OFDM CHANNEL ESTIMATION IN THE PRESENCE OF TRANSMITTER AND RECEIVER I/Q IMBALANCE

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ABSTRACT

In this paper, we consider channel estimation for orthogonal frequency division multiplexing (OFDM) systems when both the transmitter and receiver suffer from in-phase and quadrature-phase (I/Q) imbalance. By exploiting the fact that the block size of an OFDM system is usually larger than the channel order, we propose a new method which can jointly estimate the transmitter and receiver I/Q mismatch and channel response. The estimates of the transmitter and receiver I/Q imbalance parameters are given in a closed form. Using only one arbitrary OFDM block for training, the proposed method can accurately estimate these parameters and a very good performance can be achieved. Simulation results show that the bit error rate (BER) performance of the proposed method is very close to the ideal case where the I/Q mismatch and channel response are perfectly known at the receiver.

Index Terms— OFDM, channel estimation, I/Q imbalance

1. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) transmission scheme has been adopted in many standards, such as digital broadcasting, wireless local area network and digital subscriber loop transmissions, and etc. By inserting a cyclic prefix (CP) that is longer than the channel order, the orthogonality of subcarriers can be preserved even when the channel is frequency selective. Simple channel equalization can be employed at the receiver for symbol recovery. These advantages of OFDM systems are based on the assumption that the system parameters between the transmitter and receiver are perfectly matched. When there is a mismatch, the system performance can be severely degraded. One of the mismatch is the so called I/Q (in-phase and quadrature-phase) imbalance mismatch caused by the local oscillator [1].

In recent years, there are a lot of interests in the compensation of in-phase and quadrature-phase (I/Q) imbalance for orthogonal frequency division multiplexing (OFDM) systems. The I/Q mismatch problem of the local oscillator at the receiver has been studied in [2]-[5]. A blind source separation method was proposed in [2] for the compensation of the receiver I/Q mismatch for Low-IF receivers. For direct-conversion receiver, the I/Q mismatch problem has been investigated in [3]-[5]. Several compensation methods of the receiver I/Q mismatch for OFDM systems were proposed in [3]. The authors in [4] introduced a maximum likelihood (ML) method for the joint estimation of carrier frequency offset (CFO), I/Q imbalance, channel response and DC offset. By assuming that the channel frequency response of an OFDM system is smooth, the joint estimation of the channel frequency response and I/Q mismatch was proposed in [5]. Both the methods in [4][5] need only one training block for joint estimation and can achieve a very good performance. Recently, a new approach was proposed in [6] for the joint estimation of channel response and receiver I/Q imbalance in OFDM systems. By exploiting the fact that the block size is usually larger than the channel order, it was shown in [6] that the time-domain method can accurately estimate both the receiver I/Q imbalance and the channel response with only one OFDM training block. This method was applied to the case of MIMO OFDM systems in [7] and to the case of joint estimation of receiver I/Q mismatch, DC offset and channel response in [8].

In addition to the receiver I/Q mismatch, there can also be I/Q mismatch of the local oscillator at the transmitter. The effects of the transmitter I/Q mismatch on the channel estimation was investigated in [9]. An adaptive method for the joint compensation of transmitter and receiver I/Q imbalance was proposed in [10]. A least-squares method for the joint compensation of the transmitter and receiver I/Q mismatch was considered in [11]. The joint compensation of the transmitter and receiver I/Q mismatch in the presence of CFO was studied in [12]. The above methods usually need more than one OFDM training blocks for the joint compensation of the transmitter and receiver I/Q mismatch.

In this paper, we extend the time-domain method in [6] to the case where there are oscillator mismatches at both the transmitting and receiving ends. Unlike earlier reports on the joint estimation of transmitter and receiver I/Q imbalances, our method needs only one OFDM block for training. Exploiting the fact that the size of an OFDM block is usually larger than the channel order, we first show how to estimate the I/Q mismatches of the transmitter and receiver without the knowledge of the channel impulse response. The solution is given in a closed form. Then based on the estimated I/Q parameters, we obtain an estimate of the channel response. The proposed method can accurately estimate the transmitter and receiver I/Q mismatch and channel response using only one arbitrary OFDM block for training. Simulation results show that the BER performance of the proposed method is very close to the ideal case where the transmitter and receiver I/Q mismatch and channel response are perfectly known at the receiver.

