I/Q Imbalance Mitigation for STBC MIMO-OFDM Communication Systems

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***Abstract*— In this work we study the performance degradation caused by the in-phase/quadrature (I/Q) imbalance in space-time block coded (STBC) multiple-input multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) communication systems. The 2-Tx Alamouti scheme, 4-Tx quasi-orthogonal STBC (QOSTBC) scheme, and 4-Tx rotated QOSTBC (RQOSTBC) scheme with I/Q imbalance are examined in details. Our study shows that I/Q imbalance causes severe distortion in STBC MIMO-OFDM systems. By exploiting the structure of the received signal, low-complexity solutions are developed to mitigate the resultant distortion successfully.**

***Index Terms*— I/Q imbalance, STBC, MIMO, OFDM**

I. Introduction

STBC-MIMO communication systems provide reliable data transmission by exploiting the spatial diversity in fading channels. Within various space-time block codes (STBCs), orthogonal STBCs (OSTBCs) are studied extensively since they can provide full spatial diversity and a linear maximum likelihood (ML) decoder. The drawback is that only Alamouti scheme can provide full diversity and rate-one transmission for complex constellation symbols with a linear ML receiver. To achieve rate-one transmission while sacrificing full spatial diversity, quasi-orthogonal STBCs (QOSTBCs) are proposed with low-complexity pair-wise ML decoding feasible. By choosing some symbols from a rotated constellation, the rotated QOSTBCs (RQOSTBCs) can provide full diversity, rate-one, and pair-wise ML decoding properties [1].

Since STBCs are developed for flat fading channels, to achieve spatial diversity in frequency selective fading channels, MIMO-OFDM can be utilized, where STBCs are used across different antennas in conjunction with OFDM [2].

In practical applications, the I/Q imbalance always exists due to the non-ideal matching between the relative amplitudes and phases of in-phase and quadrature branches of the transceivers, especially in direct conversion transceiver design [3]. The resultant distortion in STBC MIMO-OFDM systems increases symbol error rate (SER) dramatically, especially in high order modulations or with higher carrier frequencies, and should be efficiently compensated [4] [5]. The impacts and solutions of I/Q imbalance on single-input-single-output (SISO) OFDM and MIMO-OFDM receivers are studied in [4] and [6], respectively. Taking into consideration of the I/Q imbalance of transmitter, the effect of transceiver I/Q imbalance on STBC MIMO systems is studied without providing solutions in [7]. The performance analysis in terms of signal-to-interference ratio (SIR) of STBC MIMO-OFDM systems with I/Q imbalance is studied in [5]. To obtain the results in [7] and [5], the channels and I/Q imbalance parameters should be estimated separately. However, the I/Q imbalance parameters are absorbed in the channels, and it is better to estimate the effective channels only. In this work, taking into consideration of the noise corrupted by I/Q imbalance, we examine the effect of transceiver I/Q imbalance on the STBC MIMO-OFDM communication systems in a more reasonable way, and low-complexity solutions are developed to mitigate the distortion successfully.

II. Signal and System Models

A. Transceiver I/Q Imbalance Model

In a $N \times M$ MIMO communication systems, assume the baseband signal to be transmitted from the i-th antenna is

$$x_i(t) = x_{i,I}(t) + jx_{i,Q}(t), \quad i = 1, \cdots, N, \tag{1}$$

where $x_{i,I}(t)$ and $x_{i,Q}(t)$ denote the I and Q part of $x_i(t)$, respectively, and $j = \sqrt{-1}$. Let ω_c denote the carrier frequency, $\epsilon_{T,i}$ and $\varphi_{T,i}$ denote the amplitude and phase imbalances of the local oscillator (LO) of the i-th transmit antenna, respectively. Assuming a two-parameter frequency-independent I/Q imbalance model is used in this paper, then the output of LO of the i-th transmit antenna can be expressed as $f_{T,i}(t) = \cos(\omega_c t) + j(1+\epsilon_{T,i})\sin(\omega_c t + \varphi_{T,i}) = \alpha_{T,i}e^{j\omega_c t} + \beta_{T,i}e^{-j\omega_c t}$, where $\alpha_{T,i} = \frac{1}{2}[1 + (1+\epsilon_{T,i})e^{j\varphi_{T,i}}]$, and $\beta_{T,i} = \frac{1}{2}[1 - (1+\epsilon_{T,i})e^{-j\varphi_{T,i}}]$. Consequently, the up-converted band-pass signal can be expressed as $\Re\{\bar{x}_i(t)e^{j\omega_c t}\}$. In this paper, $\Re\{x\}$ and $\Im\{x\}$ stand for real and imaginary part of x, respectively. The equivalent baseband signal can be expressed as

$$\bar{x}_i(t) = \alpha_{T,i}x_i(t) + \beta^*_{T,i}x^*_i(t), \quad i = 1, \cdots, N. \tag{2}$$

Here $(\cdot)^*$ and $(\cdot)^H$ denote complex-conjugate and complex-conjugate transpose, respectively.

