Estimation of I/Q Imblance in Mimo OFDM System

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Abstract—In this paper, we study the joint estimation of in phase and quadrature-phase (I/Q) imbalance, carrier frequency offset (CFO), and channel response for multiple-input multiple output (MIMO) orthogonal frequency division multiplexing (OFDM) systems using training sequences. A new concept called channel residual energy (CRE) is introduced. We show that by minimizing the CRE, we can jointly estimate the I/Q imbalance and CFO without knowing the channel response. The proposed method needs only one OFDM block for training and the training symbols can be arbitrary. Moreover, when the training block consists of two repeated sequences, a low complexity two-step approach is proposed to solve the joint estimation problem. Simulation results show that the mean-squared error (MSE) of the proposed method is close to the Cramer-Rao bound (CRB).

Index Terms—MIMO OFDM, CFO, I/Q imbalance, channel estimation.

I. INTRODUCTION

In recent years, direct conversion receiver has drawn a lot of attention due to its low power consumption and low implementation cost. However, some mismatches in direct conversion receiver can seriously degrade the system performance, such as in-phase and quadrature-phase (I/Q) imbalance and carrier frequency offset (CFO). The I/Q imbalance is due to the amplitude and phase mismatches between the I and Q-branch of the local oscillator whereas the CFO is due to the mismatch of carrier frequency at the transmitter and receiver. It is known that the I/Q imbalance and CFO can cause a serious inter-carrier interference (ICI) in orthogonal frequency division multiplexing (OFDM) systems. As a result, the bit error rate (BER) has an error-flooring. There have been many reports in the literature on the compensation of the I/Q imbalance and CFO. Several compensation methods for I/Q imbalance in OFDM systems have been proposed.

II. SYSTEM DESCRIPTION

In MIMO OFDM system where the numbers of the transmit and receive antenna are $N_t$and $N_r$respectively. The input vector $\mathbf{s}_j$(see Fig. 1) is an $M\times 1$ vector containing the modulation symbols. After taking the M-point IDFT of $\mathbf{s}$, we obtain the $M\times$ 1 vector $\mathbf{x}_j$. After the insertion of a CP of length $L-$ 1, the signal is transmitted from the jth transmit antenna. Let the channel impulse response from the jth transmit antenna to the kth receive antenna be $h_{k,j}$ (n). We assume that the lengths of all the channels are $\le L$and the length of the cyclic prefix (CP) is $L-$ 1. So there is no interblock interference between adjacent OFDM blocks afterCP removal. The received vector at the kth receive antenna can be written as

$$\mathbf{r}_k = \begin{bmatrix} \mathbf{H}_{k,0} & \mathbf{H}_{k,1} & \cdots & \mathbf{H}_{k,N_t-1} \end{bmatrix} \begin{bmatrix} \mathbf{x}_0 \\ \mathbf{x}_1 \\ \vdots \\ \mathbf{x}_{N_t-1} \end{bmatrix} + \mathbf{q}_k \quad (1)$$

Where $\mathbf{H}_{k,j}$ is an $M\times M$circulant matrix with the first column

$$\mathbf{h}_{k,j} \triangleq \left[\begin{array}{cccccc} h_{k,j}(0) & \cdots & h_{k,j}(L-1) & 0 & \cdots & 0 \end{array} \right]^T, \quad (2)$$

and $\mathbf{q}_k$ is the $M \times 1$ blocked version of channel noise. After passing $\mathbf{r}_k$ through the M-point DFT, we can employ a frequency domain equalizer (FEQ) to recover the transmit signal **s**.
Suppose now that the system suffers from carrier frequency offset (CFO) Δf_k. Define the normalized CFO as

$$\theta_k \triangleq \frac{\Delta f_k}{\frac{1}{MT}} = \Delta f_k MT, \quad (3)$$

Where M is the size of the DFT matrix and T is the sample spacing. The vector due to CFO is

$$\mathbf{y}_k = \mathbf{E}_k \mathbf{r}_k, \quad (4)$$

Where $\mathbf{r}_k$ is the desired baseband vector in (1) and $\mathbf{E}_k$ is an $M \times M$ diagonal matrix

$$\mathbf{E}_k \triangleq diag \left[\begin{array}{cccc} 1 & e^{j\frac{2\pi}{M}\theta_k} & \cdots & e^{j\frac{2\pi}{M}(M-1)\theta_k} \end{array} \right]. \quad (5)$$