Notation: Boldfaced lower and upper case letters represent vectors and matrices respectively. The notation $\mathbf{A}^\dagger$ denotes transpose-conjugate of $\mathbf{A}$, $\mathbf{A}^T$ denotes the transpose

This work was supported by the National Science Council, Taiwan, R.O.C., under Grant NSC 95-2752-E-002-006-PAE and NSC 95-2213-E-002-075.

of $\mathbf{A}$, and $\mathbf{A}^*$ denotes the complex conjugate of $\mathbf{A}$. $\mathbf{W}$ is the $M\times M$ normalized DFT matrix with entries $[\mathbf{W}]_{k,l} = \frac{1}{\sqrt{M}}e^{-j\frac{2\pi}{M}kl}$.

2. SYSTEM MODEL

Fig. 1 shows the block diagram of an OFDM system. First suppose that there is no I/Q imbalance, that is $\widetilde{\mathbf{x}} = \mathbf{x}$ and $\widetilde{\mathbf{r}} = \mathbf{r}$. At the transmitter, the input is an $M\times 1$ vector $\mathbf{s}$ consisting of modulation symbols. We first take an M-point IDFT to obtain an $M\times 1$ vector $\mathbf{x}$ and then a cyclic prefix (CP) of length L is appended before transmitting it through the channel. In this paper, we assume that the channel does not change during the transmission of one OFDM block and the channel order is $\leq L$ so that $h(l)$ is zero outside the interval $[0,L]$. At the receiver, after removing the CP, we apply an M-point DFT to the received vector $\mathbf{r}$. Because the CP length L is $\geq$ channel order, it is well-known that there is no interblock interference after the CP removal. Simple one-tap frequency domain equalizer can be used at the receiver for symbol recovery.

Suppose now there is I/Q imbalance. In this case, the vectors $\widetilde{\mathbf{x}}$ and $\widetilde{\mathbf{r}}$ are related to $\mathbf{x}$ and $\mathbf{r}$ respectively as [12]

$$\widetilde{\mathbf{x}} = \mu_t\mathbf{x} + \nu_t\mathbf{x}^* \tag{1}$$

$$\widetilde{\mathbf{r}} = \mu_r\mathbf{r} + \nu_r\mathbf{r}^* \tag{2}$$

where μ_t, ν_t, μ_r and ν_r are the parameters describing the I/Q imbalance. They are given by

$$\mu_i = \frac{1+\varepsilon_i e^{-j\phi_i}}{2}, \nu_i = \frac{1-\varepsilon_i e^{j\phi_i}}{2}, \text{ for } i\in\{t,r\}, \tag{3}$$

where ε_i and ϕ_i are respectively the amplitude and the phase mismatch of the oscillators. It can be shown [12] that due to the I/Q mismatch, the orthogonality of subcarriers in the OFDM system is destroyed. Therefore it is important to compensate the I/Q mismatch at the receiver. To do this, let us define the transmitter and receiver I/Q parameters respectively as

$$\alpha_t \triangleq \frac{\nu_t}{\mu_t}, \text{ and } \alpha_r \triangleq \frac{\nu_r}{\mu_r^*}. \tag{4}$$

For notational simplicity, we define a scaled version of the channel impulse response as

$$h_0(n) \triangleq \mu_t\mu_r h(n). \tag{5}$$

Below we will show how to recover the symbol $\mathbf{s}$ from the I/Q corrupted vector $\widetilde{\mathbf{r}}$ when α_t, α_r and $h_0(n)$ are known. Notice that from (2), we can write [2]

$$\mathbf{r}_0 \triangleq \mu_r\mathbf{r} = \frac{\widetilde{\mathbf{r}} - \alpha_r\widetilde{\mathbf{r}}^*}{1-|\alpha_r|^2}. \tag{6}$$

Due to the CP insertion and CP removal, the vectors $\mathbf{r}$ and $\widetilde{\mathbf{x}}$ in Fig. 1 are related by

$$\mathbf{r} = \mathbf{H}_{cir}\widetilde{\mathbf{x}} + \mathbf{q}, \tag{7}$$

where $\mathbf{q}$ is the noise vector and $\mathbf{H}_{cir}$ is an $M\times M$ circulant matrix with the first column

$$[\ h(0)\ \ \cdots\ \ h(L)\ \ 0\ \ \cdots\ \ 0\ ]^T. \tag{8}$$

Substituting (7) into (6) and using (1), we get

$$\mathbf{r}_0 = \mu_r\mu_t\mathbf{H}_{cir}(\mathbf{x} + \alpha_t\mathbf{x}^*) + \mathbf{q}. \tag{9}$$

