



Channel Estimation and Carrier Frequency Offset Compensation in Orthogonal Frequency Division Multiplexing System Using Adaptive Filters in Wavelet Transform Domain

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ABSTRACT

In this paper, a combination of channel, receiver frequency-dependent IQ imbalance, and carrier frequency offset estimation under short cyclic prefix length is considered in orthogonal frequency division multiplexing system. An adaptive algorithm based on the set-membership filtering algorithm is used to compensate for these impairments. In short CP length, per-tone equalization structure is used to avoid inter-symbol interference. Due to CFO impairment and IQ imbalance in the receiver, we will expand the PTEQ structure to a two-branch structure. This structure has high computational complexity, so using the set-membership filtering idea with variable step size while reducing the average computation of the system can also increase the convergence speed of the estimates. On the other hand, applying wavelet transform on each branch of this structure before applying adaptive filters will increase the estimation speed. The proposed algorithm will be named SMF-WP-NLMS-PTEQ. The results of the simulations show better performance than the usual adaptive algorithms. Besides, estimation and compensation of channel effects, receiver IQ imbalance and carrier frequency offset under short CP can be easily accomplished by this algorithm.

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1. INTRODUCTION

The use of high-performance wireless communication systems is steadily growing [1]. As a multicarrier modulation technique, the orthogonal frequency division multiplexing (OFDM) technique can provide reliable high data rate transmission in different communication scenarios [2, 3].

The tendency to use cutting-edge technology and smaller elements in the same silicon area can also cause problems. For example, replacing the direct conversion architecture (DCA) with a superheterodyne configuration scheme with low cost and lower power utilization in silicon compared to the super heterodyne can lead to an imbalance of Inphase and Quadrature (IQ). IQ imbalance can be described as the difference between the gain, phase,

and even frequency response of each orthogonal branch of communication systems [4, 5].

The OFDM direct conversion receiver also suffers from carrier frequency offset and DC offset in addition to IQ imbalance. On the other hand, the OFDM system is very sensitive to the carrier frequency offset (CFO) [6]. In OFDM systems, the presence of some frequency offset will disrupt the orthogonality of the sub-carriers, resulting in inter-carrier interference and a loss of performance [7, 8].

In [9], an empirical mode decomposition (EMD) based adaptive filter for channel estimation in an OFDM system is proposed. In this method, the length of the channel impulse response (CIR) is first approximated. Then, CIR is estimated using an adaptive filter through the received OFDM symbol.

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In [10], an efficient IQ imbalance compensation scheme is proposed based on pilots; this is achieved through a nonlinear least squares (NLS) analysis of the joint channel and IQ imbalance estimation, and a simple symbol detection procedure. For rigor, the Cramer-Rao lower bounds (CRLBs) for both the IQ imbalance parameters and the channel coefficients are also derived.

In [11], a novel joint channel impulse response estimation and impulsive noise mitigation algorithm based on compressed sensing theory is proposed. In this algorithm, both the channel impulse response and the impulsive noise are treated as a joint sparse vector. Then, the sparse Bayesian learning framework is adopted to jointly estimate the channel impulse response, the impulsive noise, and the data symbols, in which the data symbols are regarded as unknown parameters.

In [12], an adaptive approach to eliminate the receiver frequency IQ imbalance in OFDM systems based on a direct conversion scheme is implemented using decision feedback adaptive filtering.

In [13], the effect of receiver IQ imbalance on the OFDM system is investigated and a design based on training data in both time and frequency domain is presented. In [14], the same method is applied assuming the presence of transmitter IQ. In [15], an integrated structure for estimating transmitter, receiver, and channel IQ under short CP conditions is presented. In this structure, PTEQ structure is used to overcome the inter-symbol-interference (ISI) problem. Short CP increases bandwidth efficiency.

Employing cyclic-prefix (CP) in multi-carrier systems not only protects the signal from inter-symbol-interference but also allows circular interpretations of the channel which simplifies the estimation and equalization techniques. Nevertheless, the CP information is usually discarded at the receiver side [16].

In this paper, short CP is considered for optimal use of the channel. Therefore, PTEQ structure-based algorithms will be used. In addition to effectively estimating the channel, the receiver IQ imbalance and CFO prevent ISI and deterioration of the estimation results. In this paper, to increase the speed of estimation of the adaptive algorithms, in each sub-carrier of the PTEQ output data, a wavelet transform is implemented. Then the NLMS adaptive algorithm is developed in the new output signal. Due to the length of filters required and the number of subcarriers, the implementation of any adaptive algorithm in this structure will require a lot of computation for convergence.

