JOINT ESTIMATION AND COMPENSATION OF FREQUENCY, DC-OFFSET, I-Q IMBALANCE AND CHANNEL IN MIMO RECEIVERS

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ABSTRACT

Direct-conversion is a low-cost RF architecture that requires fewer external components in chip implementation. It introduces, however, extra RF impairments such as I-Q imbalance and DC-offset that degrade severely the receiver performance if left uncompensated. In this paper, a new method of joint estimation and compensation of frequency, DC-offset, I-Q imbalance and channel is developed for MIMO receivers. Numerical results show that the proposed method has a very good performance; less than 2 dB loss is observed at bit-error rate 10^{-4}, as compared to the ideal receiver. In addition, a new training sequence along with a low-complexity architecture is proposed to reduce largely the receiver complexity with negligible performance loss.

Index Terms— MIMO Receiver, Direct-Conversion RF Architecture, Joint Least-Squares Estimation.

1. INTRODUCTION

Using multiple transmit and receive antennas in a wireless communication system, a.k.a. the multiple-input multiple-output (MIMO) system, is capable of providing diversity gain, array gain (power gain), and/or degree-of-freedom gain over the single-input single-output (SISO) systems [1]. Space-time coding, beam-forming and spatial multiplexing are modes of operations to exploit the diversity gain, array gain, and degree-of-freedom gain, respectively. MIMO has been one of the key technologies to enable high data-rate, high spectral-efficiency transmission in wireless communications.

Using direct-conversion RF architecture, on the other hand, is a low-cost design that requires fewer external components in chip implementation [2]. Direct-conversion RF architecture, nevertheless, introduces extra RF impairments such as I-Q imbalance and DC-offset that if left uncompensated will degrade severely the receiver performance. Receiver design with direct-conversion RF has been an important research topic [3]-[7]: first for SISO systems [3]-[5] and then extended to MIMO systems [6][7].

So far, most MIMO receiver designs with direct-conversion RF treat I-Q imbalance as the only impairment, assuming all others have been estimated separately and compensated beforehand. Unfortunately, under the same optimization criterion, separate estimation of RF parameters leads to an inferior performance and may not be feasible if estimation of one parameter is seriously degraded at the presence of others.

In this paper, a new method of joint estimation and compensation of frequency, DC-offset, I-Q imbalance and channel is developed for MIMO receivers. Numerical results show that the proposed method has a very good performance; less than 2 dB loss is observed at bit-error rate 10^{-4}, as compared to the ideal receiver. In addition, a new training sequence along with a low-complexity architecture is proposed to reduce largely the receiver complexity with negligible performance loss.

2. SYSTEM MODEL AND RECEIVER ARCHITECTURE

Consider a block transmission with a prefix to avoid inter-block interference. The base-band transmitted signal from i-th transmit antenna is $s_i\,(t) = \sum_k \sum_{n=-N_g}^{N-1} s_{i,k}(n) g_T(t - (k(N+N_g)+n)T_s)$, where $s_{i,k}(n)$ is the transmitted k-th block symbols, N_g is the length of prefix, N is the length of useful data, $g_T(t)$ is the transmit filter with unit power, and T_s is the symbol time. For a data-aided estimation as considered here, $s_{i,k}(n)$ is known perfectly to the receiver. A total of $P \geq 1$ training blocks will be assumed in this paper, starting from the zero-th block ($k = 0$). Assume n_t and n_r are the number of transmit and receive antennas, respectively. The band-pass signal received from j-th antenna is $\tilde{y}_j(t) = \mathrm{Re}\{y_j(t)e^{\boldsymbol{j}2\pi(f_c+\Delta f)t}\} + \tilde{w}_j(t)$, where $y_j(t) = \sum_{i=1}^{n_t} s_i(t) \otimes h_{j,i}(t)$, $\boldsymbol{j} = \sqrt{-1}$, $h_{j,i}(t)$ denote the base-band channel response from i-th transmit to j-th receive antenna, $\tilde{w}_j(t)$ is the band-pass additive white Gaussian noise and $w_{0,j}(t)$ is its base-band equivalent, Δf is the frequency-offset which is same for all receive branches, and $\otimes$ denotes the operation of convolution. Taking into account