Suppose in addition to the CFO, there is also I/Q mismatch at the receiver. The received vector due to I/Q mismatch becomes

$$\mathbf{z}_k = \mu_k \mathbf{y}_k + \nu_k \mathbf{y}_k^*, \quad (6)$$

Where μ_k and ν_k are the I/Q parameters at thereceiver. They are related to the amplitude mismatch ϵ_k and phase mismatch ϕ_k as

$$\mu_k \triangleq \frac{1+\epsilon_k e^{-j\phi_k}}{2} \text{ and } \nu_k \triangleq \frac{1-\epsilon_k e^{j\phi_k}}{2}. \quad (7)$$

Substituting (4) into (6), we get

$$\mathbf{z}_k = \mu_k \mathbf{E}_k \mathbf{r}_k + \nu_k \mathbf{E}_k^* \mathbf{r}_k^*. \quad (8)$$

The received vector $\mathbf{z}_k$ consists of not only the desired baseband vector $\mathbf{r}_k$ but also its complex conjugate $\mathbf{r}_k k$. Moreover, the presence of $\mathbf{E}_k$ due to CFO will also destroy the subcarrier orthogonality. In later sections, we will show how to jointly estimate the I/Q imbalance, CFO and MIMO channel response using training sequences. Suppose that we have estimates of the I/Q imbalance and CFO at the receiver. We will show how to recover the desired baseband vector $\mathbf{r}_k$ from $\mathbf{z}_k$. Define a parameter α_k that is related to the I/Q imbalance parameters as

$$\alpha_k \triangleq \frac{\nu_k}{\mu_k^*}. \quad (9)$$

If α_k is known at the receiver, from (6) we can get

$$\mu_k \mathbf{y}_k = \frac{\mathbf{z}_k - \alpha_k \mathbf{z}_k^*}{1-|\alpha_k|^2}. \quad (10)$$

If θ_k is also known at the receiver, from (4) we can recover a scaled version of the desired baseband vector by

$$\mu_k \mathbf{r}_k = \mathbf{E}_k^* \mu_k \mathbf{y}_k = \mathbf{E}_k^* \frac{\mathbf{z}_k - \alpha_k \mathbf{z}_k^*}{1-|\alpha_k|^2}. \quad (11)$$

III. PROPOSED JOINT ESTIMATION METHOD

In this section, we propose a new method to estimate the channel response when there are CFO and I/Q imbalances. We will first consider the simpler problem of the joint estimation of channel response and I/Q imbalance under the assumption that there is no CFO. In this special case, the optimal solution is given in closed form. Then the joint estimation of the channel response, CFO and I/Q imbalance will be studied. Below we will show how to estimate $\alpha k$and $\theta k$from one received vector $\mathbf{z}k$at the kth receive antenna. For notational simplicity, we will drop the receive antenna index $k$as the problem can be solved separately for each receive antenna.

A. Joint Estimation of Channel Response and I/Q Imbalance

In this subsection, we assume that there is no CFO. Hence we have θ= 0 and $\mathbf{E} = \mathbf{I}$. From (11), $\mu\mathbf{r}$ is related to the received vector $\mathbf{z}$ as

$$\mu\mathbf{r} = \frac{\mathbf{z} - \alpha\mathbf{z}^*}{1 - |\alpha|^2}. \tag{12}$$

From (23) and (29), if $\alpha$is given, an estimate of the MIMO channel response can be obtained as

$$\begin{aligned} \mu\widehat{\mathbf{d}} &= \mu\left[\begin{array}{cccc} \widehat{\mathbf{d}}_0^T & \widehat{\mathbf{d}}_1^T & \cdots & \widehat{\mathbf{d}}_{N_t-1}^T \end{array}\right]^T \\ &= \mathbf{B}^{-1}\mu\mathbf{r} = \mathbf{B}^{-1}\frac{\mathbf{z} - \alpha\mathbf{z}^*}{1 - |\alpha|^2}, \end{aligned} \tag{13}$$

Where$\mathbf{B}$ is defined in (22). When $\alpha$is estimated perfectly, the first Lentries of each $\hat{\mathbf{h}}l$in the above expression will give us an estimate of the channel response and the last ($\rho-L$) entriesof $\hat{\mathbf{d}}l$are solely due to the channel noise. For moderately high SNR, the energy of these entries should be small. Let us define a quantity called the channel residual energy (CRE) as

$$\mathrm{CRE} \triangleq \sum_{l=0}^{N_t-1}\sum_{i=L}^{\rho-1} |[\mu\widehat{\mathbf{d}}_l]_i|^2, \tag{14}$$