Because $\mathbf{x} = \mathbf{W}^\dagger\mathbf{s}$, it can be shown [3] that

$$\mathbf{x}^* = \mathbf{W}^\dagger\mathbf{s}^\#, \tag{10}$$

where

$$\mathbf{s}^\# = [\ s_0^*\ \ s_{M-1}^*\ \ s_{M-2}^*\ \ \cdots\ \ s_1^*\ ]^T. \tag{11}$$

Taking the M-point DFT of $\mathbf{r}_0$, we have

$$\mathbf{y}_0 \triangleq \mathbf{W}\mathbf{r}_0 = \mathbf{W}(\mu_r\mu_t\mathbf{H})\mathbf{W}^\dagger(\mathbf{s} + \alpha_t\mathbf{s}^\#) + \mathbf{W}\mathbf{q}. \tag{12}$$

It is well known that circulant matrices are diagonalized by DFT matrices. Thus the matrix $\mathbf{D}_0$ defined below is a diagonal matrix:

$$\mathbf{D}_0 = \mathbf{W}(\mu_r\mu_t\mathbf{H})\mathbf{W}^\dagger. \tag{13}$$

The diagonal entries of $\mathbf{D}_0$ are related to $h_0(n)$ as

$$[\mathbf{D}_0]_{k,k} = \sum_{l=0}^{L} h_0(l)e^{-j\frac{2\pi}{M}kl}. \tag{14}$$

Substituting (13) into (12) and solving the equation for $\mathbf{s}$, we get

$$\mathbf{s} = \frac{1}{1-|\alpha_t|^2}(\mathbf{D}_0^{-1}\mathbf{y}_0 - \alpha_t(\mathbf{D}_0^{-1}\mathbf{y}_0)^\#). \tag{15}$$

From the above discussions, we see that the receiver only needs to know α_t, α_r and $h_0(n)$ to recover the transmitted vector $\mathbf{s}$ from the I/Q corrupted received vector $\widetilde{\mathbf{r}}$. We shall show how to estimate these parameters in the next section.

3. PROPOSED JOINT ESTIMATION METHOD

In this section, we consider the problem of channel estimation in the presence of transmitter and receiver I/Q imbalance. Suppose that we send one arbitrary vector $\mathbf{s}$ (known to receiver) for channel estimation. In other words, the vector $\mathbf{x}$, which is the IDFT of $\mathbf{s}$, is known. For channel estimation, we can rewrite (9) as

$$\mathbf{r}_0 = (\mathbf{X} + \alpha_t\mathbf{X}^*)\mathbf{h}_0 + \mathbf{q}, \tag{16}$$

where $\mathbf{X}$ is an $M\times M$ circulant matrix with the first column $\mathbf{x}$ and the $M\times 1$ vector

$$\mathbf{h}_0 = [\ h_0(0)\ \ \cdots\ \ h_0(L)\ \ 0\ \ \cdots\ \ 0\ ]^T. \tag{17}$$

Suppose that $\mathbf{X}$ is invertible (this condition holds if and only if the entries of the training vector $\mathbf{s}$ are nonzero, which is true for most practical applications). Then we can write

$$\mathbf{r}_0 = \mathbf{X}(\mathbf{I} + \alpha_t\mathbf{X}^{-1}\mathbf{X}^*)\mathbf{h}_0 + \mathbf{q}. \tag{18}$$

From the above equation, we get

$$\widehat{\mathbf{h}}_0 = (\mathbf{I} + \alpha_t\mathbf{X}^{-1}\mathbf{X}^*)^{-1}\mathbf{X}^{-1}\mathbf{r}_0. \tag{19}$$