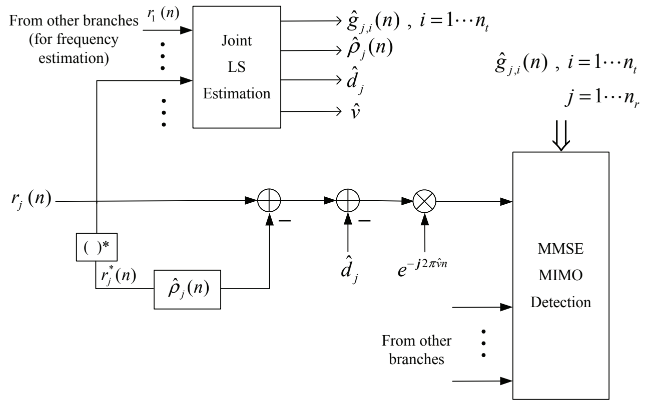

Fig. 1. MIMO receiver with joint LS estimation/compensation of RF parameters.

the frequency-independent I-Q imbalance, the mixer of j-th branch is $c_j(t) = 2\cos(2\pi f_c t) - \boldsymbol{j}2\alpha_j \sin(2\pi f_c t + \theta_j)$, where α_j and θ_j are the gain and phase imbalance, respectively. After down conversion, $h_j^I(t)$ and $h_j^Q(t)$ are the base-band filters used to remove out-of-band noise and high-frequency components. If $h_j^I(t) \neq h_j^Q(t)$, we say that there exists frequency-dependent I-Q imbalance, which is mostly encountered in a wide-band RF receiver because it is generally difficult to maintain base-band filters to have the same response over a wide frequency range [4]. After sampling, the end-to-end equivalent discrete system can be modeled as

$$\begin{aligned} r_j(n) &= h_{+,j}(n) \otimes \left[y_j(n)e^{\boldsymbol{j}2\pi\nu n} + w_{0,j}(n)\right] \\ &\quad + h_{-,j}(n) \otimes \left[y_j(n)e^{\boldsymbol{j}2\pi\nu n} + w_{0,j}(n)\right]^* + d_{0,j} \end{aligned} \quad (1)$$

where $h_{+,j}(n) = [h_j^I(n) + h_j^Q(n)\alpha_j e^{-\boldsymbol{j}\theta_j}]$, $h_{-,j}(n) = [h_j^I(n) - h_j^Q(n)\alpha_j e^{\boldsymbol{j}\theta_j}]$, $d_{0,j} = d_{0,j}^I + \boldsymbol{j} d_{0,j}^Q$ is the DC-offset, $\nu = \Delta f T_S$ is the normalized frequency-offset, and $w_{0,j}(n)$ is a zero mean additive white Gaussian noise.

Motivated by (1), one way to process the received signal is to cancel out firstly the self-image interference due to I-Q imbalance. By introducing the filter $\rho_j(n)$, we have

$$\begin{aligned} &r_j(n) - \rho_j(n) \otimes r_j^*(n) \\ &= e^{\boldsymbol{j}2\pi\nu n}\left(\sum_{i=1}^{n_t} s_i(n) \otimes \mathrm{g}_{j,i}(n)\right) + d_j + w_j(n) \end{aligned} \quad (2)$$

where $\mathrm{g}_{j,i}(n) \doteq h_{j,i}(n) \otimes (\mathrm{g}_j(n)e^{-\boldsymbol{j}2\pi\nu n})$, $\mathrm{g}_j(n) \doteq h_{+,j}(n) - \rho_j(n) \otimes h_{-,j}^*(n)$, $d_j = d_{0,j} - \rho_j(n) \otimes d_{0,j}^*$ and $w_j(n) = \mathrm{g}_j(n) \otimes w_{0,j}(n)$. To completely cancel out the self-image interference, $\rho_j(n) = (h_{+,j}^*(n))^{-1} \otimes h_{-,j}(n)$, where $(h_{+,j}^*(n))^{-1}$ is the inverse filter of $h_{+,j}^*(n)$.

