

Time-Frequency-Multiplex Preamble Design for Joint I/Q Imbalance, CFO and Channel Estimation in OFDM Systems

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Abstract— In this paper, a novel time-frequency-multiplex preamble design is proposed to estimate the in-phase and quadrature-phase (I/Q) imbalance, carrier frequency offset (CFO) and channel impulse response parameters for the zero-IF receiver of orthogonal frequency-division multiplexing (OFDM) systems. First, the contiguous time-multiplex preamble is designed to estimate CFO. Then, the two CFO compensators of the positive and negative subcarriers are proposed to separate and estimate I/Q imbalance and channel parameters. Finally, the image signal cancellation via the estimated parameters is designed to enhance the receiver performance. The advantage of the proposed time-frequency-multiplex preamble design is low computational complexity and robust performance. Simulation results confirm that the proposed estimator can provide reliable performance over the severe I/Q imbalance, CFO, and multipath fading channel environment.

Keywords— OFDM, CFO, I/Q imbalance, channel estimation, equalization.

I. INTRODUCTION

Recently, due to the requirement of low power consumption, single chip design, and low computational complexity, the zero-IF receiver has drawn a lot of attentions and exhibited great advantages in the high data rate wireless communication systems, i.e., OFDM systems. However, some gain and phase mismatches in zero-IF receiver can seriously degrade the system performance. One of these mismatches is so-called the in-phase and quadrature-phase (I/Q) imbalance which is due to the RF signal direct convert to baseband signal with the amplitude and phase mismatches between the I and Q-branch of the local oscillators. Besides, carrier frequency offset (CFO) and multipath fading channel effects still cause the performance degradation of the OFDM transceiver. CFO is induced by the transmitter and receiver with carrier frequency mismatch. In summary, the performance of OFDM system is limited by the presence of multipath fading channel, RF impairment with I/Q imbalance and CFO.

In order to remedy these problems, several schemes have been proposed to estimate and compensate for the RF impairment and channel effect. Among these works, some schemes use the data-aided methods [1]-[2], i.e., the special pilots and training symbols utilized in the transmitted data. Some papers employ the nondata-aided methods [3]-[4],

which only use the transmitted data without other information. Other filter-based method [5] and blind estimator [6] are also proposed to deal with the problem of the RF impairment and channel effects. The above schemes require a lot of converged cycles or large computational complexity to estimate and compensate I/Q imbalance, CFO, channel effects. In this paper, we propose a low computing scheme with a few preamble symbols to estimate and compensate the RF impairment and channel effects.

The processing procedures of the proposed scheme are summarized as follows. First, two sets of the same preambles can estimate the offset frequency via the difference time received data. The received signal with offset frequency can assist to cancel some CFO effect. Second, the time-multiplex preamble design can induce the OFDM received signal being divided into two parts, lower part subcarriers and upper part subcarriers. One part with the original signal and the other part with image signal involve some relative signal due to I/Q imbalance impact. Thus, we can estimate the parameters of I/Q imbalance and channel response via correlation, conjugate, and elimination techniques. Finally, to enhance the performance of the parameters estimation, a feedback cancellation technique is proposed to combat some residual interference effect. To sum of, the proposed time-frequency-multiplex preamble design method is very simple, so it can take a low computational complexity to estimate the RF impairment and channel parameters for OFDM systems. Simulation results have confirmed the validity and superiority of the proposed preamble design.

II. DATA MODEL

Reconsider an OFDM data model for a scenario in multipath fading channel with frequency offset and I/Q/imbalance. First, the transmitted OFDM signal is expressed by

$$x(t) = \sum_{k=0}^{N-1} X_k e^{j2\pi kt/T}, 0 \leq t < T \quad (1)$$

After the transmitted signal is passing through the channel, the received data can be written as

$$y_n(t) = h(t) \otimes x(t) + n(t) \quad (2)$$

where $\otimes$ denotes linear convolution, $h(t)$ is the multipath

fading channel with L paths, and $n(t)$ is an additive white gaussian noise (AWGN). When the receiver is affected by I/Q imbalance and CFO, the received sampled signal can be expressed by [7]

$$y(n) = k_1 E(n) y_n(n) + k_2 E^*(n) y_n^*(n) \quad (3)$$

where $E(n) = \exp(j2\pi\Delta f n T_s)$ is the time-domain carrier frequency offset signal with Δf being a CFO parameter and T_s being the sampling time. I/Q imbalance parameter is $k_1 = (1 + g e^{-j\phi})/2$ and $k_2 = (1 - g e^{j\phi})/2$ where g is gain mismatch and ϕ is phase mismatch. The relationship of k_1 and k_2 is given by

$$k_1 + k_2^* = 1 \quad (4)$$

$$k_1 - k_2^* = g e^{-j\phi} = \mu \quad (5)$$

After the CP removal and fast Fourier transform (FFT), the post-FFT received signal vector is expressed by