Similarly, let $\epsilon_{R,m}$ and $\varphi_{R,m}$ $(m = 1, \cdots, M)$ be the amplitude and phase imbalances of LO of the m-th receive antenna, respectively. Then the output of LO of the m-th receive antenna can be expressed as $f_{R,m}(t) = \cos(\omega_c t) - j(1+\epsilon_{R,m})\sin(\omega_c t + \varphi_{R,m}) = \alpha_{R,m}e^{-j\omega_c t} + \beta_{R,m}e^{j\omega_c t}$, where $\alpha_{R,m} = \frac{1}{2}[1+(1+\epsilon_{R,m})e^{-j\varphi_{R,m}}]$, and $\beta_{R,m} = \frac{1}{2}[1-(1+\epsilon_{R,m})e^{j\varphi_{R,m}}]$. Consequently the down-converted baseband noise-free signal from the i-th transmit antenna to the m-th receive antenna can be expressed as

$$\tilde{x}_{i,m}(t) = \alpha_{R,m}\bar{x}_i(t) + \beta_{R,m}\bar{x}_i^*(t). \tag{3}$$

As shown in (2) and (3), in time domain, the intended signals are interfered by their own complex-conjugates due to the I/Q imbalance of transmitter and/or receiver [7].

In OFDM systems, the equalization and detection are performed in frequency domain. In frequency domain, (2) and (3) can be written as

$$\bar{X}_i(k) = \alpha_{T,i}X_i(k) + \beta_{T,i}^*X_i^*(-k), \tag{4}$$

$$\tilde{X}_{i,m}(k) = \alpha_{R,m}\bar{X}_i(k) + \beta_{R,m}\bar{X}_i^*(-k), \tag{5}$$

where $k = -K/2, -K/2+1, \cdots, K/2-2, K/2-1$ denotes the subcarrier index (the mirror-subcarriers of $k = -K/2$ and $k = 0$ are themselves). Here K is the fast Fourier transform (FFT) size. It is seen that in frequency domain, the I/Q imbalance of transmitter or receiver results in the inter-carrier interference (ICI) between the mirror-subcarrier pairs [5].

B. OSTBC MIMO-OFDM System With I/Q Imbalance

To simplify the study, a 2×1 Alamouti OFDM scheme with I/Q imbalance is examined here to demonstrate the distortion.

Without loss of generality, assume the transmitter sends $S_1(k)$ from antenna-one and $S_2(k)$ from antenna-two at time 1 over subcarrier k, then at time 2, it sends $-S_2^*(k)$ and $S_1^*(k)$ from antenna-one and antenna-two over subcarrier k, respectively. We denote the corresponding transmitted codeword as

$$\mathbf{C} = \begin{pmatrix} S_1 & S_2 \\ -S_2^* & S_1^* \end{pmatrix}, \tag{6}$$

where the element of the l-th row and the i-th column of $\mathbf{C}$ denotes the symbol sent from the i-th transmit antenna at time l. Assume the cyclic prefix (CP) is longer than the channel delay spread, and the channel keeps steady within the coded block interval. Let $H_i(k)$ denote the frequency domain channel response between the i-th transmit antenna and the receive antenna over subcarrier k, take the noise corrupted by I/Q imbalance into consideration, the received signal at time 1 and 2 over subcarrier k (except for $k = -K/2$ and $k = 0$) after CP removal can be obtained from (4) and (5) as

$$\underbrace{\begin{bmatrix} Y_1(k) \\ Y_2^*(k) \end{bmatrix}}_{Y_a(k)} = \mathbf{W}_a(k)\underbrace{\begin{bmatrix} S_1(k) \\ S_2(k) \end{bmatrix}}_{S_a(k)} + \mathbf{U}_a(k)\underbrace{\begin{bmatrix} S_1^*(-k) \\ S_2^*(-k) \end{bmatrix}}_{S_a^*(-k)} + \underbrace{\begin{bmatrix} \bar{V}_1(k) \\ \bar{V}_2^*(k) \end{bmatrix}}_{V_a(k)}, \tag{7}$$