Therefore, computational optimization of the applied algorithms while maintaining the required convergence

speed and steady-state error is necessary. In the second step, to reduce the computation of the channel estimation process based on the adaptive algorithm, we need to increase the convergence speed again. For this purpose, we use set-membership filtering in adaptive filters. The use of SMF filtering can improve the estimation speed alone by using variable step-size, optimal, and dependent on noise variance. Also, in some iteration, it prevents partial sub-carriers from being updated and reduces average computation.

Therefore, the implementation of an adaptive algorithm in the PTEQ structure in combination with wavelet transform and set-membership filtering will have several important features. First, it will increase channel utilization efficiency by reducing the length of CP used. Second, the ISI can be eliminated for the short CP utilization and, ultimately, due to SMF and WP transform using, has high speed and accuracy in channel and receiver IQ imbalance estimation. The CFO effect can also be compensated in the proposed structure by existing methods independent of channel effects and IQ imbalances. Therefore, using SMF-WP-NLMS-PTEQ proposed method, good channel equalization can be performed for the data.

This paper first describes the OFDM-based receiver IQ imbalance and channel estimation model. Then the wavelet packet transform is reviewed in Section 2. The proposed adaptive compensation algorithm is presented using SMF-based adaptive filters in the wavelet transform domain and PTEQ structure under short CP in Section 3. Finally, the simulation results are presented in Section 4 and the conclusions in Section 5.

2. CHANNEL ESTIMATION MODEL IN OFDM SYSTEMS

In this section, the compensation of channel distortion under sufficient CP length will be studied and corresponding analytical equations will be introduced to model the channel effect so that based on these equations the proposed method can be presented in the next sections. In sufficient CP length, inter-symbol-interference does not occur, so a first-order filter as a compensator can minimize existing distortions. But in the case of short CP length, a structure called PTEQ is used to compensate and estimate the data.

2. 1. Sufficient CP Length

First, for a sufficient CP, the channel compensation scheme is examined. It is assumed that S represents the OFDM symbol in the frequency domain with $(N \times 1)$ length, where N is the

number of subcarriers. Thus the baseband symbol in the time domain can be written as Equation (1).

$$s = P_{CI} F_N^{-1} S \quad (1)$$

where P_{CI} is a cyclic prefix (CP) insertion matrix of length v to symbol S , and F_N^{-1} represents the inverse matrix of DFT.

When the output symbol in the transmitter passes through the semi-stationary channel with a length of L_{ch} , the received baseband symbol r can be written as Equation (2).

$$r = c \otimes s + n \quad (2)$$

In Equation (2), c represents the baseband channel model. Also, n is Gaussian additive white noise.

The OFDM system is sensitive to the CFO due to changes in the transmitter and receiver local oscillators. Therefore, its effect must be compensated. By applying the CFO effect on the received signal, Equation (3) is obtained.

$$z = r \cdot e^{j2\pi\Delta f \cdot t} \quad (3)$$

when considering the receiver IQ imbalance as the CFO and the channel effect, the received signal can be written as Equation (4).

$$\begin{aligned} y &= y_{1r} \otimes (r \cdot e^{j2\pi\Delta f \cdot t}) + y_{2r} \otimes (r^* \cdot e^{-j2\pi\Delta f \cdot t}) = \\ &y_{1r} \otimes ((c \otimes s + n) \cdot e^{j2\pi\Delta f \cdot t}) + y_{2r} \otimes ((c \otimes s + n)^* \cdot e^{-j2\pi\Delta f \cdot t}) \end{aligned} \quad (4)$$

where y_{1r} and y_{2r} are the frequency selective imbalance filter model of the receiver in branches I and Q. In the frequency domain, the CFO will result in a leakage of energy from the desired sub-carrier to all sub-carriers in one OFDM symbol. The energy leakage from sub-carrier $[l]$ to other sub-carrier $[l']$ will be in Equation (5) [17].

$$P_\zeta[l'] = e^{j\pi(\zeta-l)\frac{N-1}{N}} \frac{\sin \pi(\zeta-l)}{\sin \frac{\pi}{N}(\zeta-l')} \quad (5)$$

where ζ as normalized CFO, is the actual ratio of CFO namely Δf and the distance between the substrates $\frac{1}{T \cdot N}$ and is obtained as $\zeta = \Delta f \cdot T \cdot N$. and, T is the duration of each bit.