$$\mathbf{y} = \mathbf{F} \{ k_1 E(n) y_n(n) + k_2 E^*(n) y_n^*(n) \} \quad (6)$$

$$= k_1 \mathbf{G} \mathbf{H} \mathbf{x} + k_2 \mathbf{G}^* \mathbf{H}^* \mathbf{x}^{\#}$$

where the transmitted signal $\mathbf{x}$ is assigned to N subcarriers, which setup null transmission on the DC and $N/2$ -th subcarriers for reducing interference. And for the lower-side subcarriers, i.e., $k = 0 \sim N/2 - 1$, it denotes the positive part of the transmitted signal. On the contrary, for the upper-side subcarriers, i.e., $k = N/2 \sim N - 1$, the transmitted signal is denoted by the negative part. The $\#$ operation represents conjugate mirror operation, i.e. the carriers take conjugate and

turn upside down. Next, $\mathbf{G} = \begin{bmatrix} \mathbf{G}_1 & \mathbf{G}_2 \\ \mathbf{G}_2^* & \mathbf{G}_1^* \end{bmatrix} = \mathbf{F} \text{diag} \{ E(n) \} \mathbf{F}^H$ is a circular matrix with symmetric property, i.e., $\mathbf{G}^* = \mathbf{G}^H = \mathbf{F} \text{diag} \{ E(n) \}^* \mathbf{F}^H$. $\mathbf{H} = \mathbf{F} \text{diag} (h(n)) \mathbf{F}^H = \begin{bmatrix} \mathbf{H}_1 & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_2 \end{bmatrix}$

is a diagonal frequency-domain channel matrix.

III. DEVELOPMENT OF JOINT IQ IMBALANCE, CFO, AND CHANNEL ESTIMATION

In order to estimate the I/Q imbalance, CFO, and channel parameters jointly, the time-frequency-multiplex preamble design is proposed for OFDM systems. The overall schematic diagram of the proposed receiver is depicted in Figure 1.

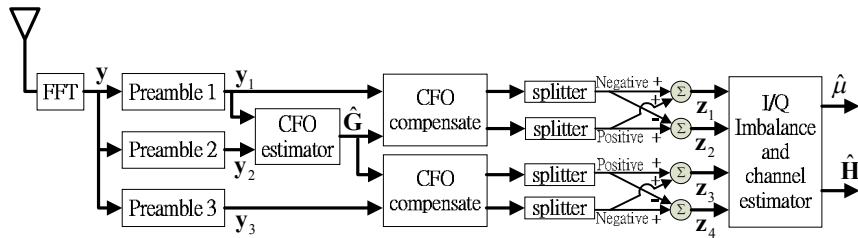


Figure 1. Block diagram of the proposed receiver.

A. Preamble Design

Preamble design is a core design in this paper, which assists receiver to estimate parameters and overcome to image interference. That is, as shown in Figure 2, the two contiguous preambles with same training sequences in preamble 1 and 2 can be used to estimate CFO parameter via cross-correlation scheme. Next, I/Q imbalance and channel parameters are estimated by two preambles with time-frequency-multiplex nulling techniques, i.e., negative subcarriers nulling in preamble 1 and positive subcarriers nulling in preamble 3, which is shown in Figure 2. The preamble sequence $\mathbf{s}_1$ is designed by Chu-sequence [8], which contains the low peak-to-average power ratio(PAPR) and orthogonal property.

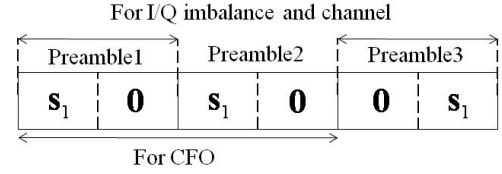


Figure 2. Time-Frequency-Multiplex Preamble Design

B. CFO Estimation

In this subsection, we use preamble 1 and preamble 2 to estimate the CFO parameter. Preamble 1 and preamble 2 are the same sequence. Time delay between contiguous preambles can provide the exponential effect in frequency domain. Then, we utilize the property to estimate the CFO parameter. Next, for detail description, substituting preamble 1 into $\mathbf{y}$, the post-FFT received signal can be expressed by

$$\mathbf{y}_1 = \begin{bmatrix} \mathbf{y}_{11} \\ \mathbf{y}_{12} \end{bmatrix} = k_1 \begin{bmatrix} \mathbf{G}_1 & \mathbf{G}_2 \\ \mathbf{G}_2^* & \mathbf{G}_1^* \end{bmatrix} \begin{bmatrix} \mathbf{H}_1 & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_2 \end{bmatrix} \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{0} \end{bmatrix} + k_2 \begin{bmatrix} \mathbf{G}_1^* & \mathbf{G}_2^* \\ \mathbf{G}_2^* & \mathbf{G}_1^* \end{bmatrix} \begin{bmatrix} \mathbf{H}_2^* & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_1^* \end{bmatrix} \begin{bmatrix} \mathbf{0} \\ \mathbf{s}_1^{\#} \end{bmatrix} \quad (7)$$

