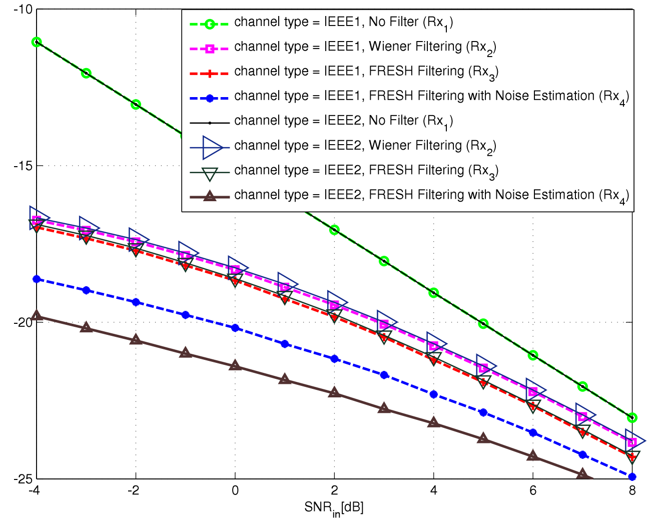

Fig. 4. TA-MSE comparison for the noise model of [7].

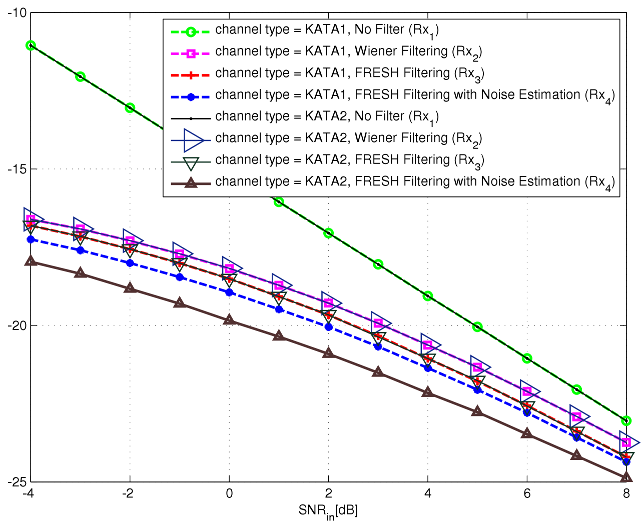

Fig. 3. TA-MSE comparison for the Katayama noise model of [5].

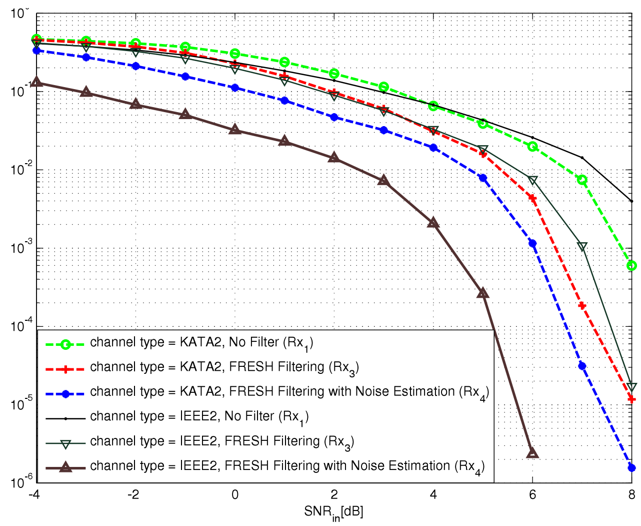

Fig. 5. BER comparison for cyclostationary channel.

stronger cyclic redundancy and therefore noise cancellation is more effective. Accordingly, for the KATA2 model we observe a TA-MSE gain compared to Rx_3 of 2.4 dB at $\mathrm{SNR}_{in} \leq 0$ dB, which decreases to 1.2 dB at $\mathrm{SNR}_{in} = 4$ dB. For the IEEE models we observe in Fig. 4 gains of $2.5 - 6$ dB compared to Rx_3 at low SNR_{in} which decreases at $\mathrm{SNR}_{in} = 4$ dB to 1.55 dB gain for IEEE1 and 2.7 dB gain for IEEE2. However, when the impulsive noise component is very short, as in KATA1 (only 25 microseconds impulse width) Rx_4 achieves relatively modest gains in the TA-MSE of about $0.35 - 1.2$ dB compared Rx_3. In all cases, as Rx_3 exploits only the cyclostationary characteristics of the OFDM signal, its performance improvement over the stationary Wiener filter (Rx_2) is the same for both noise models at all SNRs. *The benefits of noise cancellation are thus clearly observed.* We note that all TA-MSE results were confirmed by the theoretical analysis presented in Section III.

2) BER Improvements due to Noise Cancellation: The substantial gains obtained by Rx_4 in terms of TA-MSE translate directly into gain in BER. To demonstrate this point the coded BER results at the output of the different receivers for the ACGN channel are depicted in Fig. 5 for both noise models. To avoid cluttering we depict only the results with the KATA2 and the IEEE2 parameters. Observe that the new FRESH receiver with noise cancellation (Rx_4) achieves an input SNR gain of 4.9 dB compared to Rx_3 at output BER of 10^{-1} for the IEEE2 model. This gain decreases to 3.2 dB at output BER of 10^{-2} and to 2.6 dB at output BER of 10^{-3}. For the KATA2 model the corresponding input SNR gain of Rx_4 over Rx_3 are 1.6 dB, 0.65 dB and 0.4 dB, respectively. It is emphasized that this BER improvement is *only* due to noise cancellation.

V. Conclusions

In this paper, a new receiver designed for exploiting the cyclostationary characteristics of the OFDM information signal as well as *those of the narrowband PLC channel noise* is proposed. The novel aspect of the work is the insight that noise estimation in cyclostationary noise channels is *beneficial*, contrary to the widely used AWGN channels. It was shown that a substantial performance improvement can be obtained by FRESH filtering combined with noise cancellation, compared to previous approaches which focused on estimating only the signal. This gain was demonstrated for different cyclostationary noise models and in particular for the noise models specified in the IEEE standard [8].

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