where

$$\mathbf{W}_a(k) = \begin{bmatrix} w_1(k) & w_2(k) \\ w_2^*(k) & -w_1^*(k) \end{bmatrix}, \ \mathbf{U}_a(k) = \begin{bmatrix} u_1(k) & u_2(k) \\ u_2^*(k) & -u_1^*(k) \end{bmatrix},$$

and $\bar{V}(k) = \alpha_R V(k) + \beta_R V^*(-k)$. Note $S_a(k)$ and $S_a(-k)$ are two different data vectors. Since there is only one receive antenna, here the receive antenna indexes in the subscripts of $\alpha_{R,1}$ and $\beta_{R,1}$ are dropped for simplicity, and $V(k) \sim \mathcal{CN}(0, \sigma_v^2)$ denotes the proper AWGN in frequency domain at subcarrier k. The elements of $\mathbf{W}_a(k)$ and $\mathbf{U}_a(k)$ can be obtained by

$$\begin{matrix} w_i(k) = \alpha_{T,i}\alpha_R H_i(k) + \beta_{T,i}\beta_R H_i^*(-k) \\ u_i(k) = \beta_{T,i}^*\alpha_R H_i(k) + \alpha_{T,i}^*\beta_R H_i^*(-k) \end{matrix}, \quad i = 1, 2, \tag{8}$$

It shows that without the transmitter I/Q imbalance, i.e., $\{\beta_{T,i}\}_{i=1}^2 = 0$, then $H_i^*(-k)$ and $H_i(k)$ will not be involved in $w_i(k)$ and $u_i(k)$, respectively, the model reduces to the model used in [6]. Since $Y(k)$ is a linear model of the effective channel parameters $\{w_i(k)\}_{i=1}^2$ and $\{u_i(k)\}_{i=1}^2$, their corresponding estimates, $\{\hat{w}_i(k)\}_{i=1}^2$ and $\{\hat{u}_i(k)\}_{i=1}^2$, can be obtained by sending training sequences based on the assumption that $\{w_i(k)\}_{i=1}^2$ and $\{u_i(k)\}_{i=1}^2$ are constant within the period [8]. If there is no I/Q imbalance, $\{\alpha_{T,i}\}_{i=1}^2 = \alpha_R = 1$, and $\{\beta_{T,i}\}_{i=1}^2 = \beta_R = 0$, then $w_i(k) = H_i(k)$ $(i = 1, 2)$, and $\{u_i(k)\}_{i=1}^2 = 0$, respectively.

To achieve the diversity gain without considering the I/Q imbalance, the transmitted signal can be recovered by [1]

$$\hat{S}_a(k) = \mathbf{W}_a^H(k)Y_a(k). \tag{9}$$

Here $\mathbf{W}_a(k)$ is assumed known since it can be estimated. Then it can be obtained that

$$\hat{S}_1(k) = (\|w_1\|^2 + \|w_2\|^2)S_1(k) + (w_1^*u_1 + w_2u_2^*)S_1^*(-k) + (w_1^*u_2 - w_2u_1^*)S_2^*(-k) + w_1^*V_1(k) + w_2V_2^*(k), \tag{10}$$

$$\hat{S}_2(k) = (\|w_1\|^2 + \|w_2\|^2)S_2(k) + (w_2^*u_2 + w_1u_1^*)S_2^*(-k) + (u_1w_2^* - w_1u_2^*)S_1^*(-k) + w_2^*V_1(k) - w_1V_2^*(k). \tag{11}$$

Here the subcarrier index k of $\{w_i(k)\}_{i=1}^2$ and $\{u_i(k)\}_{i=1}^2$ is dropped for simplicity. As shown in (10) and (11), without I/Q imbalance, i.e., $\{u_i\}_{i=1}^2 = 0$, the transmitted signal can be detected separately by linear ML estimation. However, due to the I/Q imbalance, the detected signal at subcarrier k is corrupted by not only the conjugate of the signal at subcarrier $-k$ within the same signal vector performing inverse FFT (IFFT), but also the conjugate of the signal within the coding block at subcarrier $-k$. This distortion degrades SER performance and should be compensated to improve the detection reliability. It should be noted that the result obtained here is different to the result in [5], where there are three independent signals involved in the interference, but here only two independent signals are involved in the interference.