$$= \begin{bmatrix} k_1 \mathbf{G}_1 \mathbf{H}_1 \mathbf{s}_1 + k_2 \mathbf{G}_2^* \mathbf{H}_1^* \mathbf{s}_1^{\#} \\ k_1 \mathbf{G}_2 \mathbf{H}_2 \mathbf{s}_1 + k_2 \mathbf{G}_1^* \mathbf{H}_2^* \mathbf{s}_1^{\#} \end{bmatrix}$$

where the $\mathbf{y}_{11}$ and $\mathbf{y}_{12}$ are the positive and negative part received signal of $\mathbf{y}_1$, respectively. In (7), it can be observed that $\mathbf{y}_1$ composes of one real part and one image part received signal via the nulling design of preamble.

Next, similarly, substituting preamble 2 into $\mathbf{y}$, $\mathbf{y}_2$ can be derived by

$$\mathbf{y}_2 = \begin{bmatrix} \mathbf{y}_{21} \\ \mathbf{y}_{22} \end{bmatrix} = \begin{bmatrix} k_1 \mathbf{G}_1 \mathbf{H}_1 \mathbf{s}_1 e^{j2\pi\epsilon_f} + k_2 \mathbf{G}_2^* \mathbf{H}_1^* \mathbf{s}_1^* e^{-j2\pi\epsilon_f} \\ k_1 \mathbf{G}_3 \mathbf{H}_1 \mathbf{s}_1 e^{j2\pi\epsilon_f} + k_2 \mathbf{G}_1^* \mathbf{H}_1^* \mathbf{s}_1^* e^{-j2\pi\epsilon_f} \end{bmatrix} \quad (8)$$

where ϵ_f is the normalized carrier frequency offset, i.e., $\epsilon_f = NT_s \Delta f$. $\mathbf{y}_{21}$ and $\mathbf{y}_{22}$ are the positive and the negative part received signal of $\mathbf{y}_2$. Then, using the known property of I/Q imbalance and CFO effects, i.e., $k_2 < k_1$ and $\mathbf{G}_2 < \mathbf{G}_1$, the $\mathbf{y}_{11}$ and $\mathbf{y}_{21}$ can be approached by

$$\mathbf{y}_{11} \cong \tilde{\mathbf{y}}_{11} = k_1 \mathbf{G}_1 \mathbf{H}_1 \mathbf{s}_1 \quad (9)$$

$$\mathbf{y}_{21} \cong \tilde{\mathbf{y}}_{21} = k_1 \mathbf{G}_1 \mathbf{H}_1 \mathbf{s}_1 e^{j2\pi\epsilon_f} \quad (10)$$

Moreover, using the auto-correlation scheme between $\tilde{\mathbf{y}}_{11}$ and $\tilde{\mathbf{y}}_{21}$, the CFO parameter can be estimated by

$$\hat{\epsilon}_f = \frac{1}{2\pi} \angle \tilde{\mathbf{y}}_{11}^H \tilde{\mathbf{y}}_{21} \quad (11)$$

Based on the estimated normalized CFO $\hat{\epsilon}_f$, we can get the frequency-domain CFO matrix, i.e.,

$$\hat{\Delta f} = \hat{\epsilon}_f / (NT_s) \quad (12)$$

$$\hat{E}(n) = \exp(j2\pi\hat{\Delta f}nT_s) \quad (13)$$

$$\hat{\mathbf{G}} = \mathbf{F} \text{diag}\{\hat{E}(n)\} \mathbf{F}^H \quad (14)$$

Due to the estimated CFO matrix with accurate performance, so it can assist us to estimate I/Q imbalance and channel parameters.

C. I/Q Imbalance and Channel Estimation

The following development considers the design of I/Q imbalance and channel estimation shown in Figure 1. As seen in (14), the CFO matrix $\hat{\mathbf{G}}$ can be estimated. Then, the post-FFT received signal can be compensated by the original and conjugate CFO estimated matrices, i.e., $\hat{\mathbf{G}}$ and $\hat{\mathbf{G}}^*$, to acquire the composite signal $\mathbf{z}_1$ and $\mathbf{z}_2$ in Figure 1. First, after the conjugate CFO compensation, the compensated signal can be derived by

$$\begin{aligned} \bar{\mathbf{y}}_1 &= \hat{\mathbf{G}}^* \mathbf{y}_1 \\ &= k_1 \hat{\mathbf{G}}^* \mathbf{G} \mathbf{H} \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{0} \end{bmatrix} + k_2 \hat{\mathbf{G}}^* \mathbf{G}^* \mathbf{H}^* \begin{bmatrix} \mathbf{0} \\ \mathbf{s}_1^* \end{bmatrix} \\ &= k_1 \Delta \bar{\mathbf{G}} \mathbf{H} \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{0} \end{bmatrix} + k_2 \bar{\mathbf{G}} \mathbf{H}^* \begin{bmatrix} \mathbf{0} \\ \mathbf{s}_1^* \end{bmatrix} \end{aligned}$$