Following (2), we propose the receiver architecture as in Figure 1; after canceling the self-image interference, DC-offset is compensated next, and then compensation for frequency-offset follows. The parameters $\rho_j(n)$, $\mathrm{g}_{j,i}(n)$, ν, and d_j will be estimated jointly in the least-squares sense, as to be discussed in the next section, and the MIMO detection is done with the MMSE detector, based on the estimated channel responses $\{\hat{\mathrm{g}}_{j,i}(n)\}$ along with the compensated received signals from all branches. Some other types of MIMO detectors can be used as well.

3. JOINT LEAST-SQUARES ESTIMATION

In order to do joint LS estimation, $\rho_j(n)$ and $\mathrm{g}_{j,i}(n)$ are approximated by the FIR filters $\boldsymbol{\rho}_j$ and $\mathbf{g}_{j,i}$ with the length $L_{\boldsymbol{\rho}}$ and $L_{\mathbf{g}}$, respectively.[1] In the following, $L_{\boldsymbol{\rho}}$ and $L_{\mathbf{g}}$ are assumed to be large enough (to contain 99% of the energy).

3.1. LS Estimators

Consider the case of $L_{\mathbf{g}} \leq N_g$; thus there will be no inter-block interference. Let $\mathbf{r}_j(k) = [r_{j,k}(0), r_{j,k}(1), \cdots, r_{j,k}(N-1)]^T$ be the useful part of the k-th block of received signal, where $r_{j,k}(n) \doteq r_j(k(N_g+N)+n)$, $\mathbf{R}_j(k)$ be the $N \times L_{\boldsymbol{\rho}}$ received signal matrix with (m,l)-th entry $[\mathbf{R}_j(k)]_{m,l} = r_{j,k}(m-l)$, for $0 \leq m \leq N-1$, $0 \leq l \leq L_{\boldsymbol{\rho}}-1$, and $\mathbf{S}_i(k)$ be the $N \times L_{\mathbf{g}}$ signal matrix with $[\mathbf{S}_i(k)]_{m,l} = s_{i,k}(m-l)$, $0 \leq m \leq N-1$, $0 \leq l \leq L_{\mathbf{g}}-1$. Furthermore, let $\mathbf{r}_j = [\mathbf{r}_j^T(0), \mathbf{r}_j^T(1), \cdots, \mathbf{r}_j^T(P-1)]^T$, $\mathbf{R}_j = [\mathbf{R}_j^T(0), \mathbf{R}_j^T(1), \cdots, \mathbf{R}_j^T(P-1)]^T$, $\mathbf{S}(k) = [\mathbf{S}_1(k), \mathbf{S}_2(k), \cdots, \mathbf{S}_{n_t}(k)]$, and $\mathbf{g}_j = [\mathbf{g}_{j,1}^T, \mathbf{g}_{j,2}^T, \cdots, \mathbf{g}_{j,n_t}^T]^T$. The total useful received signal for training is

$$\mathbf{r}_j - \mathbf{R}_j^* \boldsymbol{\rho}_j = \Gamma(\nu)\mathbf{S}\mathbf{g}_j + d_j\mathbf{1} + \mathbf{w}_j \quad (3)$$

where $\mathbf{S} = [\mathbf{S}^T(0), \mathbf{S}^T(1), \cdots, \mathbf{S}^T(P-1)]^T$, $\Gamma(\nu) = \mathrm{diag}\{\Gamma_0(\nu), \Gamma_1(\nu), \cdots, \Gamma_{P-1}(\nu)\}$ with diagonal matrix $\Gamma_k(\nu) = e^{\boldsymbol{j}2\pi k(N_g+N)\nu}\mathrm{diag}\{1, e^{\boldsymbol{j}2\pi\nu}, \cdots, e^{\boldsymbol{j}2\pi\nu(N-1)}\}$, and $\mathbf{w}_j$ is the corresponding noise vector. From (3), the joint least-squares estimates of all parameters are obtained by minimizing the cost function