C. Extension to QOSTBC & RQOSTBC MIMO-OFDM System

Without loss of generality, a 4×1 QOSTBC scheme with codeword

$$\mathbf{C} = \begin{pmatrix} S_1 & S_2 & S_3 & S_4 \\ -S_2^* & S_1^* & -S_4^* & S_3^* \\ -S_3^* & -S_4^* & S_1^* & S_2^* \\ S_4 & -S_3 & -S_2 & S_1 \end{pmatrix} \tag{12}$$

is examined, then the received signal corrupted by I/Q imbalance can be expressed as

$$\underbrace{\begin{bmatrix} Y_1(k) \\ Y_2^*(k) \\ Y_3^*(k) \\ Y_4(k) \end{bmatrix}}_{Y_b(k)} = \mathbf{W}_b(k) \underbrace{\begin{bmatrix} S_1(k) \\ S_2(k) \\ S_3(k) \\ S_4(k) \end{bmatrix}}_{S_b(k)} + \mathbf{U}_b(k) \underbrace{\begin{bmatrix} S_1^*(-k) \\ S_2^*(-k) \\ S_3^*(-k) \\ S_4^*(-k) \end{bmatrix}}_{S_b^*(-k)} + \underbrace{\begin{bmatrix} \bar{V}_1(k) \\ \bar{V}_2^*(k) \\ \bar{V}_3^*(k) \\ \bar{V}_4(k) \end{bmatrix}}_{V_b(k)}, \tag{13}$$

where

$$\mathbf{W}_b(k) = \begin{bmatrix} w_1 & w_2 & w_3 & w_4 \\ w_2^* & -w_1^* & w_4^* & -w_3^* \\ w_3^* & w_4^* & -w_1^* & -w_2^* \\ w_4 & -w_3 & -w_2 & w_1 \end{bmatrix}, \mathbf{U}_b(k) = \begin{bmatrix} u_1 & u_2 & u_3 & u_4 \\ u_2^* & -u_1^* & u_4^* & -u_3^* \\ u_3^* & u_4^* & -u_1^* & -u_2^* \\ u_4 & -u_3 & -u_2 & u_1 \end{bmatrix}.$$

Here the subcarrier index k of $\{w_i(k)\}_{i=1}^4$ and $\{u_i(k)\}_{i=1}^4$ are dropped for simplicity, and $\{w_i(k)\}_{i=1}^4$ and $\{u_i(k)\}_{i=1}^4$ can be obtained from (8) by letting $i = 1, \cdots, 4$.

To achieve full diversity, assume the same codeword in (12) is used, $s_1, s_2 \in \mathcal{C}$ and $s_3, s_4 \in e^{j\theta}\mathcal{C}$, where $\mathcal{C}$ denotes the constellation set. Then the corresponding received signal becomes

$$Y_b(k) = \tilde{\mathbf{W}}_b(k) S_b(k) + \tilde{\mathbf{U}}_b(k) S_b^*(-k) + V_b(k), \tag{14}$$

where

$$\begin{aligned} \tilde{\mathbf{W}}_b(k) &= \left[\begin{array}{cccc} \mathbf{w}_{b,1}(k) & \mathbf{w}_{b,2}(k) & \mathbf{w}_{b,3}(k)e^{j\theta} & \mathbf{w}_{b,4}(k)e^{j\theta} \end{array}\right] \\ \tilde{\mathbf{U}}_b(k) &= \left[\begin{array}{cccc} \mathbf{u}_{b,1}(k) & \mathbf{u}_{b,2}(k) & \mathbf{u}_{b,3}(k)e^{-j\theta} & \mathbf{u}_{b,4}(k)e^{-j\theta} \end{array}\right] \end{aligned}.$$

Here $\mathbf{w}_{b,i}(k)$ and $\mathbf{u}_{b,i}(k)$ denote the i-th column of $\mathbf{W}_b(k)$ and $\mathbf{U}_b(k)$, respectively.

It is easy to show that the pair-wise ML decoding is not applicable to QOSTBC or RQOSTBC MIMO-OFDM scheme due to the I/Q imbalance.