$$\begin{aligned} &= k_1 \begin{bmatrix} \Delta \bar{\mathbf{G}}_1 \Delta \bar{\mathbf{G}}_2 \\ \Delta \bar{\mathbf{G}}_2 \Delta \bar{\mathbf{G}}_1 \end{bmatrix} \begin{bmatrix} \mathbf{H}_1 & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_2 \end{bmatrix} \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{0} \end{bmatrix} \\ &\quad + k_2 \begin{bmatrix} \bar{\mathbf{G}}_1 \bar{\mathbf{G}}_2 \\ \bar{\mathbf{G}}_2 \bar{\mathbf{G}}_1 \end{bmatrix} \begin{bmatrix} \mathbf{H}_2^* & \mathbf{0} \\ \mathbf{0} & \mathbf{H}_1^* \end{bmatrix} \begin{bmatrix} \mathbf{0} \\ \mathbf{s}_1^* \end{bmatrix} \\ &= \begin{bmatrix} k_1 \Delta \bar{\mathbf{G}}_1 \mathbf{H}_1 \mathbf{s}_1 + k_2 \bar{\mathbf{G}}_2 \mathbf{H}_1^* \mathbf{s}_1^* \\ k_1 \Delta \bar{\mathbf{G}}_2 \mathbf{H}_1 \mathbf{s}_1 + k_2 \bar{\mathbf{G}}_1 \mathbf{H}_1^* \mathbf{s}_1^* \end{bmatrix} \end{aligned} \quad (15)$$

where $\Delta \bar{\mathbf{G}} = \hat{\mathbf{G}}^* \mathbf{G} = \begin{bmatrix} \Delta \bar{\mathbf{G}}_1 \Delta \bar{\mathbf{G}}_2 \\ \Delta \bar{\mathbf{G}}_2 \Delta \bar{\mathbf{G}}_1 \end{bmatrix}$ and $\bar{\mathbf{G}} = \hat{\mathbf{G}}^* \mathbf{G}^* = \begin{bmatrix} \bar{\mathbf{G}}_1 \bar{\mathbf{G}}_2 \\ \bar{\mathbf{G}}_2 \bar{\mathbf{G}}_1 \end{bmatrix}$ are the residual CFO matrices of the positive and negative subcarriers, respectively. Note that if the CFO estimation is more accurate, $\Delta \bar{\mathbf{G}}$ will approach to the identity matrix and $\bar{\mathbf{G}}$ will be close to the double carrier frequency offset matrix. Moreover, using the known property of $k_2 < k_1$, $\Delta \mathbf{G}_2 < \Delta \mathbf{G}_1$, and $\bar{\mathbf{G}}_2 < \bar{\mathbf{G}}_1$, the compensated signal in (15) can ignore some smaller terms. Thus, the positive and negative parts of the compact compensated signal can be expressed by $\tilde{\mathbf{y}}_{11}$ and $\tilde{\mathbf{y}}_{12}$, respectively:

$$\begin{bmatrix} \tilde{\mathbf{y}}_{11} \\ \tilde{\mathbf{y}}_{12} \end{bmatrix} \cong \begin{bmatrix} k_1 \Delta \bar{\mathbf{G}}_1 \mathbf{H}_1 \mathbf{s}_1 \\ k_2 \bar{\mathbf{G}}_1 \mathbf{H}_1^* \mathbf{s}_1^* \end{bmatrix} \quad (16)$$

Second, similar to the previous CFO compensation, but it is compensated by the original estimated CFO matrix $\hat{\mathbf{G}}$. It will assist us to acquire two approaching matrices, i.e., the identity and double CFO matrices at negative and positive subcarriers part, respectively. Therefore, the compensated signal can be expressed by

$$\begin{aligned} \bar{\bar{\mathbf{y}}}_1 &= \hat{\mathbf{G}} \mathbf{y}_1 \\ &= k_1 \bar{\mathbf{G}}^* \mathbf{H} \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{0} \end{bmatrix} + k_2 \Delta \bar{\mathbf{G}}^* \mathbf{H}^* \begin{bmatrix} \mathbf{0} \\ \mathbf{s}_1^* \end{bmatrix} \end{aligned} \quad (17)$$

In (17), with k_2 , $\Delta \bar{\mathbf{G}}_2$, and $\bar{\mathbf{G}}_2$ smaller properties, the received signal of the positive and negative part subcarriers can be approached by

$$\begin{bmatrix} \bar{\bar{\mathbf{y}}}_{11} \\ \bar{\bar{\mathbf{y}}}_{12} \end{bmatrix} \cong \begin{bmatrix} k_1 \bar{\mathbf{G}}_1^* \mathbf{H}_1 \mathbf{s}_1 \\ k_1 \bar{\mathbf{G}}_2^* \mathbf{H}_1 \mathbf{s}_1 + k_2 \Delta \bar{\mathbf{G}}_1^* \mathbf{H}_1^* \mathbf{s}_1^* \end{bmatrix} \quad (18)$$

Because $k_1 \bar{\mathbf{G}}_2^*$ is not obviously smaller than $k_2 \Delta \bar{\mathbf{G}}_1^*$, $\bar{\bar{\mathbf{y}}}_{12}$ cannot ignore any compensated terms.