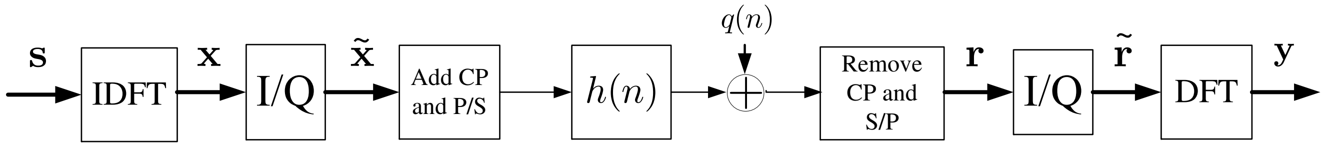


Figure 1: An OFDM system

From previous discussions, we know that $\mathbf{r}_0$ is related to the received vector $\widetilde{\mathbf{r}}$ as in (6). So we have

$$\widehat{\mathbf{h}}_0 = (\mathbf{I} + \alpha_t \mathbf{X}^{-1}\mathbf{X}^*)^{-1}\mathbf{X}^{-1}\frac{\widetilde{\mathbf{r}} - \alpha_r \widetilde{\mathbf{r}}^*}{1-|\alpha_r|^2}. \quad (20)$$

If α_t and α_r are known, one can obtain an estimate of $h_0(n)$ from the received vector $\widetilde{\mathbf{r}}$ using the above expression. Below we will show how to estimate α_t and α_r without the knowledge of $h_0(n)$. To do this, notice that the last $(M-L-1)$ entries of $\widehat{\mathbf{h}}_0$ in (20) will be zero when the channel noise is zero. In other words, for a moderate SNR, if α_t and α_r are accurately estimated, these entries will be small. Therefore we can get an accurate estimate of α_t and α_r by minimizing the last $(M-L-1)$ entries of $\widehat{\mathbf{h}}_0$ in (20). In practice, the block size M is usually larger than the CP length L so that the spectral efficiency is high. It is reasonable to assume that $M-L-1 \geq 2$. Below we will formulate the optimization problem.

Define the $(M-L-1)\times M$ matrix

$$\mathbf{E} = [\ \mathbf{0} \quad \mathbf{I}_{M-L-1}\ ]. \quad (21)$$

By premultiplying $\widehat{\mathbf{h}}_0$ by $\mathbf{E}$, we retain the last $(M-L-1)$ entries of the vector $\widehat{\mathbf{h}}_0$ and obtain the vector

$$\mathbf{E}\widehat{\mathbf{h}}_0 = \mathbf{E}(\mathbf{I} + \alpha_t \mathbf{X}^{-1}\mathbf{X}^*)^{-1}\mathbf{X}^{-1}\frac{\widetilde{\mathbf{r}} - \alpha_r \widetilde{\mathbf{r}}^*}{1-|\alpha_r|^2}. \quad (22)$$

To simplify the problem, we make three approximations:
1. $1-|\alpha_r|^2 \approx 1$.
2. $(\mathbf{I} + \alpha_t \mathbf{X}^{-1}\mathbf{X}^*)^{-1} \approx \mathbf{I} - \alpha_t \mathbf{X}^{-1}\mathbf{X}^*$.
3. $\alpha_t \alpha_r \approx 0$.

In practice, α_t and α_r are often small numbers. So these approximations are quite accurate. Using these approximations, we can write (22) as

$$\mathbf{E}\widehat{\mathbf{h}}_0 = \mathbf{A}\widetilde{\mathbf{r}} - \alpha_r \mathbf{A}\widetilde{\mathbf{r}}^* - \alpha_t \mathbf{B}\widetilde{\mathbf{r}}, \quad (23)$$

where the matrices

$$\mathbf{A} = \mathbf{E}\mathbf{X}^{-1}, \quad \mathbf{B} = \mathbf{E}\mathbf{X}^{-1}\mathbf{X}^*\mathbf{X}^{-1}. \quad (24)$$

As $\mathbf{X}$ is known, the matrices $\mathbf{A}$ and $\mathbf{B}$ can be precomputed at the receiver. We would like to find α_t and α_r so that $||\mathbf{E}\widehat{\mathbf{h}}_0||^2$ is minimized. Define the $(M-L-1)\times 2$ matrix

$$\mathbf{\Phi} = [\ \mathbf{A}\widetilde{\mathbf{r}}^* \quad \mathbf{B}\widetilde{\mathbf{r}}\ ]. \quad (25)$$