Equation (4) clearly shows that the received signal is first distorted by the channel, then the resulting signal is affected by the CFO, and is eventually re-distorted by the receiver IQ. To compensate for CFO and IQ, the CFO is first estimated. Then the signal itself and its conjugate once are multiplied by $e^{-j2\pi\Delta f \cdot t}$ and the Equations (6) and (7) are obtained.

$$y_1 = y \cdot e^{-j2\pi\Delta f \cdot t} \quad (6)$$

$$y_2 = (y)^* \cdot e^{-j2\pi\Delta f \cdot t} \quad (7)$$

Straightforwardly, we can derive symbol estimation using a suitable linear combination of two inputs in the frequency domain including $Y_1[l]$ and $Y_2^*[l_m]$ as Equation (8).

$$\tilde{S}[l] = [W_a[l] \quad W_b[l]] \begin{bmatrix} Y_1[l] \cdot e^{-j2\pi\Delta f \cdot t} \\ Y_2^*[l_m] \cdot e^{-j2\pi\Delta f \cdot t} \end{bmatrix} \quad (8)$$

Equation (8) shows a two-tap FEQ equalizer that with any adaptive algorithm, the coefficients can be learned. The coefficients $W_a[l]$ and $W_b[l]$ can be calculated using the MSE criterion, which is expressed by Equation (9).

$$\min_{W_a[l], W_b[l]} \mathcal{E} \left\{ \left| \tilde{S}[l] - \begin{bmatrix} W_a[l] \\ W_b[l] \end{bmatrix}^T \begin{bmatrix} Y_1[l] \cdot e^{-j2\pi\Delta f \cdot t} \\ Y_2^*[l_m] \cdot e^{-j2\pi\Delta f \cdot t} \end{bmatrix} \right|^2 \right\} \quad (9)$$

In the above equations, $(.)_m$ represents the mirror operator and is expressed as $Y_m[l] = Y[l_m]$ where:

$$[l_m] = \begin{cases} [N - l + 2], & \text{for } l = 2, \dots, N \\ [l], & \text{for } l = 1 \end{cases} \quad (10)$$

2. 2. Wavelet Packet Transform

Wavelets are transform methods that have received a great deal of attention over the past decades. The wavelet transform is a time-scale representation by which signals are broken down into time and scale functions in terms of basic functions. Wavelet transform has been used extensively for various applications including feature extraction, detection, data compression, signal denoising, etc. [18].

The basis of the wavelet transform is based on two basic functions shown in Equation (11).

$$\begin{aligned} \varphi_{j,k}(t) &= 2^{\frac{j}{2}} \varphi(2^j t - k) \\ \psi_{j,k}(t) &= 2^{\frac{j}{2}} \psi(2^j t - k) \end{aligned} \quad (11)$$

In which they are called the mother wavelet function and the basic scale function, respectively [18].

Using these two basic functions, any arbitrary signal can be written as Equation (12).

$$f(t) = \sum_{k=-\infty}^{+\infty} c_k \varphi(t - k) + \sum_{k=-\infty}^{+\infty} \sum_{j=0}^{+\infty} d_{j,k} \psi(2^j t - k) \quad (12)$$

Most of the results of wavelet theory are developed using filter banks. The full wavelet packet decomposition in two scales is shown in Figure 1.

3. PROPOSED ALGORITHM

In this section, the proposed algorithm for short CP length will be provided. Reducing CP length improves channel utilization. Short CP lengths in multiple channels will

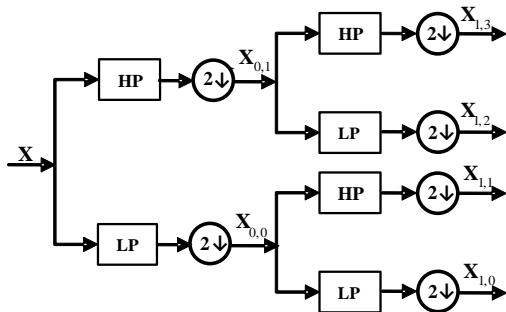


Figure 1. Wavelet packet decomposition in two scales

result in inter-symbol interference. In this paper, to increase the efficiency of channel use, inter-symbol interference is accepted and then using the PTEQ structure in the frequency domain, and simultaneously with an effective estimation of the channel, the inter-symbol interference will be eliminated. The estimation and compensation of the channel effect assuming a short CP length cannot be modeled based on the equations described in the previous section. The PTEQ compensation structure to overcome the ISI is originally derived from the integration of a time-domain equalizer (TEQ) and a frequency-domain equalizer (FEQ) [17]. The PTEQ structure, which can reduce the effective channel length and compensate for the channel effect in the frequency domain, by making adjustments commensurate with that used in this paper, is illustrated in Figure 2.

A PTEQ is a unified compensation structure, where equalization is performed individually on each subcarrier after taking the DFT of the received signal.