Next, we are interesting the negative compensated signal $\bar{\bar{\mathbf{y}}}_{12}$ in (16) and the positive compensated signal $\tilde{\mathbf{y}}_{11}$ in (18), which can be used to estimate the I/Q imbalance and channel parameters. That is, first taking the complex conjugate and reversing operation of $\tilde{\mathbf{y}}_{12}$, we can get $\tilde{\mathbf{y}}_{12}^* = k_2^* \bar{\mathbf{G}}_1^* \mathbf{H}_1^* \mathbf{s}_1^*$. Then,

using the addition and subtraction between $\tilde{\mathbf{y}}_{12}^\#$ and $\tilde{\mathbf{y}}_{11}$, the composite signal $\mathbf{z}_1$ and $\mathbf{z}_2$ can be acquired by

$$\begin{aligned}\mathbf{z}_1 &= \tilde{\mathbf{y}}_{11} + \tilde{\mathbf{y}}_{12}^\# \\ &= (k_1 + k_2^*) \bar{\mathbf{G}}_1 \mathbf{H}_1 \mathbf{s}_1\end{aligned}\quad (19)$$

$$\begin{aligned}\mathbf{z}_2 &= \tilde{\mathbf{y}}_{11} - \tilde{\mathbf{y}}_{12}^\# \\ &= (k_1 - k_2^*) \bar{\mathbf{G}}_1 \mathbf{H}_1 \mathbf{s}_1 \\ &= \mu \bar{\mathbf{G}}_1^* \mathbf{H}_1 \mathbf{s}_1\end{aligned}\quad (20)$$

Note that, using the different CFO compensated techniques in (15) and (17), the estimated parameters can be separated, i.e., I/Q imbalance parameter μ being easily determined by $\mathbf{z}_1$ and $\mathbf{z}_2$. Moreover, the I/Q imbalance parameter can be estimated by the cross-correlation and auto-correlation of $\mathbf{z}_1$ and $\mathbf{z}_2$, i.e.,

$$\hat{\mu}_1 = \mathbf{z}_1^H \mathbf{z}_2 / \|\mathbf{z}_1\|^2 \quad (21)$$

Next, referring the composite signal $\mathbf{z}_1$ in (19), the positive subcarriers part channel matrix $\mathbf{H}_1$ can be estimated by the preamble sequence $\mathbf{s}_1(k)$ and the double CFO matrix $\bar{\mathbf{G}}_1$, i.e.,

$$\hat{\mathbf{H}}_1 = \text{diag}\{\text{diag}\{\mathbf{s}_1\}^{-1} \bar{\mathbf{G}}_1 \mathbf{z}_1\} \quad (22)$$

Furthermore, in order to estimate the negative subcarriers part channel matrix $\mathbf{H}_2$, the preamble 3 design with positive subcarriers nulling scheme is proposed and shown in Figure 2. Therefore, the same estimated procedures in (15)-(22) for preamble 1 can be applied to the estimator of preamble 3.

First, substituting preamble 3 into $\mathbf{y}$, the received signal $\mathbf{y}_3$ can be derived by

$$\mathbf{y}_3 = \begin{bmatrix} k_1 \mathbf{G}_2 \mathbf{H}_2 \mathbf{s}_1 e^{j4\pi\hat{\epsilon}_f} + k_2 \mathbf{G}_1^* \mathbf{H}_2^\# \mathbf{s}_1^\# e^{-j4\pi\hat{\epsilon}_f} \\ k_1 \mathbf{G}_1 \mathbf{H}_2 \mathbf{s}_1 e^{j4\pi\hat{\epsilon}_f} + k_2 \mathbf{G}_2^* \mathbf{H}_2^\# \mathbf{s}_1^\# e^{-j4\pi\hat{\epsilon}_f} \end{bmatrix} \quad (23)$$

Then, using the estimated CFO $\hat{\epsilon}_f$ in (13) and CFO matrix $\hat{\mathbf{G}}$ in (14), the original and conjugate CFO compensation for the received signal $\mathbf{y}_3$ can be expressed by

$$\bar{\mathbf{y}}_3 = \hat{\mathbf{G}}^* \mathbf{y}_3 e^{-j4\pi\hat{\epsilon}_f} \quad (24)$$

$$\bar{\bar{\mathbf{y}}}_3 = \hat{\mathbf{G}} \mathbf{y}_3 e^{j4\pi\hat{\epsilon}_f} \quad (25)$$