Since $(M-L-1) \geq 2$, the above matrix has full column rank. From linear algebra, we know that α_t and α_r that minimize $||\mathbf{E}\widehat{\mathbf{h}}_0||^2$ are given by

$$\begin{bmatrix} \widehat{\alpha}_r \\ \widehat{\alpha}_t \end{bmatrix} = (\mathbf{\Phi}^\dagger \mathbf{\Phi})^{-1}\mathbf{\Phi}^\dagger \mathbf{A}\widetilde{\mathbf{r}}. \quad (26)$$

Substituting $\widehat{\alpha}_t$ and $\widehat{\alpha}_r$ into (20), we can get an estimate of the channel response. Once $\widehat{\alpha}_t$, $\widehat{\alpha}_r$ and $\widehat{h}_0(n) = \mu_r \mu_t \widehat{h}(n)$ are known, one can use the algorithm described in Sec. 2 to recover the input vector $\mathbf{s}$. From (26), we see that the estimates of I/Q mismatches are given in a closed form. No optimization is needed. Also notice that in the above derivation, we need only one OFDM training block. Moreover, except the mild assumption that its entries are nonzero, we do not make any assumption on the OFDM training block $\mathbf{s}$.

Complexity: Note that as $\mathbf{X}$ is circulant, the matrices $\mathbf{X}^{-1}$, $\mathbf{X}^{-1}\mathbf{X}^*\mathbf{X}^{-1}$, and $(\mathbf{I} + \alpha_t \mathbf{X}^{-1}\mathbf{X}^*)^{-1}$ are circulant. Thus the computation of $\mathbf{A}\widetilde{\mathbf{r}}^*$, $\mathbf{B}\widetilde{\mathbf{r}}$, and $\widehat{\mathbf{h}}_0$ can be implemented efficiently using circulant convolution. Moreover, $(\mathbf{\Phi}^\dagger\mathbf{\Phi})$ is 2×2 and thus the complexity of (26) is also low.

Improving the Accuracy Using a Two-Step Approach: As we will see in the simulation, when the I/Q mismatches are small, (26) yields a very accurate estimate of the I/Q parameters. When the mismatches are large, the estimation error floors at high SNR. This is because when α_t and α_r are large, ignorance of the second order term $\alpha_t\alpha_r$ affects the accuracy. To avoid this, we can use the two-step algorithm below.
1. Estimate the I/Q parameters using (26). Denote the estimates as α_t^0 and α_r^0.
2. Obtain the new estimates using the following expression.

$$\begin{bmatrix} \widehat{\alpha}_r \\ \widehat{\alpha}_t \end{bmatrix} = (\mathbf{\Phi}^\dagger \mathbf{\Phi})^{-1}\mathbf{\Phi}^\dagger (\mathbf{A}\widetilde{\mathbf{r}} + \alpha_t^0\alpha_r^0 \mathbf{B}\widetilde{\mathbf{r}}^*). \quad (27)$$

Note that the two-step approach also needs only one OFDM training block.

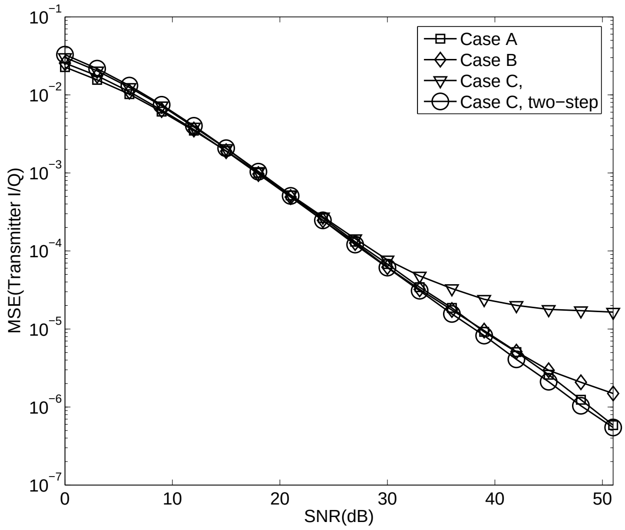
4. SIMULATION RESULTS

In this section, we carry out Monte-Carlo experiments to verify the performance of the proposed method. A total of 5000 random channels are generated in the experiments. The channel taps are i.i.d. complex Gaussian random variables. The variance of the channel taps are normalized by $\sum_{l=0}^{L} E\left\{|h(l)|^2\right\} = 1$. The training data are assumed to be BPSK symbols and they are randomly generated. The size of the DFT matrix is $M = 64$, and the CP length is L.

