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Sparse Repeated Preamble-Based Direct Current Offset and Frequency Offset Estimator in OFDM Direct Conversion Receivers

YUJIE XIA, JIANHUA CUI, AND JUNJIE ZHANG
School of Physics and Electronic Information, Luoyang Normal University, Luoyang 471022, China

Corresponding author: Yujie Xia (yjxia_2001@163.com)

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ABSTRACT Direct conversion receiver (DCR) architecture has drawn considerable attention in orthogonal frequency-division multiplexing (OFDM) systems. However, OFDM DCRs are sensitive to non-ideal front-end effects and suffer from direct current offset (DCO) and carrier frequency offset (CFO). In this paper, a novel method to estimate DCO and CFO is presented in OFDM DCRs by exploring the sparse symbols embedded in the preamble. In the proposed method, CFO and DCO estimations can be easily decoupled. Next, a robust low-complexity DCO estimation method is proposed by using the sparse properties of the preamble in the frequency domain. Moreover, a novel CFO estimation scheme independent of the DCO is presented based on the repeated preamble structure in which the DCO component focuses on the DC subcarrier when the received preamble signal was not CFO compensated in the frequency domain. The simulation results show that the proposed CFO and DCO estimators can achieve better performance than any available methods.

INDEX TERMS Direct current offset (DCO), carrier frequency offset (CFO), orthogonal frequency division multiplexing (OFDM), direct conversion receiver (DCR).

I. INTRODUCTION

In orthogonal frequency division multiplexing (OFDM) systems, direct conversion receiver (DCR) architecture has attracted extensive attention due to its simplicity and ease of integration [1]. Unfortunately, OFDM DCRs are sensitive to non-ideal radio frequencies (RFs) on the front end. Moreover, these devices suffer from direct current offset (DCO) and carrier frequency offset (CFO) [2]. Consequently, without proper DCO cancellation and CFO compensation, the estimation of important parameters is biased when performing signal demodulation and detection. Therefore, it is important to remove the DCO and compensate for the CFO in an OFDM DCR system to achieve good performance.

Some existing works have examined the problem of CFO and DCO estimation in OFDM DCR systems. In [3], the Bit Error Rate (BER) analysis was investigated for an OFDM DCR system impaired by both DCO and CFO over multipath Rayleigh fading channels. The performance of the CFO estimator is degraded by the DCO. The large residual CFO leads to inter-carrier interference (ICI), which severely decreases the overall system performance. Moreover, when the residual DCO is large, the DCO energy is spread to all the subcarriers after CFO compensation, further degrading the BER performance. In [4]–[7], many CFO estimation methods in the absence of any DCO were proposed for an OFDM DCR system. In these methods, the performance of the CFO estimation is satisfied, and the BER degradation is mostly negligible. However, those methods are invalid in the presence of DCO. In [8], a simple DCO estimation and cancellation scheme in the frequency domain was proposed with the assumption of CFO perfect compensation. However, the accuracy of the DCO estimation decreases if the CFO estimation is imperfect. Meanwhile, the residual DCO spreads the DCO energy to all the subcarriers and the residual CFO causes ICI, which decreases the system performance. In [9], the maximum likelihood (ML) estimator in an OFDM DCR system was discussed by taking in-phase and quadrature-phase (IQ) imbalance, channel, DCO and CFO into account. The ML estimator can achieve excellent performance for frequency selective channels, but the computational complexity is high.

Pilot-aided CFO and DCO estimation schemes were presented for OFDM DCR systems in [10]–[14]. The CFO estimator with a varying DCO was proposed based on differential filtering in [10]. The DCO can be almost removed by the received signal passing through the differential filter. However, the CFO performance decreases due to the extra noise introduced by the differential procedure. In [11], a time-domain average (TDA)-based DCO estimator was proposed using a simple average of the cumulative sum of multiple pilot symbols in the time domain. The CFO estimation is obtained after subtracting the coast estimated DCO power from the pilot correlation. Nevertheless, the CFO estimator is biased, and the BER performance decreased due to the residual CFO and DCO. In [12], the best linear unbiased estimator (BLUE) for the DCO was derived in an OFDM-based wireless local area network (WLAN) system with the CFO modeled as a random variable. Since the CFO is generally assumed to be deterministic, this modeling mismatch causes system performance loss. To improve the estimation accuracy of the DCO, a two-stage DCO estimation scheme was proposed in the presence of a CFO in [13]. However, the scheme is only limited to estimations of the DCO without considering the CFO estimation. In [14]–[18], blind joint CFO and DCO estimators were presented by exploring the subspace structure of the received OFDM preamble. In those methods, estimating the CFO in the time domain needs either global searching or singular value decomposition (SVD). Therefore, the high computational complexity limits their applications. In [19], the least squares (LS) CFO estimator was proposed in the presence of the DCO based on the TDA of the periodic pilot symbols. However, the accuracy of the CFO and DCO estimator is insufficient and the inversion of the estimated CFO matrix leads to high complexity in the DCO estimation. In [20], a novel null subspace (NSP) based the CFO and DCO estimation method was proposed according to the IEEE 802.11a standard [21] for an OFDM DCR system. A special separation matrix of a null subspace was designed to decouple the CFO and DCO variables. However, the CFO estimation degrades, as it suffers from energy loss based on the segmental correlation of the periodic preamble. Moreover, the residual CFO decreases the accuracy of the DCO estimation. In addition, the DCO estimation is still obtained as in [19] by using an inversion operation of the estimated CFO compensation matrix, which is highly complex.

The above discussion shows that the DCO and CFO are coupled with each other in an OFDM DCR system. When the CFO is compensated for, the residual CFO introduces ICI into the desired frequency signal, whereas the residual DCO leads to energy spreading to all the subcarriers. Therefore, to effectively solve the problems of DCO and CFO of an OFDM DCR, the key issue is how to decouple the DCO and CFO.