Note that, similar to preamble 1, we are interesting to the positive part signal $\tilde{\mathbf{y}}_{31}$ of $\bar{\mathbf{y}}_3$ and the negative part signal $\tilde{\bar{\mathbf{y}}}_{32}$ of $\bar{\bar{\mathbf{y}}}_3$, which can assist us to estimate the I/Q imbalance and channel matrix. Thus, based on (23)-(25) and the ignoring operation in (16) and (18), $\tilde{\mathbf{y}}_{31}^\#$ and $\tilde{\bar{\mathbf{y}}}_{32}$ can be derived by

$$\tilde{\mathbf{y}}_{31}^\# = k_2^* \bar{\mathbf{G}}_1^* \mathbf{H}_2 \mathbf{s}_1 \quad (26)$$

$$\tilde{\bar{\mathbf{y}}}_{32} = k_1 \bar{\mathbf{G}}_1^* \mathbf{H}_2 \mathbf{s}_1 \quad (27)$$

By the addition and subtraction of $\tilde{\mathbf{y}}_{31}^\#$ and $\tilde{\bar{\mathbf{y}}}_{32}$, the I/Q imbalance parameter can be separated by $\mathbf{z}_3$ and $\mathbf{z}_4$, i.e.,

$$\mathbf{z}_3 = \tilde{\bar{\mathbf{y}}}_{32} + \tilde{\mathbf{y}}_{31}^\# = \bar{\mathbf{G}}_1^* \mathbf{H}_2 \mathbf{s}_1 \quad (28)$$

$$\mathbf{z}_4 = \tilde{\bar{\mathbf{y}}}_{32} - \tilde{\mathbf{y}}_{31}^\# = \mu \bar{\mathbf{G}}_1^* \mathbf{H}_2 \mathbf{s}_1 \quad (29)$$

Thus, the parameter of I/Q imbalance can be estimated again by the cross-correlation and auto-correlation of $\mathbf{z}_3$ and $\mathbf{z}_4$, i.e.,

$$\hat{\mu}_2 = \mathbf{z}_3^H \mathbf{z}_4 / \|\mathbf{z}_3\|^2 \quad (30)$$

Moreover, the negative subcarriers part channel matrix $\mathbf{H}_2$ can be estimated by the preamble sequence $\mathbf{s}_1$ and the double CFO matrix $\bar{\mathbf{G}}_1$, i.e.,

$$\hat{\mathbf{H}}_2 = \text{diag}\{\text{diag}\{\mathbf{s}_1\}^{-1} \bar{\mathbf{G}}_1 \mathbf{z}_3\} \quad (31)$$

By taking the average of $\hat{\mu}_1$ and $\hat{\mu}_2$, the accurate estimated parameter of I/Q imbalance can be obtained by

$$\hat{\mu} = \frac{1}{2}(\hat{\mu}_1 + \hat{\mu}_2) \quad (32)$$

For channel estimation, cascading the estimated frequency-domain channel matrix of positive and negative subcarriers part, the complete channel matrix can be acquired by

$$\hat{\mathbf{H}} = \begin{bmatrix} \hat{\mathbf{H}}_1 & \mathbf{0} \\ \mathbf{0} & \hat{\mathbf{H}}_2 \end{bmatrix} \quad (33)$$

D. Enhancement of Channel Estimation

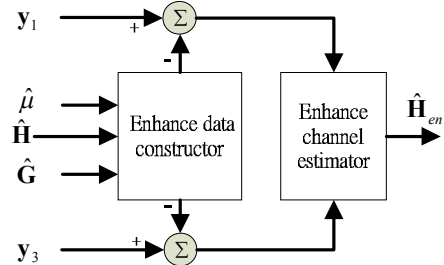


Figure 3. Block diagram of enhanced channel estimation

In order to enhance the channel estimation, the negative subcarriers part of the received signal $\mathbf{y}_1$ can be cancelled by the estimated parameters, which is shown in Figure 3. Thus, the received signal with image cancellation of preamble 1 can be expressed by

$$\begin{aligned}\mathbf{y}_{en1} &= \left(\mathbf{y}_1 - \hat{k}_2 \hat{\mathbf{G}}^* \hat{\mathbf{H}}^\# \begin{bmatrix} \mathbf{0} \\ \mathbf{s}_1^\# \end{bmatrix} \right) / \hat{k}_1 \\ &\cong \mathbf{G} \mathbf{H} \begin{bmatrix} \mathbf{s}_1 \\ \mathbf{0} \end{bmatrix}\end{aligned}\quad (34)$$