$$\Lambda(\tilde{\nu}, \tilde{\boldsymbol{\rho}}_j, \tilde{d}_j, \tilde{\mathbf{g}}_j, j = 1\cdots n_r) = \sum_{j=1}^{n_r} \Lambda_j(\tilde{\nu}, \tilde{\boldsymbol{\rho}}_j, \tilde{d}_j, \tilde{\mathbf{g}}_j) \quad (4)$$

with $\Lambda_j(\tilde{\nu}, \tilde{\boldsymbol{\rho}}_j, \tilde{d}_j, \tilde{\mathbf{g}}_j) = \|\mathbf{r}_j - \mathbf{R}_j^*\tilde{\boldsymbol{\rho}}_j - \tilde{d}_j\mathbf{1} - \Gamma(\tilde{\nu})\mathbf{S}\tilde{\mathbf{g}}_j\|^2$. The LS solution of (4) is shown below, which can be obtained successively as in [5] and [11].

$$\hat{\boldsymbol{\rho}}_j = [(\mathbf{I} - \mathbf{C}(\hat{\nu}))(\mathbf{I} - \mathbf{1}\mathbf{f}^H(\hat{\nu}))\mathbf{R}_j^*]^\dagger(\mathbf{I} - \mathbf{1}\mathbf{f}^H(\hat{\nu}))\mathbf{r}_j \quad (5)$$

$$\hat{d}_j = \mathbf{f}^H(\hat{\nu})\left(\mathbf{r}_j - \mathbf{R}_j^*\hat{\boldsymbol{\rho}}_j\right) \quad (6)$$

$$\hat{\mathbf{g}}_j = (\Gamma(\hat{\nu})\mathbf{S})^\dagger(\mathbf{r}_j - \mathbf{R}_j^*\hat{\boldsymbol{\rho}}_j - \hat{d}_j\mathbf{1}) \quad (7)$$

and

$$\hat{\nu} = \arg\min_{\tilde{\nu}}\left\{\Lambda(\tilde{\nu}) = \sum_{j=1}^{n_r}\Lambda_j(\tilde{\nu})\right\} \quad (8)$$

[1] Throughout this paper, $(\cdot)^T$ and $(\cdot)^H$ represent the operations of conjugate transpose and transpose of a matrix or vector, respectively. In addition, $\mathbf{I}$ and $\mathbf{1}$ denote $\mathbf{I}_{N \cdot P}$ and $\mathbf{1}_{N \cdot P}$, respectively.

with

$$\Lambda_j(\tilde{\nu}) = \|(\mathbf{I}-\mathbf{C}(\tilde{\nu}))(\mathbf{I}-\mathbf{1f}^H(\tilde{\nu}))(\mathbf{r}_j - \mathbf{R}_j^* \hat{\boldsymbol{\rho}}_j(\tilde{\nu}))\|^2$$
$$= \mathbf{r}_j^H \mathbf{Q}(\tilde{\nu})\mathbf{r}_j - \left(\mathbf{R}_j^T \mathbf{Q}(\tilde{\nu})\mathbf{r}_j\right)^H \left[\mathbf{R}_j^T \mathbf{Q}(\tilde{\nu})\mathbf{R}_j^*\right]^{-1} \mathbf{R}_j^T \mathbf{Q}(\tilde{\nu})\mathbf{r}_j \quad (9)$$

where $(\mathbf{X})^\dagger$ denotes the pseudo-inverse of $\mathbf{X}$, $\mathbf{C}(\tilde{\nu}) = \Gamma(\tilde{\nu})\mathbf{B}\Gamma^H(\tilde{\nu})$ with $\mathbf{B} = \mathbf{S}(\mathbf{S}^H\mathbf{S})^{-1}\mathbf{S}^H, \mathbf{f}^H(\tilde{\nu}) \doteq ((\mathbf{I}-\mathbf{C}(\tilde{\nu}))\mathbf{1})^\dagger$, and $\mathbf{Q}(\tilde{\nu}) = (\mathbf{I}-\mathbf{1f}^H(\tilde{\nu}))^H(\mathbf{I}-\mathbf{C}(\tilde{\nu}))(\mathbf{I}-\mathbf{1f}^H(\tilde{\nu}))$. Generally, no close-form is available for $\hat{\nu}$; an exhaustive search has to be performed for the solution. However, the exhaustive search can be implemented by fast Fourier transform (FFT), as in [5], [8], and [11].