To improve the estimation performance of the DCO and CFO, we propose a novel DCO and CFO estimator for OFDM DCR systems. Considering the received frequency preamble signal without CFO compensation, the DCO component is focused on the DC subcarrier and the energy of the DCO is not spread over all the subcarriers, except for the DC subcarrier. Therefore, CFO and DCO estimations can be decoupled by exploring the sparse symbols embedded in the preamble. In the proposed scheme, a low-complexity DCO estimation method that is irrespective of the CFO presence is proposed in the frequency domain by employing the sparse symbols of the preamble. Furthermore, a novel CFO estimation scheme independent of the DCO is presented based on the repeated structure of the preamble. Simulation results show that the proposed estimator outperforms other existing methods, even in presence of large DCO and CFO.

This paper is organized as follows. Section II describes the signal model for the OFDM DCR in the presence of a CFO and DCO. We analyze the effect of the CFO and DCO for OFDM DCR with two cases, including CFO compensation and no CFO compensation presented in Section III. In Section IV, a sparse preamble-based DCO estimation method is presented and an effective CFO estimation scheme independent of the DCO is proposed, respectively. The performance of the proposed DCO and CFO estimator is analyzed in Section V. In Section VI, simulation-based performance comparisons are presented. Finally, our conclusions are drawn in Section VII.

II. SYSTEM MODEL

We consider a typical OFDM system with N subcarriers and M transmitted symbol blocks. The mth input OFDM symbol block in the frequency domain is denoted by $\mathbf{X}^{(m)} = [X_{-N/2}^{(m)}, \cdots, X_k^{(m)}, \cdots, X_{N/2-1}^{(m)}]^T$, where $X_k^{(m)}$ for $-N/2 \leq k \leq N/2-1$ is the complex data transmitted at the kth subcarrier and $m = 0, 1, \cdots, M-1$. Performing the N-point inverse discrete Fourier transform (IDFT), we obtain the discrete baseband signal for the mth OFDM block $\mathbf{x}^{(m)} = [x_0^{(m)}, \cdots, x_n^{(m)}, \cdots, x_{N-1}^{(m)}]^T$ in the time domain as

$$x_n^{(m)} = \frac{1}{\sqrt{N}} \sum_{k=-N/2}^{N/2-1} X_k^{(m)} e^{j2\pi kn/N}, \quad n = 0, 1, \cdots, N-1. \tag{1}$$

The baseband OFDM signal $\mathbf{x}^{(m)}$ is formed by adding a cyclic prefix (CP) to avoid inter-symbol interference (ISI) and then passes through a quasi-static multipath Rayleigh fading channel. To eliminate ISI, we assume that the duration of the CP is longer than the maximum delay spread of a multipath fading channel.

Let ε represent the normalized CFO between the transmitter and receiver, and d denote the static complex-valued DCO generated by the OFDM DCR. At the receiver, after down conversion and timing synchronization, the nth received sample of the mth OFDM block in the time domain after CP removal is given by

$$y_n^{(m)} = \frac{1}{\sqrt{N}} \sum_{k=-N/2}^{N/2-1} H_k X_k^{(m)} e^{j2\pi(k+\varepsilon)n/N} e^{j\phi^{(m)}} + d + w_n^{(m)}, \tag{2}$$

where $\phi^{(m)} = 2\pi\varepsilon(m(N+L_{CP})+L_{CP})/N$, L_{CP} is the length of CP, H_k is the frequency response at the kth subcarrier in the wireless channel and $w_n^{(m)}$ is the independently zero-mean additive white Gaussian noise (AWGN) with a variance of $\sigma_z^2 = E[|w_n^{(m)}|^2]$.

The matrix form of the above equation can be written as

$$\begin{aligned}\mathbf{y}^{(m)} &= [y_0^{(m)}, \cdots, y_n^{(m)}, \cdots y_{N-1}^{(m)}]^T \\ &= e^{j\phi^{(m)}}\Gamma(\varepsilon)\mathbf{F}^H\mathbf{D}^{(m)} + d\mathbf{1}_N + \mathbf{w}^{(m)}, \end{aligned} \tag{3}$$

where $\mathbf{F}$ is the $N \times N$ discrete Fourier transform (DFT) matrix with the elements $e^{-j2\pi nk/N}$ $(-N/2 \le k \le N/2-1$ and $0 \le n \le N-1)$ and $\Gamma(\varepsilon)$ being a diagonal matrix with a diagonal element $e^{j2\pi n\varepsilon/N}(0 \le n \le N-1)$. $\mathbf{D}^{(m)} = [D_{-N/2}^{(m)}, \cdots, D_k^{(m)}, \cdots, D_{N/2-1}^{(m)}]^T$ where $D_k^{(m)} = H_k X_k^{(m)}$ and $\mathbf{H} = [H_{-N/2}, \cdots, H_k, \cdots, H_{N/2-1}]^T$ is the frequency response in the wireless channel. The notation $\mathbf{1}_N$ represents an all 1 column vector with a size of $N \times 1$ and $\mathbf{w}^{(m)} = [w_0^{(m)}, \cdots, w_n^{(m)}, \cdots, w_{N-1}^{(m)}]^T$ denotes the AWGN vector in the time domain.

III. IMPACT ANALYSIS OF THE CFO AND DCO IN THE DCR

The analysis in [3] indicates that the DCR is sensitive to the CFO and DCO between the received waveform and the local oscillator signals because this introduces interference and results in severe performance degradation.

At the receiver, let $\hat{\varepsilon}$ denote the estimated normalized CFO and $y_n^{(m)}$ represent the received sample in the time domain. It is assumed that $y_n^{(m)}$ is compensated for by the estimated CFO and the compensation is performed by the DFT operation. Then, the DFT output in the frequency domain is given by

$$\begin{aligned}Y_k^{(m)} &= \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1} y_n^{(m)} e^{-j2\pi\hat{\varepsilon}n/N} e^{-j2\pi kn/N} \\ &= \alpha e^{j\phi^{(m)}} D_k^{(m)} + e^{j\phi^{(m)}} I_k^{(m)} + I_k^{DC} + W_k^{(m)}, \end{aligned} \tag{4}$$