Similarly, the received signal with image cancellation of preamble 3 can be given by

$$\mathbf{y}_{en3} \equiv \mathbf{G}\mathbf{H} \begin{bmatrix} \mathbf{0} \\ \mathbf{s}_1 \end{bmatrix} \quad (35)$$

Next, using the post-cancelled received signal, preamble sequence, and CFO estimated matrix, the positive and negative part channel matrices can be estimated by

$$\hat{\mathbf{H}}_{en1} = \text{diag} \left\{ \text{diag} \{ \mathbf{s}_1 \}^{-1} \hat{\mathbf{G}}^* \mathbf{y}_{en1} \right\} \quad (36)$$

$$\hat{\mathbf{H}}_{en2} = \text{diag} \left\{ \text{diag} \{ \mathbf{s}_1 \}^{-1} \hat{\mathbf{G}}^* \mathbf{y}_{en3} \right\} \quad (37)$$

Thus, the complete CFO matrix can be given by

$$\hat{\mathbf{H}}_{en} = \begin{bmatrix} \hat{\mathbf{H}}_{en1} & \mathbf{0} \\ \mathbf{0} & \hat{\mathbf{H}}_{en2} \end{bmatrix} \quad (38)$$

IV. SIMULATION RESULTS

In the section, simulation results are conducted to demonstrate the performance of the proposed estimators. For simulation parameters, the number of subcarriers and the preamble Chu sequence are chosen to be 64 and 32, respectively. The $L=4$ independent Rayleigh fading paths are generated by the delay chosen from $[0, 3T_s]$ with T_s being sample duration, which is smaller than the guard interval. A quasi-static fading channel is assumed and the fading gains are i.i.d. complex Gaussian random variable with a zero-mean and unit variance. Also, the normalized frequency offset is assumed to be $\varepsilon_f = 0.1$. The amplitude and phase mismatch of I/Q imbalance are chosen to be $g=1.3$ and $\phi = 10^\circ$ and 20° . Note that all I/Q imbalance and CFO parameters are assumed to be time-invariant.

In the first set of simulations, the mean square errors (MSE) of the estimated parameters, i.e., CFO, channel, I/Q imbalance, are evaluated as a function of E_b/N_0 for the proposed estimators in Section III under different gain and phase scenarios. The results shown in Figure 4 indicate that the proposed CFO estimator can provide the robust performance over the serious mismatches ($g=1.3$ and $\phi = 20^\circ$) of I/Q imbalance. Next, Figure 4(b) also shows the reliable MSE performance of CFO estimator as E_b/N_0 increases. For the worse case of I/Q imbalance, it only degrades about 1 dB than the ideal case scenario. Then, for the MSE performance of channel estimator, the results given in Figure 4(c) show that the worse case of I/Q imbalance provides the error floor degradation at $E_b/N_0 > 25$ dB, which is due to the image signal interference.