3.2. Low-Complexity Implementation

From (5), (6), and (9), it is observed that the calculation of the projection matrix $\mathbf{B} = \mathbf{S}\left(\mathbf{S}^H\mathbf{S}\right)^{-1}\mathbf{S}^H$ in $\mathbf{C}(\tilde{\nu}) = \Gamma(\tilde{\nu})\mathbf{B}\Gamma^H(\tilde{\nu})$ plays a key role in determining the complexity of the estimators of $\hat{\boldsymbol{\rho}}_j$, $\hat{d}_j$, and $\hat{\nu}$. Motivated by the design for SISO systems in [8], we design the following training format

$$\mathbf{S}^H = \left[\underbrace{\left[e^{\boldsymbol{j}\phi_0}\mathbf{D}\right]^H \left[e^{\boldsymbol{j}\phi_1}\mathbf{D}\right]^H \cdots \left[e^{\boldsymbol{j}\phi_{P-1}}\mathbf{D}\right]^H}_{P}\right] \quad (10)$$

with

$$\mathbf{D}^H = \left[\underbrace{\mathbf{A}^H\mathbf{A}^H\cdots\mathbf{A}^H}_{K}\right], \quad (11)$$

where $\mathbf{A}$ is an $n_tL_\mathbf{g} \times n_tL_\mathbf{g}$ full-rank matrix, $N = K \cdot n_tL_\mathbf{g}$ with $K \geq 1$, and $\{\phi_k\}_{k=0}^{P-1}$ are parameters to optimize the estimation performance. Example $\{\phi_k\}$ for $K = P = 2$ will be given in Section 4; nevertheless, the issue of optimum design of them will not be pursued any further in this paper. In this way, the projection matrix $\mathbf{B}$ becomes

$$\mathbf{B} = \begin{bmatrix} \mathbf{F} & e^{\boldsymbol{j}(\phi_0-\phi_1)}\mathbf{F} & \cdots & e^{\boldsymbol{j}(\phi_0-\phi_{P-1})}\mathbf{F} \\ e^{\boldsymbol{j}(\phi_1-\phi_0)}\mathbf{F} & \mathbf{F} & \cdots & e^{\boldsymbol{j}(\phi_1-\phi_{P-1})}\mathbf{F} \\ \vdots & \vdots & & \vdots \\ e^{\boldsymbol{j}(\phi_{P-1}-\phi_0)}\mathbf{F} & e^{\boldsymbol{j}(\phi_{P-1}-\phi_1)}\mathbf{F} & \cdots & \mathbf{F} \end{bmatrix} \quad (12)$$

with

$$\mathbf{F} = \frac{1}{P\cdot K}\begin{bmatrix} \mathbf{I}_{n_tL_\mathbf{g}} & \mathbf{I}_{n_tL_\mathbf{g}} & \cdots & \mathbf{I}_{n_tL_\mathbf{g}} \\ \mathbf{I}_{n_tL_\mathbf{g}} & \mathbf{I}_{n_tL_\mathbf{g}} & \cdots & \mathbf{I}_{n_tL_\mathbf{g}} \\ \vdots & \vdots & & \vdots \\ \mathbf{I}_{n_tL_\mathbf{g}} & \mathbf{I}_{n_tL_\mathbf{g}} & \cdots & \mathbf{I}_{n_tL_\mathbf{g}} \end{bmatrix}_{N\times N},$$