In the PTEQ structure, a multi-tap filter is used for each subcarrier as an equalizer. Using this mechanism, the estimation of data in each subcarrier is based on the optimal design of the coefficients of these filters, showing Equations (13) to (15) of the corresponding equations.

$$\tilde{S}^{(i)}[l] = W^{(i)}[l] \left(F_{ext}^{(i)}[l] z \right) \quad (13)$$

where the matrix of $F_{ext}[l]$ is defined as Equation (14) and $L' = L'' + L_r - 1$. Where, L_r and L'' is considered as the length of the receiver imbalance and PTEQ structure branch filters, respectively.

$$F_{ext}[l] = \begin{bmatrix} I_{L'-1} & 0_{L'-1 \times N-L'+1} & -I_{L'-1} \\ 0_{1 \times L'-1} & F_N[l] \end{bmatrix} \quad (14)$$

In this matrix, the first row represents the difference, and the second row the DFT matrix. And again, the MSE criterion according to Equation (15) is used to obtain the PTEQ coefficients.

$$\min_{W_a[l], W_b[l]} \mathbb{E} \left\{ \left| \tilde{S}[l] - \begin{bmatrix} W_a[l] \\ W_b[l] \end{bmatrix}^T \begin{bmatrix} F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta f \cdot t} \\ F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta f \cdot t} \end{bmatrix} \right|^2 \right\} \quad (15)$$

Equation (15) shows the compensator criteria in the two-branch structure of PTEQ, in which the coefficients can be learned by any adaptive algorithm.

The speed and accuracy of channel estimation are essential. Therefore, before the development of adaptive filters, one of the orthogonal transforms in the proposed structure will be applied. Figure 3 shows the block diagram of the receiver section of the proposed SMF-WP-NLMS-PTEQ algorithm, which is described below.

Due to the benefits of wavelet transform and considering that the convergence speed of the LMS algorithm in this domain is faster than the LMS algorithm applied in the time domain, discrete cosine and Fourier transform domain [19], the wavelet transform in the proposed structure will be used. It has been shown previously in reference [20] and the system identification scenario that the use of adaptive filtering in orthogonal transform domain such as wavelet and DCT can increase convergence speed, and reduce steady-state error. This improvement is due to the decrease in the self-correlation of the input signals. Therefore, to increase the speed of the algorithm in channel estimation, and using the wavelet transform denoising property, in the proposed algorithm, the wavelet transform is first applied to the data behind the branches in each subcarrier. The appropriate adaptive algorithm is then extended to each branch of the PTEQ structure.

After applying the wavelet transform on the PTEQ branches and the data $F_{ext}^{(i)}[l]y$, new data will be shown as $(F_{ext}^{(i)}[l]y)_{WP}$.

Figure 3 shows that each subcarrier must be equalized using a multi-tap adaptive filter.