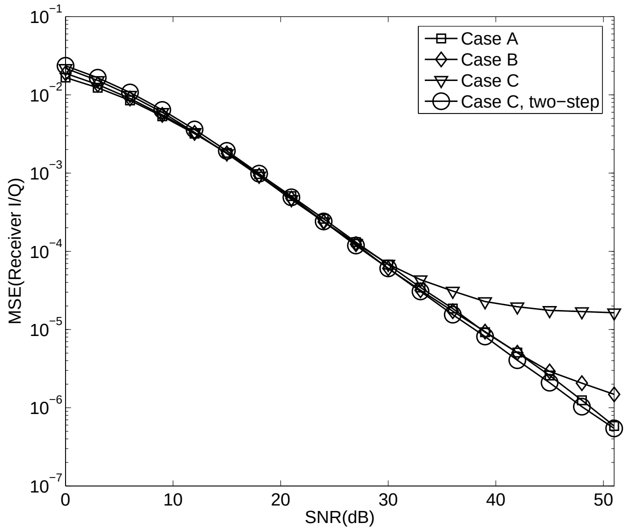
Fig. 2 shows mean-squared error (MSE). In Fig.2, we assume that $L = 3$. The MSE of the I/Q mismatch parameter α_i and channel response are respectively defined as

$$\begin{aligned} MSE(\alpha_i) &= E\left\{|\alpha_i - \widehat{\alpha}_i|^2\right\}, \\ MSE(\text{channel}) &= \frac{1}{L+1}\sum_{l=0}^{L} E\left\{|h(l) - \widehat{h}(l)|^2\right\}. \quad (28) \end{aligned}$$

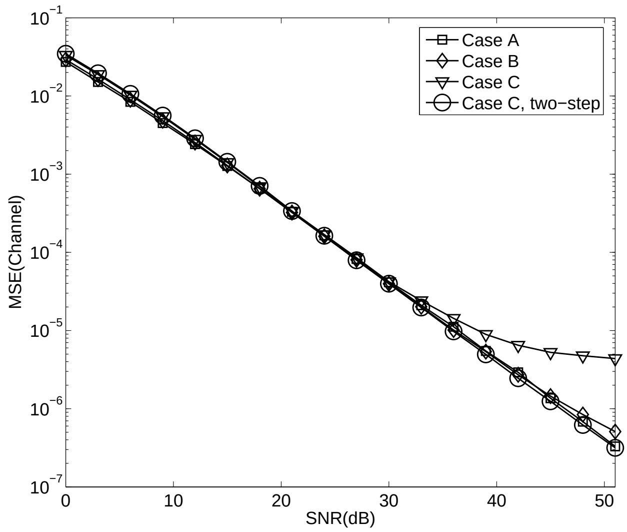
Three different pairs of transmitter and receiver I/Q mismatch are considered: (A) $\epsilon_t = \epsilon_r = 1$, $\phi_t = \phi_r = 0°$, (B) $\epsilon_t =$



(a)



(b)



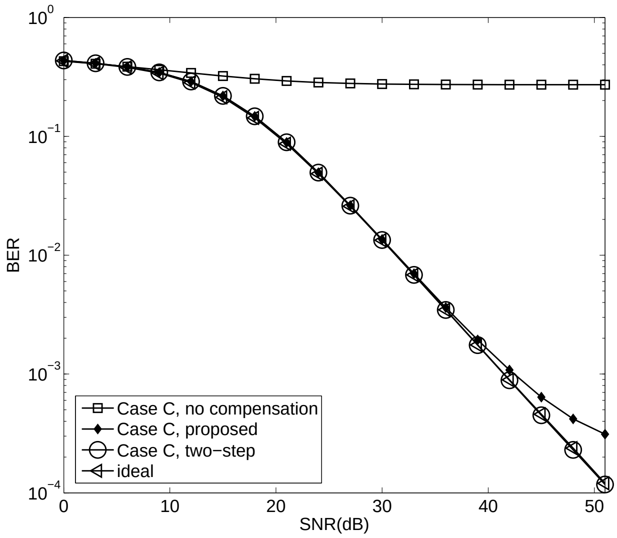
(c)

Figure 2: (a) MSE of the estimated transmitter I/Q factor α_t; (b) MSE of the estimated receiver I/Q factor α_r; (c) MSE of the estimated channel.