where $\alpha = e^{j\pi\frac{(\varepsilon-\hat{\varepsilon})(N-1)}{N}}\text{sinc}(\frac{\varepsilon-\hat{\varepsilon}}{N})$ and $\text{sinc}(x) = \sin(\pi x)/(\pi x)$. The ICI at the kth subcarrier of the mth block is $I_k^{(m)} = \sum_{i=-N/2, i\neq k}^{N/2} D_i^{(m)} e^{j\pi\frac{(N-1)(i-k+\varepsilon-\hat{\varepsilon})}{N}}\text{sinc}(i - k + \varepsilon - \hat{\varepsilon})/\text{sinc}(\frac{i-k+\varepsilon-\hat{\varepsilon}}{N})$ with $D_i^{(m)} = H_i X_i^{(m)}$, $I_k^{DC} = d\sqrt{N}e^{-j\pi\frac{(N-1)(k+\hat{\varepsilon})}{N}}\text{sinc}(k + \hat{\varepsilon})/\text{sinc}(\frac{k+\hat{\varepsilon}}{N})$ being the DCO interference component at the kth subcarrier and $W_k^{(m)} = \frac{1}{\sqrt{N}}\sum_{n=0}^{N-1} w_n^{(m)} e^{-j2\pi\frac{(k+\hat{\varepsilon})n}{N}}$ defined as an independent complex Gaussian random variable with a zero mean and a variance of σ_z^2.

To illustrate the impact of the CFO and DCO on OFDM DCR systems, the following two cases are discussed.

(1) $\hat{\varepsilon} = 0$. This means that the received signal in the time domain has performed on the DFT without CFO compensation. According to (4), the ICI and DCO interference component on the received frequency OFDM signals can be expressed as

$$I_k^{(m)} = \sum_{i=-N/2, i\neq k}^{N/2} D_i^{(m)} e^{j\pi\frac{(N-1)(i-k+\varepsilon)}{N}} \frac{\text{sinc}(i-k+\varepsilon)}{\text{sinc}(\frac{i-k+\varepsilon}{N})} \tag{5}$$

and

$$\begin{aligned}I_k^{DC} &= d\sqrt{N}e^{-j\pi k(N-1)/N}\frac{\text{sinc}(k)}{\text{sinc}(k/N)} \\ &= \begin{cases} d\sqrt{N} & k=0 \\ 0 & otherwise. \end{cases}\end{aligned} \tag{6}$$

The DCO is only focused on the DC subcarrier and the DCO component has no effect on the system performance of the OFDM DCR. However, the ICI caused by the CFO significantly decreases the system performance.

(2) $\hat{\varepsilon} = \varepsilon$. This means that the received signal in the time domain is first compensated for with a perfect CFO estimation performed by DFT. According to (4),

$$I_k^{(m)} = 0, \tag{7}$$

and

$$I_k^{DC} = d\sqrt{N}e^{-j\pi(k+\varepsilon)(N-1)/N}\frac{\text{sinc}(k+\varepsilon)}{\text{sinc}((k+\varepsilon)/N)}. \tag{8}$$

The ICI is equal to zero and has no effect on the system performance of an OFDM DCR. However, the energy of the DCO is spread over all subcarriers and the interference from the DCO will cause severe performance degradation.

From the two cases discussed above, it is clear that the interference induced from the DCO and CFO will severely degrade system performance. Therefore, the DCO and CFO of the OFDM DCR must be estimated and compensated for using the received signal in the time domain to avoid the interference $I_k^{(m)}$ and I_k^{DC} after the DFT operation. However, the residual interference $I_k^{(m)}$ and I_k^{DC} depend on how perfectly the DCO cancellation and CFO compensation are performed in the time domain.

IV. PROPOSED CFO AND DCO ESTIMATOR

In this paper, we consider an OFDM burst-mode communication system where each burst is preceded by a consecutive preamble to assist in synchronization and channel estimation functions. We assume that the consecutive preambles consist of M identical OFDM blocks and each block is sparse symbols loaded. The preamble (especially if $M = 2$) can be generated by a format such as a short training sequence and a long training sequence according to the IEEE 802.11a standard [21].

Without the loss of generality, the total number of subcarriers N for each OFDM preamble block, are divided into Q active subcarriers (modulated subcarriers) and N-Qinactive subcarriers (null subcarriers). The indices of the modulated and null subcarriers are denoted by the sets $\mathbf{J} = \{J_1, \cdots, J_Q\}$ and $\Omega = \{\Omega_1, \cdots, \Omega_{N-Q}\}$, respectively. The transmitted symbols of the preamble defined by $\mathbf{J}$ are $\{X_{J_i}^{(m)} \neq 0, i = 1, 2, \cdots, Q\}$, whereas the transmitted symbol of the preamble defined by Ω, which is defined as $\{X_{\Omega_i}^{(m)} = 0, i = 1, 2, \cdots, N-Q\}$. In particular, the subcarriers -1, 0 and 1 are unloaded, which implies that $\{-1, 0, 1\} \in \Omega$ and $\{X_k^{(m)} = 0, k = -1, 0, 1\}$. Clearly, when N is large, the loss of spectral efficiency caused by those null subcarriers can be neglected.

From (5) and (6), when the normalized CFO is not compensated for in the received OFDM blocks, the energy of the DCO is only focused on the DC subcarrier and cannot spread over the other subcarriers in the frequency domain. Taking into account the sparse symbols embedded in the preamble with Q null subcarriers, including the subcarriers -1, 0 and 1, the energy of the ICI at the subcarrier 0 is almost the same as in the subcarriers 1 and -1 of the received frequency signal. This means that the DCO and CFO estimations can be solved by making full use of the sparse repeated preambles. The basic structure of the proposed DCO and CFO estimator is shown in Fig. 1.

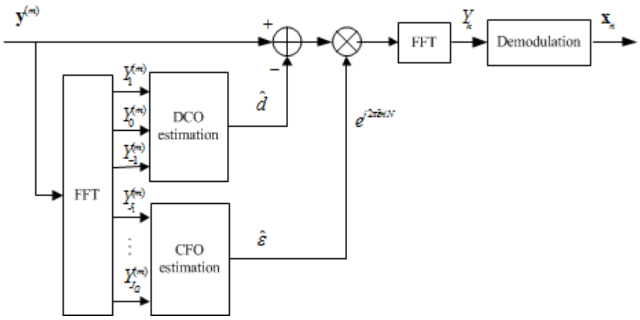

FIGURE 1. Basic structure of the proposed DCO and CFO estimator.