III. Low-complexity Solutions

It can be obtained from (7) that

$$\underbrace{\begin{bmatrix} Y_a(k) \\ Y_a^*(-k) \end{bmatrix}}_{\vec{Y}_a(k)} = \underbrace{\begin{bmatrix} \mathbf{W}_a(k) & \mathbf{U}_a(k) \\ \mathbf{U}_a^*(-k) & \mathbf{W}_a^*(-k) \end{bmatrix}}_{\mathbf{H}_a(k)} \underbrace{\begin{bmatrix} S_a(k) \\ S_a^*(-k) \end{bmatrix}}_{\vec{S}_a(k)} + \begin{bmatrix} V_a(k) \\ V_a^*(-k) \end{bmatrix}. \tag{15}$$

The least squares (LS) estimate of $\vec{S}_a(k)$ can be given as

$$\hat{\vec{S}}_a(k) = (\mathbf{H}_a^H(k)\mathbf{H}_a(k))^{-1}\mathbf{H}_a^H(k)\vec{Y}_a(k). \tag{16}$$

Since the 2×2 sub-matrices to be inverted to obtain $(\mathbf{H}_a^H(k)\mathbf{H}_a(k))^{-1}$ are diagonal due to the special structure induced by the Alamouti coding scheme, a low-complexity LS estimator is available, and it is applicable to other OSTBC systems with multiple receive antennas [6].

Similarly, it can be obtained from (13) that

$$\underbrace{\begin{bmatrix} Y_b(k) \\ Y_b^*(-k) \end{bmatrix}}_{\vec{Y}_b(k)} = \underbrace{\begin{bmatrix} \mathbf{W}_b(k) & \mathbf{U}_b(k) \\ \mathbf{U}_b^*(-k) & \mathbf{W}_b^*(-k) \end{bmatrix}}_{\mathbf{H}_b(k)} \underbrace{\begin{bmatrix} S_b(k) \\ S_b^*(-k) \end{bmatrix}}_{\vec{S}_b(k)} + \underbrace{\begin{bmatrix} V_b(k) \\ V_b^*(-k) \end{bmatrix}}_{\vec{V}_b(k)}. \tag{17}$$

The 8×8 matrix $\mathbf{H}_b^H(k)\mathbf{H}_b(k)$ can be expressed by four 4×4 sub-matrices as

$$\mathbf{H}_b^H(k)\mathbf{H}_b(k) = \begin{bmatrix} \mathbf{A} & \mathbf{B} \\ \mathbf{B}^H & \mathbf{D} \end{bmatrix}. \tag{18}$$

Let $\mathbf{E} = \mathbf{A} - \mathbf{B}\mathbf{D}^{-1}\mathbf{B}^H$, according to the matrix inversion formula, $\mathbf{D}^{-1}$ and $\mathbf{E}^{-1}$ are involved in the inversion of (18). It is easy to show that $\mathbf{D}$ and $\mathbf{E}$ have the similar property of the matrix $\mathbf{H}_a^H(k)\mathbf{H}_a(k)$ in (16). Consequently, the 2×2 sub-matrices to be inverted to obtain $\mathbf{D}^{-1}$ and $\mathbf{E}^{-1}$ are diagonal, resulting in a low computational matrix inversion.

However, for RQOSTBC MIMO-OFDM system, the distortion can not be mitigated satisfactorily only by LS method. To further improve the performance, based on the LS estimate, the polygon local searching ML estimator proposed in [9] can be applied.

The received signal of a 4×1 RQOSTBC scheme can be obtained from (14) and (17) as

$$\vec{Y}_b(k) = \tilde{\mathbf{H}}_b(k)\vec{S}_b(k) + \vec{V}_b(k), \tag{19}$$

where

$$\tilde{\mathbf{H}}_b(k) = \begin{bmatrix} \tilde{\mathbf{W}}_b(k) & \tilde{\mathbf{U}}_b(k) \\ \tilde{\mathbf{U}}_b^*(-k) & \tilde{\mathbf{W}}_b^*(-k) \end{bmatrix}.$$

Let $p(\vec{Y}_b(k); \vec{S}_b(k))$ denotes the probability mass function (PMF) of $\vec{Y}_b(k)$ parameterized by the unknown parameter $\vec{S}_b(k)$, according to (19), the ML estimate of $\vec{S}_b(k)$ from $\vec{Y}_b(k)$ is

$$\hat{\vec{S}}_{b,\mathrm{M}}(k) = \arg \max_{\vec{S}_b(k) \in \mathcal{C}} p(\vec{Y}_b(k); \vec{S}_b(k)). \tag{20}$$