Where $[\mu\widehat{\mathbf{d}}_l]_i$ denotes the ith entry of $\mu\widehat{\mathbf{d}}_l$. Any error in the estimation of αwill increase the CRE (see the analysis at the end of this section). Based on this observation, by minimizingthe CRE we are able to estimate the I/Q parameter αwithout knowing the channel response. To do this, we first define the ($M-N_tL$) $\times M$matrix

$$\mathbf{P} = \begin{bmatrix} \mathbf{0} & \mathbf{I}_{\rho-L} & \mathbf{0} & \cdots & & \cdots & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & \mathbf{0} & \mathbf{I}_{\rho-L} & \mathbf{0} & \cdots & \mathbf{0} \\ \vdots & \ddots & & & \cdots & & \vdots \\ \mathbf{0} & \mathbf{0} & \cdots & & \cdots & \mathbf{0} & \mathbf{I}_{\rho-L} \end{bmatrix} \tag{15}$$

Suppose that $\rho>L$so that $\mathbf{P}$ is not a zero matrix. Multiplying $\mu\hat{\mathbf{d}}$ by $\mathbf{P}$, we can rewrite the CRE as

$$\mathrm{CRE} = \|\mathbf{P}\mu\widehat{\mathbf{d}}\|^2 = \left\|\mathbf{P}\mathbf{B}^{-1}\frac{\mathbf{z} - \alpha\mathbf{z}^*}{1 - |\alpha|^2}\right\|^2 \tag{16}$$

Our goal is to find αthat minimizes the CRE. Since for most applications, $\alpha$is small, (33) can be approximated as

$$\mathrm{CRE} \approx \|\mathbf{PB}^{-1}(\mathbf{z} - \alpha\mathbf{z}^*)\|^2. \quad (17)$$

From linear algebra, it is known that the optimal αthat minimizes the CRE is

$$\alpha_{opt} = \frac{(\mathbf{PB}^{-1}\mathbf{z}^*)^\dagger(\mathbf{PB}^{-1}\mathbf{z})}{\|\mathbf{PB}^{-1}\mathbf{z}^*\|^2}. \quad (18)$$

By substituting αoptinto (30), we get the estimated MIMO channel response $\mu\hat{\mathbf{d}}$
For the compensation of I/Q imbalance, one can employ (29) to obtain $\mu\mathbf{r}$. Notice that there is no need to compensate the factor μbecause it will be canceled when we use $\mu\hat{\mathbf{d}}$to implement the FEQ. From (35), we see that to get αopt, we only need to compute $\mathbf{B}^{-1}\mathbf{z}$ and perform vector inner products at the numerator and denominator2. When the training sequence in [9] is used, **B** becomes unitary and circulant. As B^(-1) is also circulant and unitary,$\mathbf{B}^{-1}\mathbf{z}$ can be efficiently realized using circular convolution.

B. Joint Estimation of Channel Response, I/Q Imbalance and CFO

When the receiver suffers from both CFO and I/Q mismatch, the received vector **z** is given by (8). From Sec. 2, we know that if $\theta$and $\alpha$are known, we can recover the desired basebandvector from **z** using (11) and it is given by

$$\mu\mathbf{r} = \mathbf{E}^*\mu\mathbf{y} = \mathbf{E}^*\frac{\mathbf{z} - \alpha\mathbf{z}^*}{1 - |\alpha|^2}, \quad (19)$$

where the diagonal matrix **E** is given in (5). We can obtain an estimate of the MIMO channel response as

$$\mu\hat{\mathbf{d}} = \mathbf{B}^{-1}\mu\mathbf{r} = \mathbf{B}^{-1}\mathbf{E}^*\frac{\mathbf{z} - \alpha\mathbf{z}^*}{1 - |\alpha|^2}. \quad (20)$$

From the above equation, when $\alpha$and $\theta$are perfectly estimated, the last $\rho-L$entries of ˆ**d**$/$are again solely due to the channel noise. By summing up the energy of these entries, we have the CRE

$$\mathrm{CRE}(\alpha, \theta) = \|\mathbf{P}\mu\hat{\mathbf{d}}\|^2, \quad (21)$$

Where**P** is defined in (32). Notice that the CRE is a function o0f both $\alpha$and θ. Substituting (37) into the above equation, we can rewrite the CRE as

$$\mathrm{CRE}(\alpha, \theta) = \left\|\mathbf{F}\frac{\mathbf{z} - \alpha\mathbf{z}^*}{1 - |\alpha|^2}\right\|^2, \quad (22)$$