It is assumed that the sparse preamble consists of two identical OFDM blocks with $\mathbf{X} = \mathbf{X}^{(m)}$, $m = 0, 1$. According to (2), the nth received sample of the mth block of the preamble in the time domain after CP removal can be expressed as

$$y_n^{(m)} = \frac{1}{\sqrt{N}} \sum_{k=-N/2}^{N/2-1} D_k e^{j2\pi(k+\varepsilon)n/N} e^{j\theta} e^{j\varphi^{(m)}} + d + w_n^{(m)}, \quad (9)$$

where $\theta = 2\pi\varepsilon L_{CP}/N$, $\varphi^{(m)} = 2\pi\varepsilon m$ and $D_k = H_k X_k$. Note that $e^{j\theta}$ is a constant and is neglected for simplicity in the following discussion.

The received signal vector in the time domain can be written as

$$\begin{aligned} \mathbf{y}^{(m)} &= [y_0^{(m)}, \cdots, y_n^{(m)}, \cdots y_{N-1}^{(m)}]^T \\ &= e^{j\varphi^{(m)}} \Gamma(\varepsilon) \mathbf{F}^H \mathbf{D} + d\mathbf{1}_N + \mathbf{w}^{(m)}, \end{aligned} \quad (10)$$

where $\mathbf{D} = [D_{-N/2}, \cdots, D_k, \cdots, D_{N/2-1}]^T$.

At the receiver, since the CFO is unknown and the received signal in the time domain is not compensated by the CFO, the received preamble vector of the mth block in the frequency domain is

$$\begin{aligned} \mathbf{Y}^{(m)} &= [Y_{-N/2}^{(m)}, \cdots, Y_k^{(m)}, \cdots, Y_{N/2-1}^{(m)}]^T \\ &= \alpha e^{j\varphi^{(m)}} \mathbf{F}\Gamma(\varepsilon)\mathbf{F}^H \mathbf{D} + \sqrt{N} d\eta + \mathbf{W}^{(m)} \\ &= \alpha e^{j\varphi^{(m)}} \Lambda + \sqrt{N} d\eta + \mathbf{W}^{(m)}, \end{aligned} \quad (11)$$

where $Y_k^{(m)} = \alpha e^{j\varphi^{(m)}} D_k + e^{j\varphi^{(m)}} I_k^{(m)} + I_k^{DC} + W_k^{(m)}$, $\Lambda = [\Lambda_{-N/2}^{(m)}, \cdots, \Lambda_k^{(m)}, \cdots, \Lambda_{N/2-1}^{(m)}]^T = \mathbf{F}\Gamma(\varepsilon)\mathbf{F}^H\mathbf{D}$, $\eta = [\underbrace{0, \cdots, 0}_{N/2}, 1, \underbrace{0, \cdots, 0}_{N/2-1}]^T$ and $\mathbf{W}^{(m)} = [W_{-N/2}^{(m)}, \cdots, W_{N/2-1}^{(m)}]^T$.

A. SPARSE PREAMBLE-BAED DCO ESTIMATION

Bear in mind that since $\{X_k^{(m)} = 0, k = -1, 0, 1\}$, the received samples at the subcarriers 1 and -1 of the mth block of the preamble from (11) are given by

$$Y_1^{(m)} = e^{j\varphi^{(m)}} I_1^{(m)} + W_1^{(m)}, \quad (12)$$

and

$$Y_{-1}^{(m)} = e^{j\varphi^{(m)}} I_{-1}^{(m)} + W_{-1}^{(m)}. \quad (13)$$

Consequently, (12) and (13) can be rewritten as

$$\begin{bmatrix} Y_1^{(m)} \\ Y_{-1}^{(m)} \end{bmatrix} = \begin{bmatrix} e^{j\varphi^{(m)}} I_1^{(m)} \\ e^{j\varphi^{(m)}} I_{-1}^{(m)} \end{bmatrix} + \begin{bmatrix} W_1^{(m)} \\ W_{-1}^{(m)} \end{bmatrix}. \quad (14)$$

Considering that the sparse symbols $\{X_{\Omega_i}^{(m)} = 0, i = 1, 2, \cdots, N-Q\}$ are unloaded in the preamble, the following equation approximately holds (see appendix) for $N \gg \pi$:

$$I_1^{(m)} \approx I_0^{(m)} \approx I_{-1}^{(m)}. \quad (15)$$

Combining (14) and (15) yields

$$\begin{bmatrix} Y_1^{(m)} \\ Y_{-1}^{(m)} \end{bmatrix} = \begin{bmatrix} 1 \\ 1 \end{bmatrix} e^{j\varphi^{(m)}} I_0^{(m)} + \begin{bmatrix} W_1^{(m)} \\ W_{-1}^{(m)} \end{bmatrix}. \quad (16)$$

Then, we can obtain

$$e^{j\varphi^{(m)}} \hat{I}_0^{(m)} = \frac{1}{2} \left[Y_1^{(m)} + Y_{-1}^{(m)} \right]. \quad (17)$$

The received sample at the subcarriers 0 is given by

$$Y_0^{(m)} = e^{j\varphi^{(m)}} I_0^{(m)} + W_0^{(m)} + d\sqrt{N}. \quad (18)$$

For $m = 0, 1$, we have

$$\begin{bmatrix} Y_0^{(0)} - e^{j\varphi^{(0)}} I_0^{(0)} \\ Y_0^{(1)} - e^{j\varphi^{(1)}} I_0^{(1)} \end{bmatrix} = d\sqrt{N} \begin{bmatrix} 1 \\ 1 \end{bmatrix} + \begin{bmatrix} W_0^{(0)} \\ W_0^{(1)} \end{bmatrix}. \quad (19)$$

The term d can be estimated as

$$\hat{d} = \frac{1}{2\sqrt{N}} \sum_{m=0}^{1} \left[Y_0^{(m)} - e^{j\varphi^{(m)}} I_0^{(m)} \right]. \quad (20)$$

Substituting (17) into (20), the final estimated DCO $\hat{d}$ is given by

$$\hat{d} = \frac{1}{2\sqrt{N}} \left[\sum_{m=0}^{1} Y_0^{(m)} - \frac{1}{2} \sum_{m=0}^{1} \left[Y_1^{(m)} + Y_{-1}^{(m)} \right] \right]. \quad (21)$$