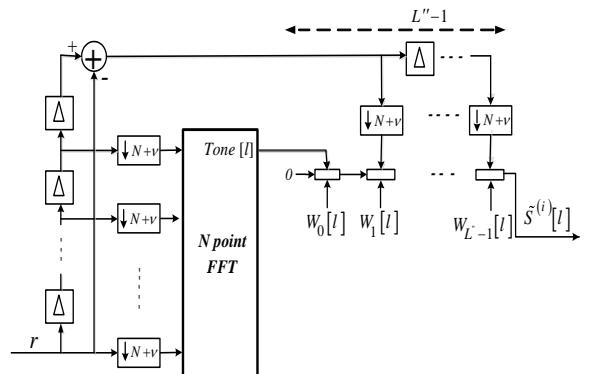


Figure 2. PTEQ-OFDM Receiver Structure

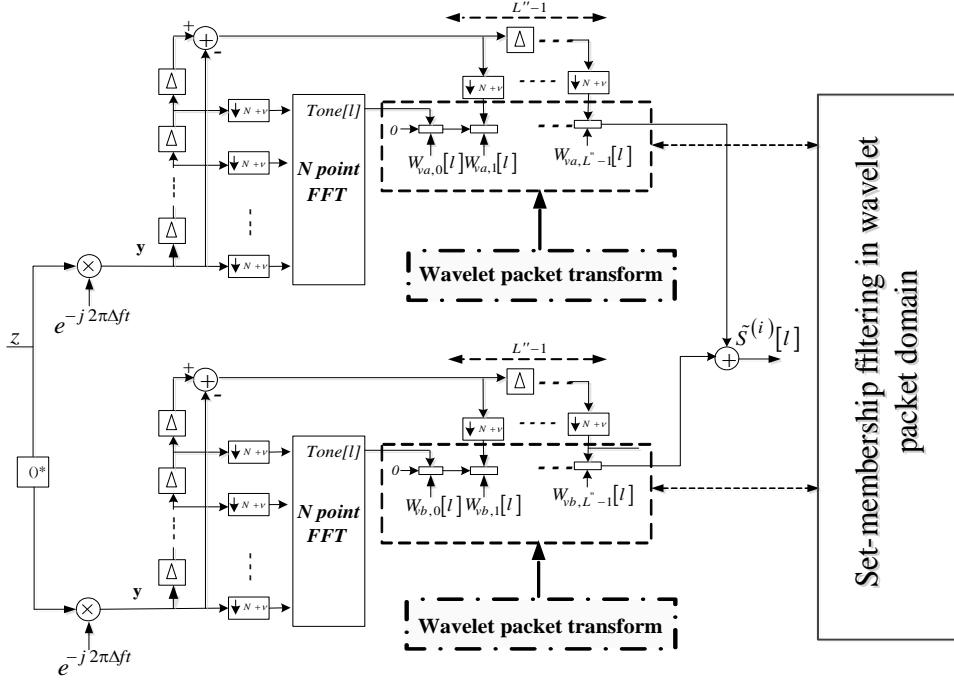


Figure 3. Receiver block diagram of proposed SMF-WP-NLMS-PTEQ algorithm

In this paper, first, the proposed WP-NLMS-PTEQ algorithm is developed for high-speed channel estimation. Then, using set-membership filtering along with wavelet packet analysis is proposed to increase the coefficient training speed and reduce the average computation rate. This algorithm is called SMF-WP-NLMS-PTEQ.

In the SMF method, when the error is less than the predetermined bound value, it is prevented from updating the coefficients and will consequently reduce the amount of computation. On the other hand, the convergence speed is also increased due to the use of the variable and noise dependent step size. The proposed algorithm will consider the estimation problem in the presence of a short CP length.

In the case of short CP, the NLMS-type algorithm is extended to the framework of the PTEQ structure for channel-effect compensation. Here, we obtain the NLMS technique in the PTEQ structure for a given error in the adaptive filter output. In a PTEQ structure, we can derive the output of the adaptive filter as $\tilde{S}^{(i)}[l]$ instant i by using Equation (17), where $W[l] = [W_0[l], W_1[l], \dots, W_{L''-1}[l]]^T$, is defined as an $L'' \times 1$ filter coefficient vector, and $F_{ext}^{(i)}[l]y$, is the $L'' \times 1$ data vector. The NLMS algorithm is computable for each branch in the PTEQ structure and is derived by using the following constrained minimization problem similar to [21]:

$$\min_{W^{(i+1)}[l]} \|W_a^{(i+1)}[l] + W_b^{(i+1)}[l] - W_a^{(i)}[l] - W_b^{(i)}[l]\|_2^2 \quad (16)$$

The limitation of the above criterion is as follows, in which $S[l]$ is a known pilot data.

$$(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}.t})_{WP}^T W_a^{(i+1)}[l] + \\ (F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}.t})_{WP}^T W_b^{(i+1)}[l] = S[l] \quad (17)$$

In Equation (17), $(F_{ext}^{(i)}[l]y)_{WP}$ represents the transform of $(F_{ext}^{(i)}[l]y)$ using the wavelet packet.

The above equations are solved by minimizing the Lagrange coefficients and the cost function of Equation (18):

$$J^{(i)}(l) = \|W_a^{(i+1)}[l] + W_b^{(i+1)}[l] - W_a^{(i)}[l] - \\ W_b^{(i)}[l]\|_2^2 + \lambda(S^{(i)}[l] - \\ (F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}.t})_{WP}^T W_a^{(i+1)}[l] + \\ (F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}.t})_{WP}^T W_b^{(i+1)}[l]) \quad (18)$$

These equations are very complicated. So for simplicity, we assume that in any adaptive filter each branch can estimate the output after the training procedure. Therefore, the above equations can be decomposed into two independent Equations as (19) and (20) to solve the problem in simple form:

$$\min_{W_a^{(i+1)}[l]} \|W_a^{(i+1)}[l] - W_a^{(i)}[l]\|_2^2 \quad (19)$$

$$(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T W_a^{(i+1)}[l] = S[l]$$

$$\min_{W_b^{(i+1)}[l]} \|W_b^{(i+1)}[l] - W_b^{(i)}[l]\|_2^2 \quad (20)$$

$$(F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T W_b^{(i+1)}[l] = S[l]$$

Based on the above decomposed equations and NLMS algorithm, the initial update equations to calculating the filter taps can be written in Equations (21) and (22):

$$W_a^{(i+1)}[l] = W_a^{(i)}[l] + \frac{(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T (W_a^{(i+1)}[l] - W_a^{(i)}[l]) (F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}}{\|(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}\|_2^2} \quad (21)$$

$$W_b^{(i+1)}[l] = W_b^{(i)}[l] + \frac{(F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T (W_{vx}^{(i+1)}[l] - W_{vx}^{(i)}[l]) (F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}}{\|(F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}\|_2^2} \quad (22)$$