Where$\mathbf{F} = \mathbf{PB}^{-1}\mathbf{E}^*$. Following the argument in the previous subsection, we get an estimate of $\theta$and $\alpha$by minimizing the CRE. The joint optimization problem is solved in 2 steps: Fora given θ, we derive the optimal α, based on that we optimize θ. Since $|\alpha|^2 \ll 1$, the above expression can be approximated as

$$\mathrm{CRE}(\alpha, \theta) \approx \|\mathbf{Fz} - \alpha \mathbf{Fz}^*\|^2. \quad (23)$$

Given θ, the optimal $\alpha$is given by

$$\alpha_{opt}(\theta) = \frac{(\mathbf{Fz}^*)^\dagger(\mathbf{Fz})}{\|\mathbf{Fz}^*\|^2}. \quad (24)$$

Note that $\alpha(\theta)$ is a function of θbecause **F** depends on θ. Substituting $\alpha op(\theta)$ into (40), the CRE can be written as

$$\mathrm{CRE}(\theta) = \|\mathbf{Fz}\|^2 - \frac{|(\mathbf{Fz})^\dagger(\mathbf{Fz}^*)|^2}{\|\mathbf{Fz}^*\|^2}. \quad (25)$$

Then the optimal estimate of CFO is given by

$$\theta_{opt} = \arg\min_{\theta} \mathrm{CRE}(\theta). \quad (26)$$

Once the optimal $\theta opt$is obtained from the above optimization, the optimal αoptcan be obtained by substituting θoptinto (41) and the estimated channel response is found by substituting $\alpha opt$and θoptinto (37). Then we can use (36) for symbol recovery. Note that no iteration is needed in the above optimization process. However, a one-dimensional search is needed to obtain the CFO estimate. In many practical applications, the training data often consist of repeated sequences. In this case, the one-dimensional search problem in (42) can be avoidedand the joint optimization problem can be solved efficiently using a two-step approach as demonstrated later.

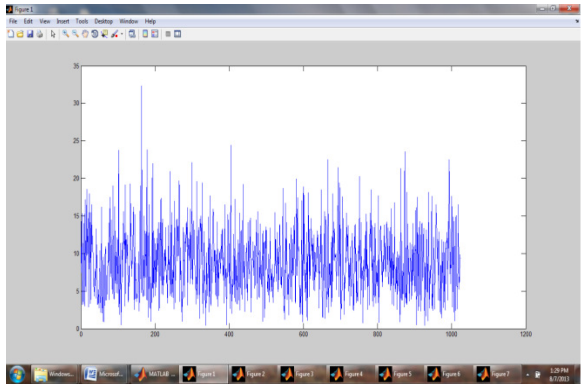
IV. JOINT ESTIMATION USING TWO REPEATED TRAINING SEQUENCES

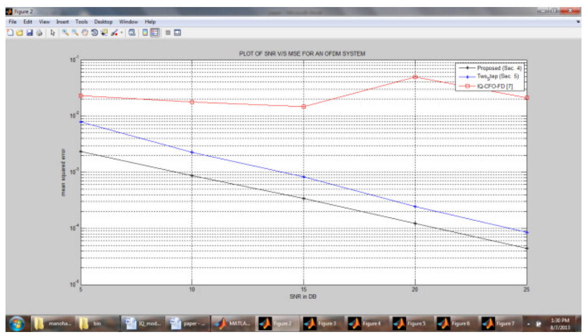
Suppose that two repeated training sequences are available. That means the training block is a $(M+L-1)\times 1$ vector in theform of $[\ \mathrm{CP} \quad \mathbf{x}_i^T \quad \mathbf{x}_i^T\ ]^T$, where the training sequence $\mathbf{x}i$ is an $\frac{M}{2} \times 1$ vector, and the CP length is $L-1$. This repeated structure has been proposed to solve the problem of CFO estimation. In what follows, we will exploit the repeated structure to solve the joint estimation of CFO, I/Q and channel response. Suppose that there are CFO and I/Q mismatch. From(8), the two received $\frac{M}{2} \times 1$ vectors are in the form of