B. CFO ESTIMATION INDEPENDENT OF THE DCO

After finishing the DCO estimation, a direct idea is to estimate the CFO from the received signal in the time domain via DCO cancellation. According to (10), the two $N \times 1$ vectors $\mathbf{y}^{(0)}$ and $\mathbf{y}^{(1)}$ from the received signal in the time domain can be expressed as

$$\mathbf{y}^{(0)} = \Gamma(\varepsilon)\mathbf{F}^H \mathbf{D} + d\mathbf{1}_N + \mathbf{w}^{(0)}, \quad (22)$$

and

$$\mathbf{y}^{(1)} = e^{j2\pi\varepsilon} \Gamma(\varepsilon)\mathbf{F}^H \mathbf{D} + d\mathbf{1}_N + \mathbf{w}^{(1)}. \quad (23)$$

Removing the estimated DCO from $\mathbf{y}^{(0)}$ and $\mathbf{y}^{(1)}$, we obtain

$$\mathbf{z}^{(0)} = \mathbf{y}^{(0)} - \hat{d}\mathbf{1}_N = \Gamma(\varepsilon)\mathbf{F}^H\mathbf{D} + \mathbf{w}^{(0)}, \tag{24}$$

and

$$\mathbf{z}^{(1)} = \mathbf{y}^{(1)} - \hat{d}\mathbf{1}_N = e^{j2\pi\varepsilon}\Gamma(\varepsilon)\mathbf{F}^H\mathbf{D} + \mathbf{w}^{(1)}. \tag{25}$$

The correlation variable is defined as

$$R = (\mathbf{z}^{(0)})^H\mathbf{z}^{(1)} = e^{j2\pi\varepsilon}|A|^2 + R_z, \tag{26}$$

where $R_z = e^{j2\pi\varepsilon}(\mathbf{w}^{(0)})^H\mathbf{A} + \mathbf{A}^H\mathbf{w}^{(1)} + (\mathbf{w}^{(0)})^H\mathbf{w}^{(1)}$ and $|A|^2 = \mathbf{A}^H\mathbf{A}$ with $\mathbf{A} = \Gamma(\varepsilon)\mathbf{F}^H\mathbf{D}$.

The CFO can be estimated based on (26) as

$$\hat{\varepsilon} = \frac{1}{2\pi}\arg\{R\}. \tag{27}$$

According to (24) and (25), the accuracy of the CFO estimation strongly depends on how completely the DCO is cancelled.

To improve the estimation accuracy of the CFO in the presence of the DCO, we propose a novel CFO estimator independent of the DCO in the frequency domain. Note that the modulated symbols $\{X_{J_i}^{(m)} \neq 0, i = 1, 2, \cdots, Q\}$ are embedded in the preamble and the received frequency vector of the preamble at the modulated subcarriers defined by $\mathbf{J}$ can be written based on (11) as

$$\begin{aligned}\bar{\mathbf{Y}}^{(m)} &= [Y_{J_1}^{(m)}, \cdots, Y_{J_Q}^{(m)}]^T \\ &= e^{j2\pi m\varepsilon}\bar{\mathbf{\Lambda}} + \bar{\mathbf{W}}^{(m)},\end{aligned} \tag{28}$$

where $\bar{\mathbf{\Lambda}} = [\Lambda_{J_1}^{(m)}, \cdots, \Lambda_{J_Q}^{(m)}]^T$ and $\bar{\mathbf{W}}^{(m)} = [W_{J_1}^{(m)}, \cdots, W_{J_Q}^{(m)}]^T$.

For m =0, 1, we have

$$\bar{\mathbf{Y}}^{(0)} = \bar{\mathbf{\Lambda}} + \bar{\mathbf{W}}^{(0)}, \tag{29}$$

and

$$\bar{\mathbf{Y}}^{(1)} = e^{j2\pi\varepsilon}\bar{\mathbf{\Lambda}} + \bar{\mathbf{W}}^{(1)}. \tag{30}$$

We define a function:

$$\bar{R} = \left(\bar{\mathbf{Y}}^{(1)}\right)^H\bar{\mathbf{Y}}^{(2)} = e^{j2\pi\varepsilon}|B|^2 + \bar{R}_z, \tag{31}$$

where $\bar{R}_z = e^{j2\pi\varepsilon}(\bar{\mathbf{W}}^{(0)})^H\bar{\mathbf{\Lambda}} + \bar{\mathbf{\Lambda}}^H\bar{\mathbf{W}}^{(1)} + (\bar{\mathbf{W}}^{(0)})^H\bar{\mathbf{W}}^{(1)}$ and $|B|^2 = \bar{\mathbf{\Lambda}}^H\bar{\mathbf{\Lambda}}$.

The CFO can be estimated as

$$\hat{\varepsilon} = \frac{1}{2\pi}\arg\{\bar{R}\}. \tag{32}$$

Once $\hat{d}$ and $\hat{\varepsilon}$ are estimated by (21) and (32), the DCO component can be removed and the CFO is compensated in the time domain for the received signals. Consequently, the received signals can be corrected as

$$\tilde{\mathbf{y}}^{(m)} = e^{j2\pi m\hat{\varepsilon}}\Gamma^H(\hat{\varepsilon})(\mathbf{y}^{(m)} - \hat{d}\mathbf{1}_N). \tag{33}$$

V. PERFORMANCE ANALYSES

A. PERFORMANCE OF THE DCO ESTIMATION

Substituting (12) and (13) into (21), we have

$$\hat{d} = d + \frac{1}{2\sqrt{N}}\left\{\sum_{m=0}^{1}W_0^{(m)} + \frac{1}{2}\sum_{m=0}^{1}(W_1^{(m)} + W_{-1}^{(m)})\right\}. \tag{34}$$

We let $\eta = \hat{d} - d$ denote the estimation error of the DCO and can be given as

$$\eta = \frac{1}{2\sqrt{N}}\left\{\sum_{m=0}^{1}W_0^{(m)} + \frac{1}{2}\sum_{m=0}^{1}(W_1^{(m)} + W_{-1}^{(m)})\right\}. \tag{35}$$