On the other hand, we have:

$$(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T W_a^{(i+1)}[l] - (F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T W_a^{(i)}[l] = S[l] - (F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T = e^{(i)}(l) \quad (23)$$

$$(F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T W_{vx}^{(i+1)}[l] - (F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T W_{vx}^{(i)}[l] = S[l] - (F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}^T = e^{(i)}(l) \quad (24)$$

Thus, Equations (21) and (22) can be written as (25) and (26). Also, the step size μ is inserted into the equations to control the stability and convergence rate.

$$W_a^{(i+1)}[l] = W_a^{(i)}[l] + \mu \frac{(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP} e^{(i)}(l)}{\|(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}\|_2^2} \quad (25)$$

$$W_b^{(i+1)}[l] = W_b^{(i)}[l] + \mu \frac{(F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP} e^{(i)}(l)}{\|(F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}\|_2^2} \quad (26)$$

In a PTEQ structure, due to the large number of branches and, as a result, the high number of filter coefficients in the branches, the amount of computation needed to train the system is high. Therefore, it is necessary to reduce the computational complexity of this system. Therefore, the SMF-based algorithm for reducing the average computation and increasing the convergence speed is presented in the following section. Also, in the NLMS algorithm, the step size μ is limited to $0 < \mu < 2$.

3. 1. Proposed SMF-WP-NLMS-PTEQ Algorithm

To obtain the SMF-WP-NLMS-PTEQ algorithm in the PTEQ structure, the step size used in Equations (25) and

(26) are considered as a variable. The update equations of this algorithm are provided in Equations (27) and (28). Equation (29) also shows the relation of the variable step size.

$$W_a^{(i+1)}[l] = W_a^{(i)}[l] + \alpha^{(i)} \frac{(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP} e^{(i)}(l)}{\|(F_{ext}^{(i)}[l] y \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}\|_2^2} \quad (27)$$

$$W_b^{(i+1)}[l] = W_b^{(i)}[l] + \alpha^{(i)} \frac{(F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP} e^{(i)}(l)}{\|(F_{ext}^{(i)}[l] y^* \cdot e^{-j2\pi\Delta\tilde{f}t})_{WP}\|_2^2} \quad (28)$$

$$\alpha^{(i)} = \begin{cases} 1 - \frac{\gamma}{|e^{(i)}[l]|}, & \text{if } \gamma > |e^{(i)}[l]| \\ 0, & \text{otherwise} \end{cases}, \quad \gamma = \sqrt{5\sigma_n^2} \quad (29)$$

In Equation (29), $\alpha^{(i)}$ is a variable step size and γ represents the error threshold value which can be calculated based on of noise variance as $\gamma = \sqrt{5\sigma_n^2}$. In the SMF-NLMS algorithm, an upper threshold γ is assumed to control and constrain the estimation error [22].

In the SMF (Set-membership filtering) method, the coefficients are not updated at times when the error is less than the threshold value, thus reducing the amount of computation, significantly. Also, the algorithm uses a noise-dependent variable step size, which can increase the convergence speed.

4. SIMULATION RESULTS

In this section, the results of several simulations are shown to demonstrate the efficiency of the proposed algorithm. In the simulations, the FFT size is set to 64, the CP length is $v = 5$, and the data modulation is 64QAM and 16QAM. Multipath channel with $(L_{ch} + 1 = 18)$ taps is used, in which the taps are independently selected and have a complex Gaussian distribution. Also, the upper band of SMF-based algorithms is considered to be $\gamma = \sqrt{5\sigma_n^2}$, where σ_n^2 is the noise variance and it is assumed to be known. In curves that do not use the SMF technique, the adaptive filter step size $\mu = 0.05$ is used. In all simulations, the decomposition level of the wavelet transform packet will be 2. All results are simulated in a Rayleigh multipath channel with 18-tap length.

Figure 4 shows a plot of bit error rate (BER) versus signal-to-noise ratio (SNR) for the proposed SMF and WP-based methods for the NLMS-PTEQ algorithm in the PTEQ structure. This simulation result is derived considering both CFO and receiver frequency selective IQ. The modulation used is 16QAM and the results are compared with LS algorithm with sufficient CP and unequal conditions. The proposed SMF-NLMS-PTEQ, WP-NLMS-PTEQ, and SMF-WP-NLMS-PTEQ algorithms, in addition to the LS algorithm, show good

results in terms of BER than the conventional NLMS-PTEQ algorithm. Also in the curves (c), (e), and (f) which used SMF technique, updating was done in 61, 55, and 52% of the iterations, respectively. Therefore, the average computation of the estimator system has also decreased in proportion to its update rate. The best results are obtained when both the SMF and WP techniques are combined. All of these curves are compared to the ideal 16QAM modulation on the AWGN channel. Curve (e) is performed in the absence of the CFO and only in the presence of the receiver and channel IQ and shows better results than the same algorithm (f).