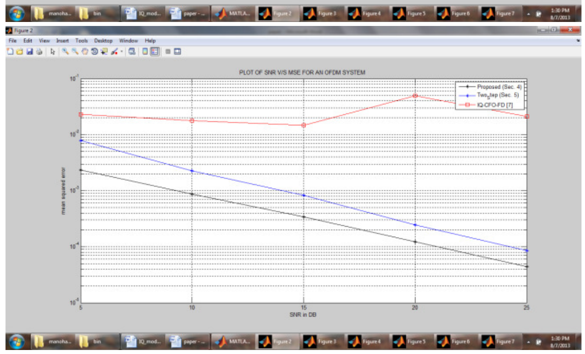
$$\begin{aligned} \mathbf{z}_a &= \mu \mathbf{y} + \nu \mathbf{y}^* + \mathbf{q}_a, \\ \mathbf{z}_b &= \mu e^{j\pi\theta}\mathbf{y} + \nu(e^{j\pi\theta}\mathbf{y})^* + \mathbf{q}_b, \end{aligned} \quad (27)$$

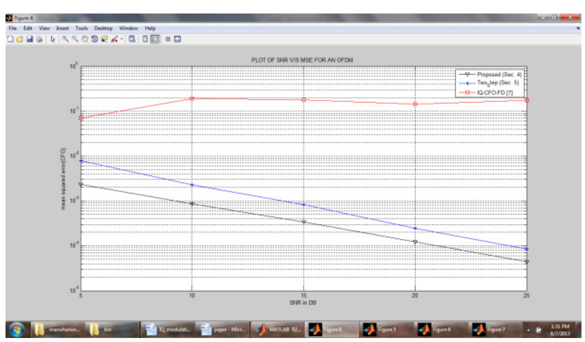
where $a$and $b$are the OFDM block indexes and **y** is an $\frac{M}{2} \times 1$ vector in (4). Our goal is to jointly estimate CFO, I/Q and channel response from $\mathbf{z}a$and $\mathbf{z}b$. Below we will first show how to solve the two sub problems: (A) given α, estimate $\theta$and (B) given θ, estimate $\alpha$and $h(n)$. Then the joint estimation of α, $\theta$and $h(n)$ will be solved by a two-step approach.

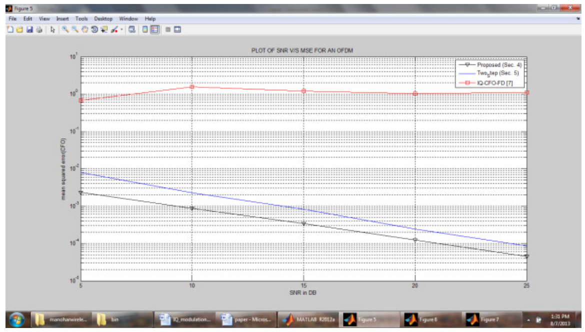
RESULTS:

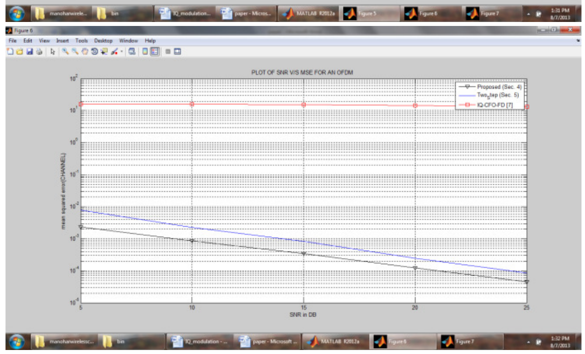



PLOT OF SNR V/S MSE FOR AN OFDM SYSTEM
Proposed (Sec. 4)
Two_step (Sec. 5)
IQ-CFO-FD [7]
SNR in DB


PLOT OF SNR V/S MSE FOR AN OFDM SYSTEM
Proposed (Sec. 4)
Two_step (Sec. 5)
IQ-CFO-FD [7]
SNR in DB


PLOT OF SNR V/S MSE FOR AN OFDM
Proposed (Sec. 4)
Two_step (Sec. 5)
IQ-CFO-FD [7]
SNR in DB





VII. CONCLUSION

In this paper, we propose new methods for the joint estimation of the I/Q imbalance, CFO and channel response for MIMO OFDM systems by using training sequences. When only one OFDM block is available for training, the first method is able to give an accurate estimate of the CFO,I/Q parameter and channel responses. The CFO is obtained through one-dimensional search algorithm. When two repeated OFDM blocks are available for training, a low complexity two step approach is proposed to solve the joint estimation problem.

Simulation results show that the MSEs of the proposed methods are very close to the CRB.

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