For $E\{W_k^{(m)}, m = 0, 1\} = 0$ and $E[|W_k^{(m)}|^2] = \sigma_z^2$, $E[\eta] = 0$. The DCO estimator is unbiased and the mean squared error (MSE) can be calculated as

$$E[|\eta|^2] = \frac{3\sigma_z^2}{4N}. \tag{36}$$

B. PERFORMANCE OF THE CFO ESTIMATON

Equation (31) can be rewritten as

$$\bar{R} = e^{j2\pi\varepsilon}|B|^2\left(1 + e^{-j2\pi\varepsilon}\frac{\bar{R}_z}{|B|^2}\right). \tag{37}$$

From (32), we see that the CFO estimation error $\hat{\varepsilon} - \varepsilon$ is caused by the angle of the bracketed term. For $|\hat{\varepsilon} - \varepsilon| \ll \frac{1}{2\pi}$ and a high SNR, the CFO estimation error can be approximated as

$$\hat{\varepsilon} - \varepsilon \approx \frac{1}{2\pi}Im\left\{\frac{e^{-j2\pi\varepsilon}\bar{R}_z}{|B|^2}\right\}. \tag{38}$$

This equation shows that $E[\hat{\varepsilon} - \varepsilon] = 0$, which indicates that the proposed CFO estimator is unbiased for small errors and a high SNR, and the MSE of the CFO estimation error is given as

$$E[|\hat{\varepsilon} - \varepsilon|^2] = \frac{\sigma_z^2}{(4\pi)^2|B|^2}. \tag{39}$$

TABLE 1. Chosen parameters for the simulations.

Parameters	Value
Number of subcarriers	64
CP length	16
Frame length	16 OFDM blocks
Bandwidth	20 MHz
constellation order	16 QAM / 64 QAM
channel	Rayleigh fading

VI. SIMULATION RESULTS

In this section, we evaluate the performance of the proposed DCO and CFO estimator by comparing it with the LS estimator in [19] and NSP estimator in [20]. The system parameters of the OFDM system are taken from the 802.11a standard [1] and some of the main parameters are listed in Table I. The sparse preamble is generated in accordance with short training sequence in the IEEE 802.11a standard. The frequency-selective fading channel has six paths, with an exponential power delay profile, and the root mean square delay spread is equal to 100 ns. The power of the DCO is set at $|d|^2 \in [0, 1]$ and the range of the CFO is $\varepsilon \in (-0.5, 0.5)$. We used the MSE of the CFO and DCO, which is defined as $E\{|\varepsilon - \hat{\varepsilon}|^2\}$ and $E\{|d - \hat{d}|^2\}$, to assess the performance of the estimators.

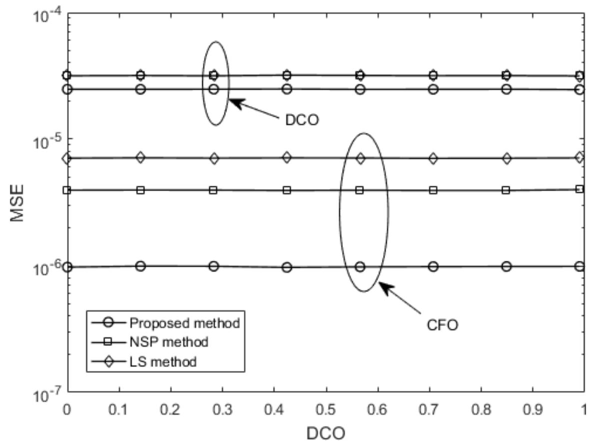

FIGURE 2. MSE versus different values of the DCO for CFO = 0.3, SNR = 20 dB.

Figure 2 shows that the MSE of the DCO and CFO versus different DCO d values at a normalized $\varepsilon = 0.3$ and SNR = 20 dB. The proposed estimator can achieve better DCO and CFO performances than both the LS and NSP methods. As expected, the DCO and CFO performances for the proposed estimator do not vary with different DCO values. Using the sparse property of the preambles, the proposed DCO method has robust performance irrespective of the CFO. Both the LS and NSP estimators have the same DCO performance, as they suffer from the ICI introduced by the residual CFO. The simulation results also indicate that the proposed CFO method can achieve satisfactory performance independent of the DCO and based on the repeated properties of the two identical preambles. When the received signals in the time domain are transformed via FFT without CFO compensation, the DCO component cannot spread over all the other subcarriers, except for the DC subcarrier from (6). Therefore, the proposed CFO estimator can completely remove the effect of the DCO component by using the loaded subcarriers of the repeated preambles. However, both the LS and NSP estimators degrade the CFO estimated performance, as the LS method for CFO estimation introduces extra noise due to the DCO component, whereas the NSP method causes energy loss via the segmental correlation operation.

Figure 3 shows that the MSE of the DCO and CFO versus different the normalized CFO at $d = 0.6(1 + j)$ and SNR = 20 dB. The proposed estimator is robust in terms of the DCO and CFO and has better CFO and DCO estimation performances than both the LS and NSP methods.

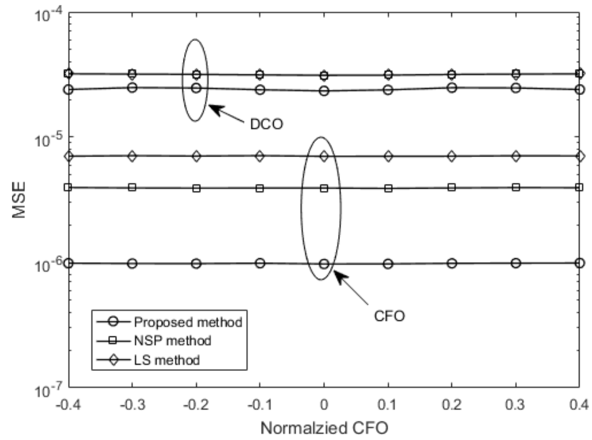

FIGURE 3. MSE versus a normalized CFO for DCO = 0.6(1+j), SNR = 20 dB.