Figure 5 shows a plot of BER versus SNR for the proposed SMF and WP-based methods for the NLMS-PTEQ algorithm in the PTEQ structure. This simulation result is derived considering both CFO and receiver frequency selective IQ. The modulation used is 64QAM and the results are compared with the LS algorithm with sufficient CP and unequal conditions. The proposed SMF-NLMS-PTEQ, WP-NLMS-PTEQ, and SMF-WP-NLMS-PTEQ algorithms, in addition to the LS algorithm, show good results in terms of BER than the conventional NLMS-PTEQ algorithm. Also in the curves (d), (e), (g), and (h) which used SMF technique, updating was done in 67, 63, 58, and 61% of the iterations, respectively. Therefore, the average computation of the estimator system has also decreased in proportion to its update rate. The best results are obtained when both the SMF and WP techniques are combined. All of these curves are compared to the ideal 64QAM modulation on the AWGN channel. The curves (e) and (h) are performed in the absence of the

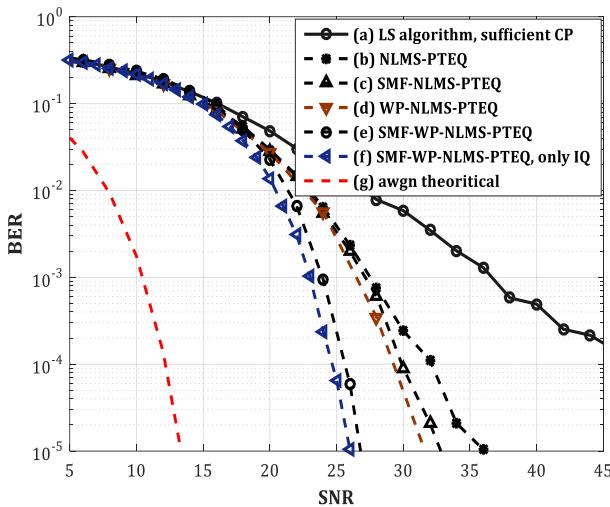


Figure 4. Error performance curve, 16QAM signal, training with NLMS and SMF-WP-NLMS-PTEQ under short CP in presence of CFO and receiver IQ

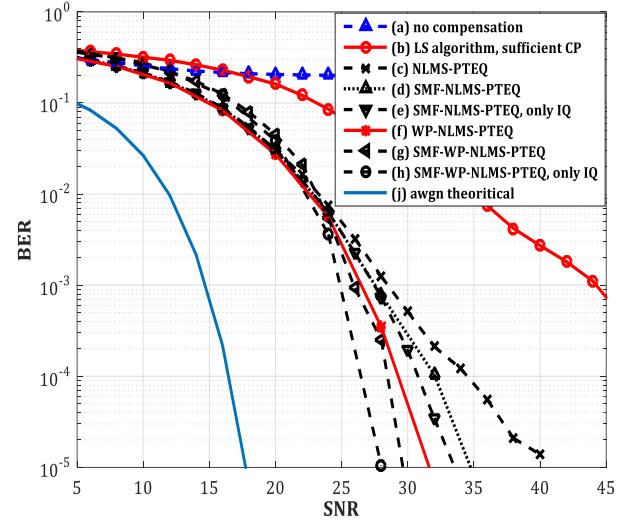


Figure 5. Error performance curve, 64QAM signal, training with NLMS and SMF-WP-NLMS-PTEQ under short CP in presence of CFO and receiver IQ

CFO and only in the presence of the receiver IQ and channel effect and show better results than the identical algorithms (d) and (h).

In Figure 6, the effect of increasing the value of the parameter L is investigated. By increasing the value of L , the performance is improved and ISI can be eliminated. Only 400 symbols are used to train the system in simulated curves. Also in the simulations where the CFO exists, the value of the normalized CFO is taken to be $\zeta = 0.1$. Also, the CFO is first estimated in the presence of IQ imbalances using the NLS algorithm proposed in [23]. The impulse response of the mismatched filters caused by the IQ imbalance in the receiver with 2-tap length $L_r = 2$, are considered $y_{r1} = [0.80.1]$ and $y_{r2} = [0.10.8]$. The branch lengths of the PTEQ structure are also considered to be 20. In the simulated figures, in addition to the normal state of the algorithm, the curves are compared with the LS algorithm used for channel estimation and with sufficient CP length.