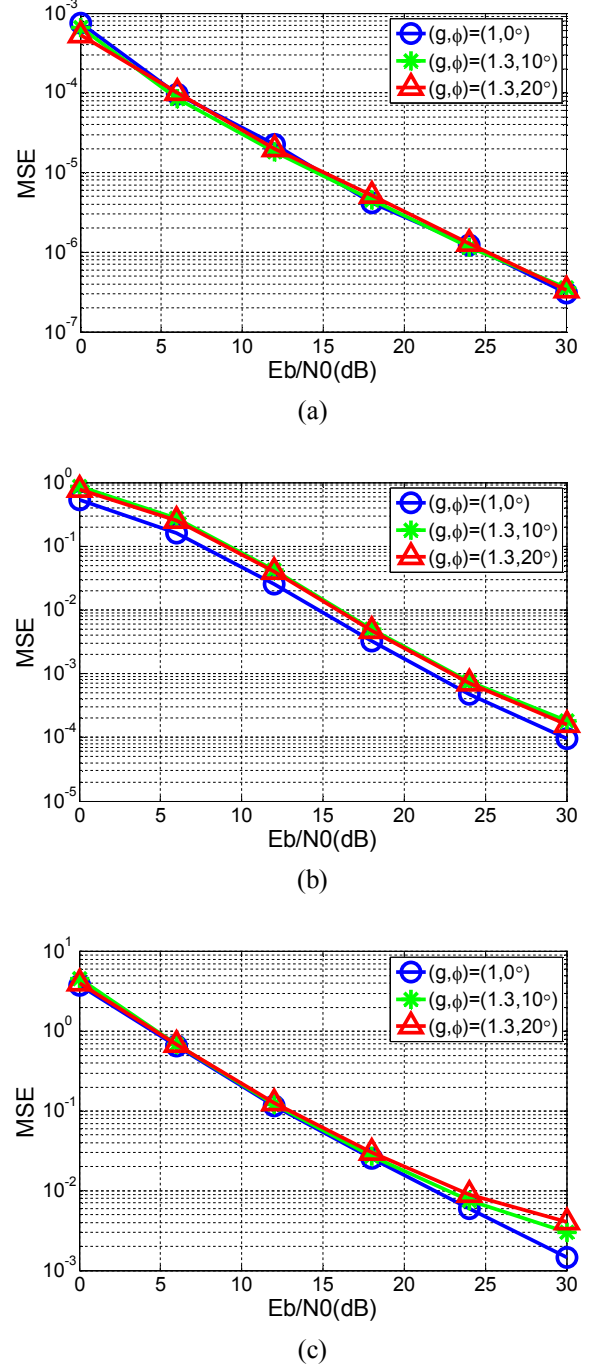


Figure 4. MSE versus E_b/N_0 for proposed estimators (a) CFO (b) I/Q imbalance (c) channel, with different I/Q imbalance and CFO $\varepsilon_f = 0.1$.

In the second set of simulations, the BER performance of the proposed receiver is evaluated at a function of E_b/N_0 . As shown in Figure 5, the ideal result is simulated under perfect parameters assumption. Moreover, the proposed receiver is degraded about 3 dB loss than the ideal bound. Note that, comparing to the MSE error floor performance of channel estimator in Figure 4(c), BER performance of the proposed receiver in Figure 5 only shows a little degradation.

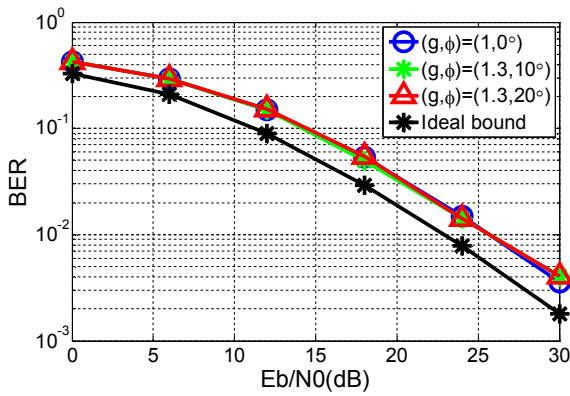


Figure 5. BER versus E_b/N_0 for proposed receiver with different I/Q imbalance and CFO $\varepsilon_f = 0.1$.

In the third set of simulations, to investigate the robust performance of the proposed method, the BER performance as a function of phase and gain mismatches is shown in Figures 6 and 7. As observed, the proposed receiver can provide the reliable BER performance under different phase and gain mismatch effects. Note that the performance shows a slight degradation at $E_b/N_0 = 30\text{dB}$, phase mismatch $\geq 30^\circ$, and gain mismatch ≥ 1.4 due to the image interference of the severe I/Q imbalance effect.

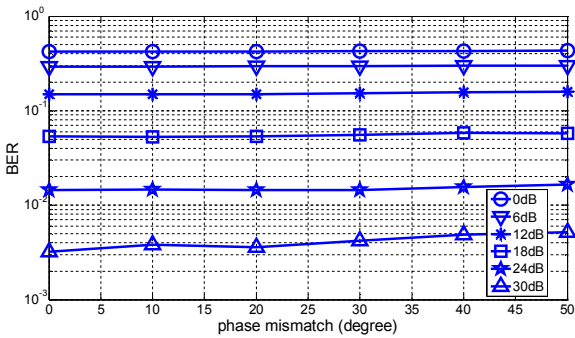


Figure 6. BER versus phase mismatch for proposed receiver with different E_b/N_0 , gain mismatch $g = 1.3$, and CFO $\varepsilon_f = 0.1$.

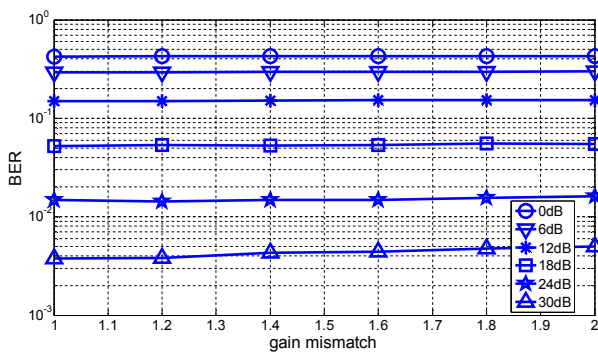


Figure 7. BER versus gain mismatch for proposed receiver with different E_b/N_0 , phase mismatch $\phi = 20^\circ$, and CFO $\varepsilon_f = 0.1$.

V. CONCLUSIONS

A low complexity time-frequency-multiplex preamble design is proposed which is suitable for OFDM systems over

frequency selective multipath channels with severe I/Q imbalance and CFO environment. Using the simple time-frequency-multiplex design, it can separate the joint parameters. Then, the low computational complexity technique, i.e., addition, subtraction, and correlation, can assist the receiver to estimate parameters individually. From simulation results, it is shown that the proposed estimators offer robust performance with tolerance to large I/Q imbalance and resistance to strong image interference.

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