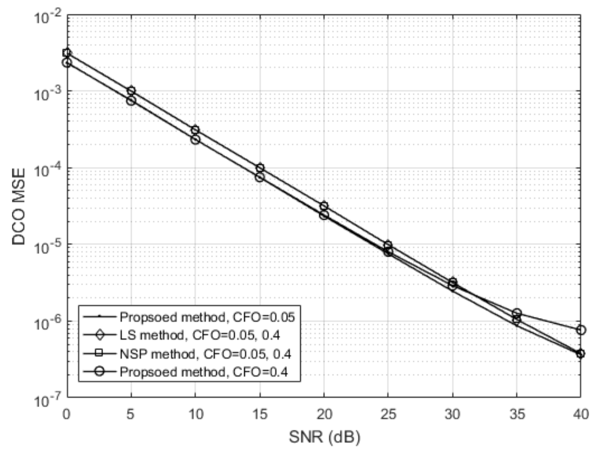

FIGURE 4. DCO MSE versus SNRs, DCO = 0.2(1+j).

When examined using varying normalized CFO values, the proposed DCO and CFO estimator maintains stable performance.

Figure 4 shows the MSE of the DCO estimation versus different SNRs at $d = 0.2(1 + j)$ with $\varepsilon = 0.05$ and $\varepsilon = 0.4$. At low SNR regions, the proposed CFO estimator has better performance than both the LS and NSP estimators for both large and small normalized CFO values. The reason for this is that at low SNR regions, the magnitudes of the AWGN at the subcarriers 1, 0 and -1 are dominant in the proposed DCO estimator, and the estimated value of $I_0^{(m)}$ in (16) results in extra noise that can be neglected. However, with the normalized CFO increasing (e.g., $\varepsilon = 0.4$), the proposed DCO method exhibits an error floor at high SNR regions and causes a small performance loss compared to both the LS and NSP methods. This is because the magnitudes of the ICI introduced by the CFO at subcarriers 1, 0 and -1 are dominant in the proposed DCO estimator at high SNR regions. For a given ε, the estimated value of $I_0^{(m)}$ in (16) leads to a deviation and causes the error floor to occur.

Figure 5 shows the MSE of the CFO estimation versus different SNRs at a normalized $\varepsilon = 0.3$ and $d = 0.2(1 + j)$.

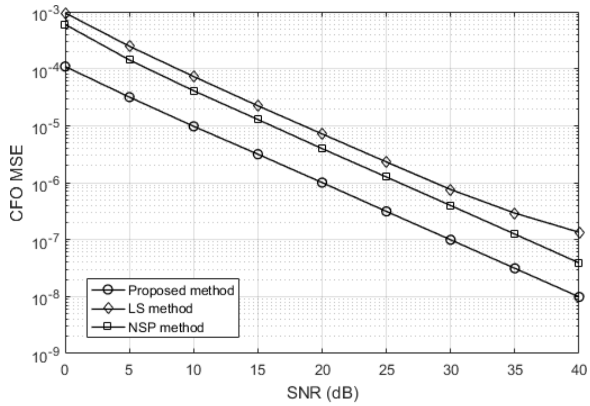

FIGURE 5. CFO MSE versus SNRs, CFO = 0.3, DCO = 0.2(1+j).

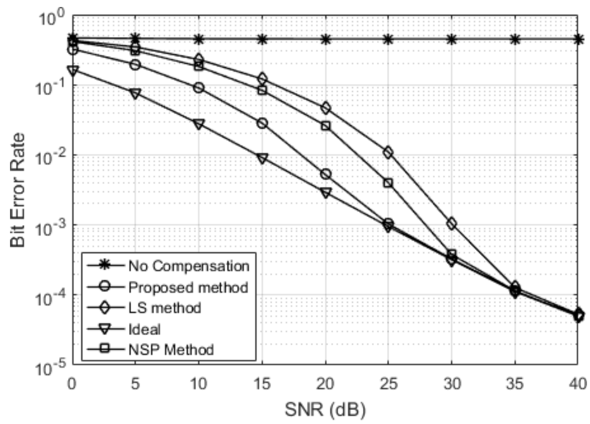

FIGURE 6. BERs of 16QAM modulated OFDM signals in multipath Rayleigh fading channels, CFO = 0.3 and DCO = 0.2(1+j).

Since the proposed CFO method is independent of the DCO, this method exhibits the best CFO performance over all three of the methods. For the LS method, the remaining DCO after the TDA-based DCO cancellation degrades the CFO performance. The NSP method uses a null subspace matrix to remove the DCO, but the frequency-selective fading channel results in CFO performance degradation.

The overall reception performance is measured in terms of the BER after the DCO cancellation and CFO compensation using the obtained estimations. Figures 6 and 7 show the BERs of 16QAM and 64QAM modulated OFDM signals in multipath Rayleigh fading channels for $\varepsilon = 0.3$ and $d = 0.2(1+j)$, respectively. In the simulation, the performance comparison is also made using an ideal system with no DCO and CFO distortion in the DCR and with a system without DCO cancellation and CFO compensation scheme included. It can be seen that, without DCO cancellation and CFO compensation scheme, the OFDM system is completely unusable. The BER performance of the proposed estimator, which can achieve satisfactory CFO and DCO estimations, outperforms that of both the LS and NSP methods. When the SNR is greater than 30 dB, the BER performance of the proposed method for the 64QAM modulated OFDM signals is very close to the ideal case.

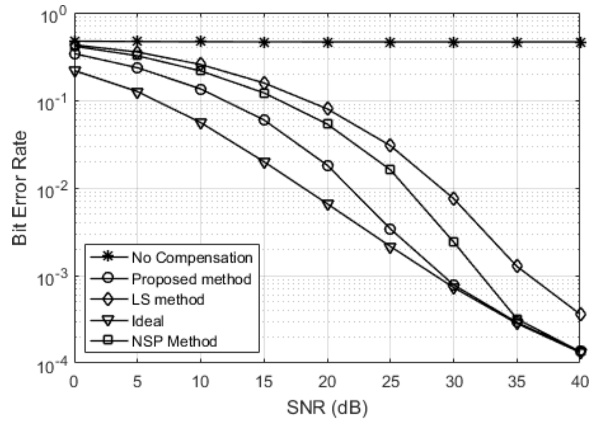

FIGURE 7. BERs of the 64QAM modulated OFDM signals in multipath Rayleigh fading channels, CFO = 0.3 and DCO = 0.2(1+j).