As shown in Figure 6, the effect of decreasing L is investigated. By decreasing L , the bit error rate performance will deteriorate and the ISI can not be eliminated in the presence of a short CP. The error performance in the short CP approaches the non-compensation state, and by increasing the branch lengths of the PTEQ structure, the ISI effects of the short CP length are eliminated. In this simulation, the SMF-WP-NLMS-PTEQ algorithm is used for the curves (e) and (f). Also in curves (b), (c), and (d) the SMF-NLMS-PTEQ algorithm is used and the WP technique is not used due to

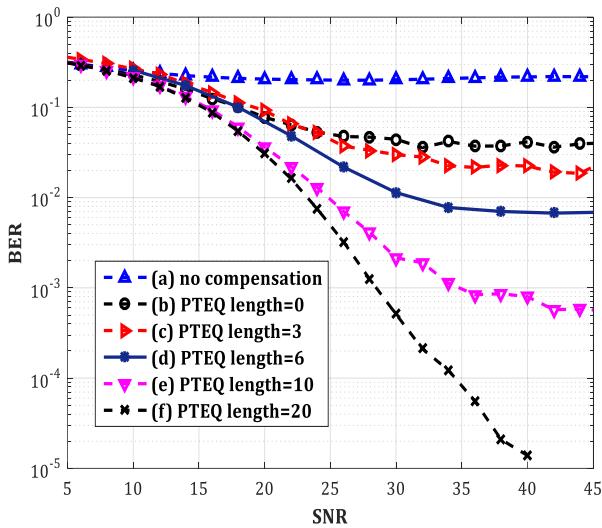


Figure 6. Error performance curve, 64QAM signal, Investigation of PTEQ branch lengths under short CP

the short length of the branches of the structure. However, the curves show that the use of PTEQ structure in short CP conditions is necessary to ISI effect elimination.

5. CONCLUDING REMARKS

In this paper, a new effective adaptive based method has been proposed for the joint estimation of the receiver IQ imbalance, CFO, and channel effects under short CP conditions. In this method, before estimating the channel, the CFO is first estimated and eliminated by existing methods. This algorithm is implemented using the structure of PTEQ. The PTEQ structure is capable of compensating channel and CFO efficiently under short CP length. Two parallel PTEQ structures are used to joint estimation of impairments. The proposed method employs an adaptive algorithm more optimally and its implementation in the wavelet transform domain has increased the convergence speed of the channel estimation. The use of PTEQ and wavelet transform along with the set-membership filtering concept, has made the proposed algorithm well capable of estimating and compensating channel effects in the presence of receiver IQ and CFO in short CP length. The average computational cost of the system has also decreased and the use of the system has become more cost-effective and implementable in real applications. Overall, the simulation results for the proposed 'SMF-WP-NLMS-PTEQ' algorithm under short CP and in the presence of receiver IQ and CFO showed sufficient improvement in BER along

with reduced computation, which makes this algorithm nearly ideal in terms of its performance.

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Persian Abstract

چکیده

در این مقاله، تخمین توازن کانال، عدم توازن IQ گیرنده وابسته به فرکانس و آفست فرکانسی حامل با طول CP کوتاه در سیستم OFDM در نظر گرفته شده است. برای جبرانسازی این خرابی‌ها، الگوریتم تطبیقی مبتنی بر ترکیب فیلترینگ set-membership مورد استفاده قرار گرفته است. در حالت CP کوتاه برای جلوگیری از تداخل بین سمبلي از ساختار PTEQ استفاده می‌شود. بخاطر وجود خرابی CFO و عدم توازن IQ در گیرنده، ساختار PTEQ را به حالت دو شاخه ای بسط خواهیم داد. این ساختار پیچیدگی محاسباتی بالایی داشته لذا در چنین شرایطی استفاده از ایده فیلترینگ set-membership با ضرب گام متغیر ضمن کاهش متوسط محاسبات سیستم، می‌تواند سرعت همگرایی تخمین‌ها را نیز افزایش دهد. از طرفی بهره‌گیری از تبدیل موجک بر روی هر شاخه این ساختار قبل از بکارگیری فیلترهای تطبیقی، سرعت تخمین را به نوبه‌ی خود افزایش خواهد داد. الگوریتم پیشنهادی با نام SMF-WP-NLMS-PTEQ ارائه خواهد شد. نتایج شبیه‌سازی‌ها نشان دهنده عملکرد بهتر از الگوریتم‌های تطبیقی معمول بوده است. علاوه بر این تخمین و جبرانسازی اثرات کانال، عدم توازن IQ گیرنده و آفست فرکانسی تحت CP کوتاه به سهولت بوسیله این الگوریتم قابل انجام می‌باشد.