Since $\bar{V}(k)$ is a linear combination of two independent proper Gaussian, $\bar{V}(k)$ is still a proper Gaussian. However, $\bar{V}(k)$ and $\bar{V}^*(-k)$ are correlated. It can be obtained that $E[\bar{V}_I(k)\bar{V}_I(-k)] = E[\bar{V}_Q(k)\bar{V}_Q(-k)] = -\frac{\sigma_v^2}{4}(2\epsilon_R + \epsilon_R^2)$, and $E[\bar{V}_I(k)\bar{V}_Q(-k)] = -E[\bar{V}_Q(k)\bar{V}_I(-k)] = \frac{\sigma_v^2}{2}(1+\epsilon_R)\sin\phi_R$. Consequently, the unknown I/Q imbalance parameters ϵ_R and ϕ_R are involved in the expression of the ML objective function. Considering the practical values of the amplitude and phase imbalances are around $1\% \sim 5\%$ and $1^\circ \sim 5^\circ$, respectively, it is reasonable to treat $\bar{V}(k)$ and $\bar{V}^*(-k)$ as independent, especially in high SNR scenario, then (20) boils down to

$$\hat{\vec{S}}_{b,\mathrm{M}}(k) = \arg \min_{\vec{S}_b(k) \in \mathcal{C}} \|\vec{Y}_b(k) - \tilde{\mathbf{H}}_b(k)\vec{S}_b(k)\|^2. \tag{21}$$

Let $\hat{\vec{S}}_b(k) = (\tilde{\mathbf{H}}_b^H(k)\tilde{\mathbf{H}}_b(k))^{-1}\tilde{\mathbf{H}}_b^H(k)\vec{Y}_b(k) = \mathbf{G}^H(k)\vec{Y}_b(k)$, to avoid the global searching within $\mathcal{C}$, a polygon local searching detector based on $\hat{\vec{S}}_b(k)$ can be applied. It holds that

$$\|\vec{Y}_b(k) - \tilde{\mathbf{H}}_b(k)\hat{\vec{S}}_{b,\mathrm{M}}(k)\|^2 \leq \|\vec{Y}_b(k) - \tilde{\mathbf{H}}_b(k)\hat{\vec{S}}_b(k)\|^2 = d^2. \tag{22}$$

Let $\{\mathbf{g}_i(k)\}_{i=1}^8$ denotes the i-th column of $\mathbf{G}(k)$, according to [9], the boundary of the shrunk area can be given by

$$-d \leq \Re\left\{\frac{\vec{Y}_{b,i}(k) - \hat{\vec{Y}}_{b,i}(k)}{\|\mathbf{g}_i(k)\|}\right\} \leq d, \quad -d \leq \Im\left\{\frac{\vec{Y}_{b,i}(k) - \hat{\vec{Y}}_{b,i}(k)}{\|\mathbf{g}_i(k)\|}\right\} \leq d \ , \tag{23}$$

where $\vec{Y}_{b,i}(k)$ and $\hat{\vec{Y}}_{b,i}(k)$ denote the i-th elements of $\vec{Y}_b(k)$ and $\hat{\vec{Y}}_b(k)$, respectively.

IV. Simulation Results

In the simulations, the worst case, amplitude imbalance $\epsilon = 5\%$ and phase imbalance $\varphi = 5^\circ$ at both of the transmitter and receiver sides are examined. In addition, FFT size $K = 64$ and 4-tap frequency-selective channel (each tap is Rayleigh fading) are used. The channels remain constant during the encoded symbol block intervals, and the effective channels are assumed known. Moreover, $\theta = \pi/4$, the optimal rotation for QAM constellation is chosen for RQOSTBC scheme [10].

Fig.1 shows that the I/Q imbalance of receiver causes more distortion than that of transmitter. Fig.2∼Fig.4 show that by using the linear ML estimator for Alamouti scheme or the pair-wise ML estimator for QOSTBC and RQOSTBC schemes with I/Q imbalance without compensation, respectively, the SER of the estimated signal increases dramatically compared to that of the estimated signal without I/Q imbalance, resulting in an error floor. However, the resultant distortion can be successfully mitigated by the proposed solutions.