which contains K^2 matrices of $1/PK \cdot \mathbf{I}_{n_tL_\mathbf{g}}$. Recall that $\mathbf{A}$ is a full-rank square matrix, and therefore its projection matrix $\mathbf{A}\left(\mathbf{A}^H\mathbf{A}\right)^{-1}\mathbf{A}^H$ is equal to $\mathbf{I}_{n_tL_\mathbf{g}}$. Since $\mathbf{B}$ is a sparse matrix now, $\mathbf{C}(\tilde{\nu}) = \Gamma(\tilde{\nu})\mathbf{B}\Gamma^H(\tilde{\nu})$ can be calculated much easily. Hence, the complexity of CFO estimation is reduced by about $n_tL_\mathbf{g}$ times. However, the frequency range that can be estimated is also reduced by the same factor with this design. For SISO systems ($n_t = n_r = 1$) with $P = 1$, (10) degenerates to the one in [8]. Furthermore, for the undesirable case of $\phi_0 = \phi_1 = \cdots = \phi_{P-1}$, $\|(\mathbf{I}-\mathbf{C}(\tilde{\nu}))\mathbf{1}\|^2|_{\tilde{\nu}=0} = 0$ in (6). In other words, with this design of training sequence, it is not able to estimate DC-offset when frequency-offset is zero because $d_j\mathbf{1}$ is now located in the space spanned by the column vectors of $\mathbf{S}$.

In addition, from the simulation results in [11], frequency-dependent I-Q imbalance has little effect on the estimation of ν and d_j. With this observation, we propose the simplified estimators for frequency and DC-offset by setting $\mathbf{R}_j = \mathbf{r}_j$ in (9) and (6) as follows;

$$\hat{\nu}_s = \arg\min_{\tilde{\nu}}\left\{\Lambda_s(\tilde{\nu}) = \sum_{j=1}^{n_r}\Lambda_{j,s}(\tilde{\nu})\right\} \quad (13)$$

with $\Lambda_{j,s}(\tilde{\nu}) = \mathbf{r}_j^H\mathbf{Q}(\tilde{\nu})\mathbf{r}_j - \left|\mathbf{r}_j^T\mathbf{Q}(\tilde{\nu})\mathbf{r}_j\right|^2/\mathbf{r}_j^T\mathbf{Q}(\tilde{\nu})\mathbf{r}_j^*$ where no matrix inversion is needed, and

$$\hat{d}_{j,s}(\hat{\nu}_s) = \mathbf{f}^H(\hat{\nu}_s)\left(\mathbf{r}_j - \mathbf{r}_j^*\hat{\rho}_j(\hat{\nu}_s)\right) \quad (14)$$

with $\hat{\rho}_j(\hat{\nu}_s) = \mathbf{r}_j^T\mathbf{Q}(\hat{\nu}_s)\mathbf{r}_j/\mathbf{r}_j^T\mathbf{Q}(\hat{\nu}_s)\mathbf{r}_j^*$. It will be shown in Section 4 that no degradation on the BER performance when using the simplified estimators in (13) and (14).

Table 1. RF Parameter Value

(α_j, θ_j)	$(1.08, 5^o)$, $(1.09, 6^o)$, $(1.1, 7^o)$
(f_j^I, f_j^Q)	$(8, 8.3)$, $(7.9, 8.2)$, $(8.1, 8.4)$
Δf	uniform over -0.5 and 0.5 subcarrier spacing
$\lvert d_{0,j}\rvert$	0.2, 0.15, 0.1

4. NUMERICAL RESULTS

The performance of the proposed estimators is evaluated for an un-coded MIMO OFDM system. The system parameters are set as FFT length $N = 64$, cyclic prefix length $N_g = 16$, $L_\mathbf{g} = 16$, and symbol time $T_s = 50ns$. The base-band filters $h_j^I(t)$ and $h_j^Q(t)$ are modeled as 2rd order Butterworth filters with different cut-off frequencies (f_j^I, f_j^Q) MHz. The DC-offset is given by $d_{0,j} = |d_{0,j}|\cdot(1+\boldsymbol{j})/\sqrt{2}$ with signal power normalized to 1. Table 1 summarizes the total RF impairment parameters of the three receive antenna branches. The transmission is done on a packet-by-packet basis with the training portion consisting of two OFDM symbols at the beginning of each packet. An exponential decay multipath channel is considered with root-mean square delay spread $T_{RMS} = 50ns$. The length of channel is 10 taps, and each tap is zero mean independently complex Gaussian random variable.