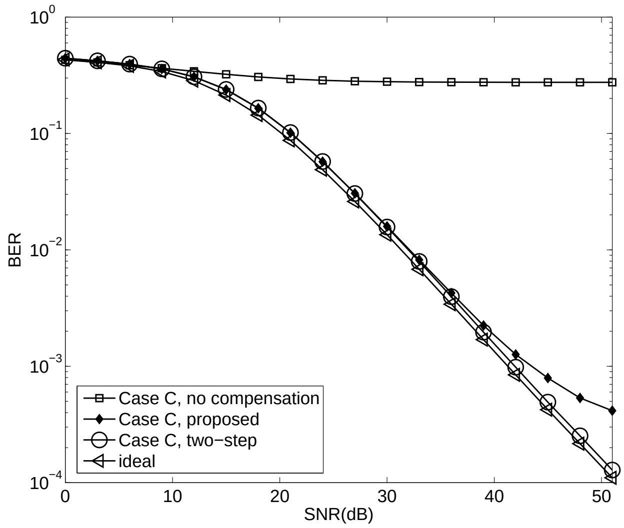
$\varepsilon_r = 1.1$, $\phi_t = \phi_r = 10^\circ$ and (C) $\varepsilon_t = \varepsilon_r = 1.2$, $\phi_t = \phi_r = 15^\circ$. The corresponding I/Q parameters are (A) $\alpha_t = \alpha_r = 0$, (B) $\alpha_t = -0.0313 - 0.0946j$, $\alpha_r = -0.048 - 0.0873j$, and (C) $\alpha_t = -0.0519 - 0.1513j$, $\alpha_r = -0.0925 - 0.1305j$ respectively. Notice that in Case A, there is no I/Q mismatch. From Fig. 2, it is found that when the I/Q mismatches are relatively small (Case A and B), the formula in (26) gives a very accurate estimate. The MSEs decrease monotonically with the increase of SNR in both Case A and B. When the SNR is $38dB$, the MSEs is about 10^{-5}. In Case C, it is found that the MSE performance suffers from an error-flooring due to the ignorance of the second order term $\alpha_t\alpha_r$. With the proposed two-step algorithm, we see that the performance can be significantly improved. The MSEs no longer floor at high SNR. Note that we did not compare our method with other existing methods on joint compensation of transmitter and receiver I/Q imbalances. This is because the existing methods need more than one OFDM blocks for training. For the purpose of comparison, we compare the BER performance of our system with the ideal case where the I/Q mismatch parameters and channel response are perfectly known at the receiver (which is the bound on BER).

Fig. 3 shows the BER performance of the proposed method for Case C. The channel order L is assumed to be 3 and 15 in Fig. 3(a) and (b) respectively. The transmitted data are 64 QAM. In Fig. 3, it is seen that when the I/Q mismatches are not compensated, the BER is very large because the I/Q mismatches in Case C are relatively large. When the proposed two-step approach is employed, the BER performance can be significantly improved. In Fig. 3(a), it is found that when $L = 3$, the BER performance is almost indistinguishable from the ideal case where the I/Q mismatch parameters and channel response are perfectly known at the receiver. Without the two-step algorithm, the BER of the proposed method deviates slightly from the ideal case at high SNR [1]. In Fig. 3(b), it is found that when $L = 15$, with the

[1]For Case A and B, experiment results show that the BER performance is very close to the ideal case even when the two-step algorithm is not employed.



(a)



(b)

Figure 3: BER performance of OFDM systems in the case of (a) $L = 3$; (b) $L = 15$.

two-step algorithm, the BER performance of the proposed method is very close to the ideal case. From Fig. 3, we can conclude that the BER performance of our method is very close to the bound of the ideal case.

5. CONCLUSION

In this paper, we consider joint estimation of the transmitter and receiver I/Q mismatch and channel response for OFDM systems. With only one OFDM block for training, we can accurately estimate the transmitter and receiver I/Q parameters and channel response. The BER performance of the proposed method is very close the ideal case where the I/Q mismatch parameters and channel response are perfectly known at the receiver.

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