V. Conclusions

In this work we study the performance degradation and the solution for STBC MIMO-OFDM communication systems with I/Q imbalance. Our study shows that I/Q imbalance causes severe distortion in STBC MIMO-OFDM communication systems, and the proposed low-complexity LS method and the polygon local search ML method can mitigate the resultant distortion successfully.

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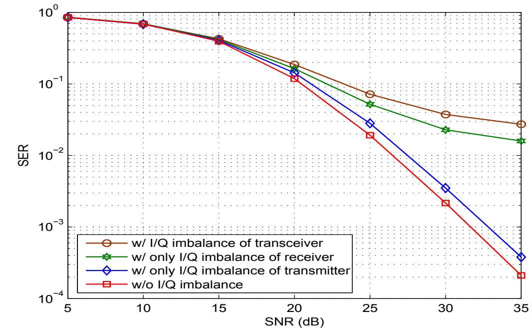

Fig. 1. SER versus SNR with 2×1 Alamouti scheme. $\{\epsilon_{T,i}\}_{i=1}^2 = 5\%$, $\{\varphi_{T,i}\}_{i=1}^2 = 5^\circ$, $\epsilon_R = 5\%$, and $\varphi_R = 5^\circ$. 64-QAM constellation is used. Number of trial is 10^6.

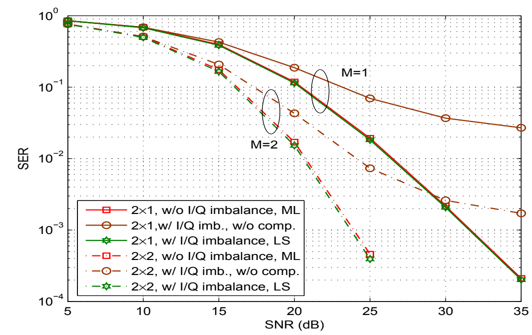

Fig. 2. SER versus SNR with Alamouti scheme. $\{\epsilon_{T,i}\}_{i=1}^2 = 5\%$, $\{\varphi_{T,i}\}_{i=1}^2 = 5^\circ$, $\{\epsilon_{R,i}\}_{i=1}^2 = 5\%$, and $\{\varphi_{R,i}\}_{i=1}^2 = 5^\circ$. 64-QAM constellation is used. Number of trial is 10^6.

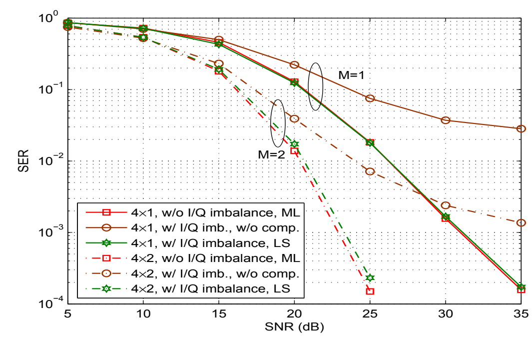

Fig. 3. SER versus SNR with 4-Tx QOSTBC scheme. $\{\epsilon_{T,i}\}_{i=1}^4 = 5\%$, $\{\varphi_{T,i}\}_{i=1}^4 = 5^\circ$, $\{\epsilon_{R,i}\}_{i=1}^2 = 5\%$, and $\{\varphi_{R,i}\}_{i=1}^2 = 5^\circ$. 64-QAM constellation is used. Number of trial is 10^6.

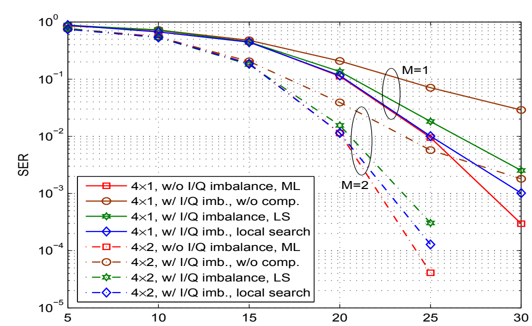

Fig. 4. SER versus SNR with 4-Tx RQOSTBC scheme. $\{\epsilon_{T,i}\}_{i=1}^4 = 5\%$, $\{\varphi_{T,i}\}_{i=1}^4 = 5^\circ$, $\{\epsilon_{R,i}\}_{i=1}^2 = 5\%$, $\{\varphi_{R,i}\}_{i=1}^2 = 5^\circ$, and $\theta = \frac{\pi}{4}$. 64-QAM constellation is used. Number of trial is 5×10^6.