Fig. 2 shows the impact of $L_{\boldsymbol{\rho}}$ on 64-QAM BER performance in the Rayleigh fading channel. The training sequence is the one given in [9] for the case of $n_t = 2$. Ideal receiver is the one with perfect RF compensation. Clearly, the modeling error due to using a small $L_{\boldsymbol{\rho}}$ incurs error floor in BER

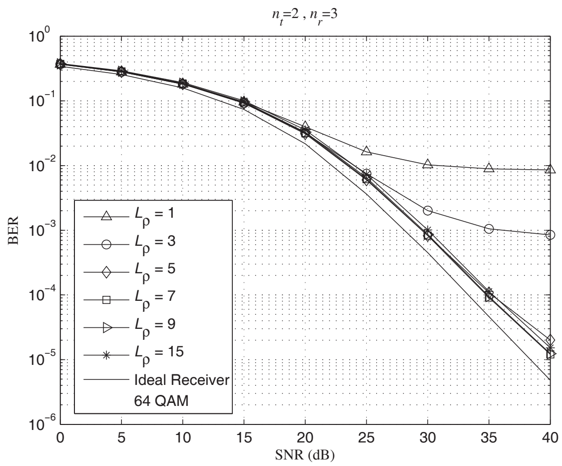

Fig. 2. The effects of $L_{\boldsymbol{\rho}}$ on BER performance in Rayleigh fading channels.

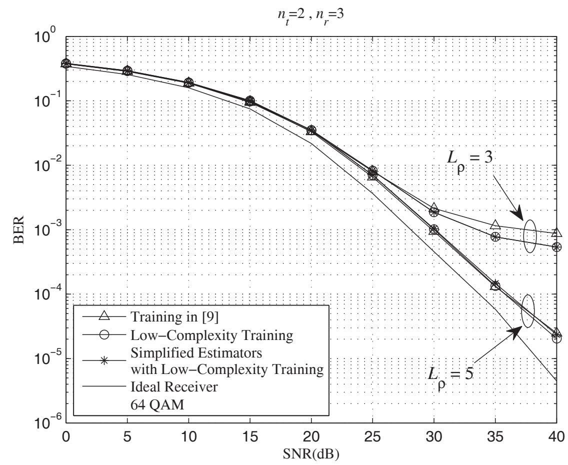

Fig. 3. BER performance with low-complexity and/or simplified estimators in Rayleigh fading channels.

performance, particularly compared with the case of $L_{\boldsymbol{\rho}} = 1$ proposed in [5]. On the other hand, too large a $L_{\boldsymbol{\rho}}$, e.g., $L_{\boldsymbol{\rho}} = 15$ degrades slightly the BER performance due to the extra noise induced by using a large filter length. In Fig. 3, we show the BER performance by using low-complexity training sequence and/or simplified frequency and DC-offset estimators. The square matrix $\mathbf{A}$ of the low-complexity training sequence is designed as $5230F641_H$ given in [10] for the first transmit antenna and its circular shift by $L_{\mathbf{g}}$ for the second transmit antenna. In addition, $K = P = 2$, $\phi_0 = 0$, and $\phi_1 = \pi/2$. The low-complexity training works very well and almost no performance loss is observed with the low-complexity designs.

5. CONCLUSION

The theory of joint least-squares estimation and compensation of frequency, I-Q imbalance, DC-offset and channel is developed for MIMO receivers with direct-conversion RF architecture. Both frequency-independent and dependent I-Q imbalances are included. Special attention is paid to the implementation complexity issue; several measures are proposed to reduce receiver complexity, including a special training-sequence design and low-complexity estimators for frequency and DC-offset. Simulation results show that the performance degradation is negligible when using the low-complexity designs.

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