VII. CONCLUSIONS

In this paper, the DCO and CFO estimation and compensation problems in OFDM DCR were addressed. A model of the received signal with the DCO and CFO was first derived and the impacts of the DCO and CFO were analyzed for system performance. By exploring the sparse symbols embedded in the preamble, a robust low-complexity DCO estimation method was proposed in the frequency domain. Considering that the DCO component focuses on the DC subcarrier when the received preamble signal does not have CFO compensation in the frequency domain, a novel CFO estimation scheme independent of the DCO was presented based on the repeated preamble structure. Simulation results show that the performances of the proposed estimator outperform available methods, even in the presence of large DCO and CFO. The proposed DCO and CFO estimator is a practical solution in an OFDM DCR according to the IEEE 802.11 and 802.16 standards. In addition, this method can also be used in other DCR applications with a sparse repeated preamble.

APPENDIX

The ICI on the kth subcarrier is

$$I_k^{(m)} = \sum_{i=-N/2, i\neq k}^{N/2} D_i^{(m)} e^{j\pi \frac{(N-1)(i-k+\varepsilon)}{N}} \frac{\mathrm{sinc}(i-k+\varepsilon)}{\mathrm{sinc}\left[\frac{(i-k+\varepsilon)}{N}\right]}. \tag{40}$$

For $k = 0$, the ICI at the subcarrier 1 can be written as

$$\begin{aligned} I_1^{ICI} &= \sum_{i=-N/2, m\neq 1}^{N/2} D_i^{(m)} e^{j\pi \frac{(N-1)(i-1+\varepsilon)}{N}} \frac{\mathrm{sinc}(i-1+\varepsilon)}{\mathrm{sinc}[(i-1+\varepsilon)/N]} \\ &= \sum_{i=-N/2, i\neq 1}^{N/2} D_i^{(m)} e^{j\pi \frac{(N-1)(i+\varepsilon)}{N}} e^{j\pi/N} \frac{\frac{\sin[\pi(i+\varepsilon)]}{\pi(i+\varepsilon-1)}}{\frac{\sin[\pi(i+\varepsilon-1)/N]}{\pi(i+\varepsilon-1)/N}}. \end{aligned} \tag{41}$$

For $N \gg \pi$, the following equations $\cos(\pi/N) \approx 1$ and $\sin(\pi/N) \approx 0$ hold. When $|i| \gg 1$, we obtain

$$\begin{aligned} I_1^{(m)} &\approx \sum_{i=-N/2, i\neq 1}^{N/2} D_i^{(m)} e^{j\pi \frac{(N-1)(i+\varepsilon)}{N}} \frac{\frac{\sin[\pi(i+\varepsilon)]}{\pi(i+\varepsilon)}}{\frac{\sin[\pi(i+\varepsilon)/N]}{\pi(i+\varepsilon)/N}} \\ &= \sum_{i=-N/2, i\neq 1}^{N/2} D_i^{(m)} e^{j\pi \frac{(N-1)(i+\varepsilon)}{N}} \frac{\operatorname{sinc}(i+\varepsilon)}{\operatorname{sinc}[(i+\varepsilon)/N]} \\ &= I_0^{ICI}. \end{aligned} \tag{42}$$

Similarly, the ICI at the subcarrier -1 can be approximately expressed as

$$\begin{aligned} I_{-1}^{(m)} &= \sum_{i=-N/2, i\neq -1}^{N/2} D_i^{(m)} e^{j\pi \frac{(N-1)(i+1+\varepsilon)}{N}} \frac{\operatorname{sinc}(i+1+\varepsilon)}{\operatorname{sinc}[(i+1+\varepsilon)/N]} \\ &\approx \sum_{i=-N/2, i\neq -1}^{N/2} D_i^{(m)} e^{j\pi \frac{(N-1)(i+\varepsilon)}{N}} \frac{\operatorname{sinc}(i+\varepsilon)}{\operatorname{sinc}[(i+\varepsilon)/N]} \\ &= I_0^{ICI}. \end{aligned} \tag{43}$$

Comparing (42) with (43), it is clear that $I_1^{(m)} \approx I_0^{(m)}$ and $I_{-1}^{(m)} \approx I_0^{(m)}$ hold true for $N \gg \pi$.

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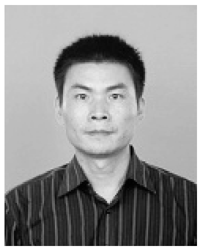
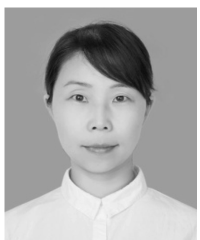
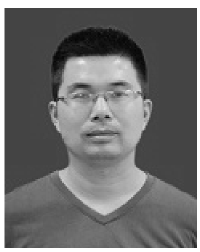


YUJIE XIA received the B.S. degree in electronic engineering from Henan Normal University, China, in 2001, the M.S. degree in communication and information systems from Harbin Engineering University, China, in 2004, and the Ph.D. degree in communication and information systems from Xidian University, China, in 2014. Since 2004, he has been with the School of Physics and Electronic Information, Luoyang Normal University, China. His research interests include broadband wireless access, orthogonal frequency-division multiple access techniques, and signal processing in communication systems.



JIANHUA CUI received the B.S. degree in electronic engineering and the M.S. degree in circuits and systems from Zhengzhou University, China, in 2003 and 2006, respectively, and the Ph.D. degree in information and communication engineering from the National Digital Switching System Engineering and Technological Research Center, in 2017. She has been a Teacher with the School of Physics and Electronic Information, Luoyang Normal University, since 2006. Her research interests include cooperative positioning and wireless network signal processing.



JUNJIE ZHANG received the B.S. degree in electronic engineering from Shaanxi Normal University, China, in 2001, and the M.S. degree in communication and information systems from Xiangtan University, China, in 2008. Since 2001, he has been with the School of Physics and Electronic Information, Luoyang Normal University, China. His research interests include wireless communication and PAPR for OFDM communication systems.