Preamble Design for Joint Estimation of Channel and I/Q Imbalance in MIMO-OFDM Systems

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***Abstract*—In this paper, preamble design for estimation of frequency selective channels and In-phase/Quadrature-phase (I/Q) imbalance in multiple input multiple output orthogonal frequency division multiplexing (MIMO-OFDM) systems is proposed. First we utilize convex optimization to optimize power of all active subcarriers, then we employ cross entropy (CE) optimization techniques to select optimal preamble sequence that minimizes the channel estimate mean squared error (MSE) while suppressing the effect of I/Q mismatch. To mitigate inter-antenna interferences, disjoint preamble sequences are utilized for each transmit antenna. An algorithm to guarantee that the preamble sequences are disjoint for each transmitter is proposed. Numerical simulations are provided to verify the advantages and effectiveness of the proposed preamble sequences over the conventional sequences.**

I. Introduction

The advantages offered by combining orthogonal frequency division multiplexing (OFDM) with multiple input multiple output (MIMO) techniques are manifold. The most remarkable of them are robustness of OFDM systems against frequency selective fading channels, obtained by converting the OFDM channel into flat fading subchannels [1], [2] and the significant information capacity gain together with improved BER performance of the MIMO systems [1], [3].

Additionally, the adoption of direct conversion receivers (DCRs) in MIMO-OFDM systems is the promising technic for future wireless communication systems. DCRs are prominent candidates for physical layer applications due to its low-cost and low-power implementation on silicon as compared to the heterodyne receivers. DCRs can convert the RF signals directly into baseband without any intermediate frequencies (IF).

In contrast to these appealing attributes, DCRs are sensitive to In-phase/Quadrature-phase (I/Q) imbalance when compared to heterodyne receivers. In DRCs, the baseband signals are more severely distorted by imbalances within the I and Q branches [4]–[9]. Likewise, contrary to a single input single output (SISO) system where only one channel is to be estimated, a MIMO system with N_t transmit and N_r receive antennas requires $N_t \times N_r$ channels to be estimated [2].

The I/Q imbalance is a critical impairment as it causes inter-carrier interferences (ICI) in OFDM systems that results to severe degradation of the system's performance. Generally I/Q imbalances exist at both transmitter and receiver, and may vary over time (e.g., due to sudden changes in temperature or the dynamics of the cooperating/relaying users) [10]–[12].

The frequency selective nature of the MIMO channels together with I/Q imbalance are the main factors that degrade the performance of OFDM systems. Since I/Q imbalance problem is originated from the unavoidable differences in the relative amplitudes and phases of the physical analog I and Q signal paths, the effects of I/Q imbalances in OFDM systems cannot be rectified by increasing signal power, thus, to obtain better quality of the high rate communications systems, efficient channel estimation and I/Q imbalance compensation techniques are crucial [8], [9], [11], [12].

Several preamble sequence designs for channel estimation and I/Q imbalance compensation for OFDM systems have been predominantly developed for both MIMO and SISO systems (see [7], [11]–[13] and the references there in). However, most of the existing techniques utilize all active subcarriers to estimate the channel and the IQ imbalances, and therefore call for proper phase design technique to ensure that I/Q imbalance is suppressed.

In order to mitigate the effect of co-channel interference between the transmitting antennas, it is necessary for the training signals from each antenna to be orthogonal. In [14], [15], the orthogonality of the preamble sequences is achieved by ensuring that the preamble sequences of one antenna are disjoint from the preamble sequences of any other antenna.

Although the orthogonality of the training signals mitigates the problem of inter-antenna interference, it does not guarantee the suppression of the distortion caused by the I/Q imbalance. In [4], an estimator that utilizes special training pattern for both distortion and channel estimation is presented. The adopted training patterns in [4] can be utilized in single input single output (SISO) as well as in MIMO systems without introducing inter-antenna interferences. The scheme effectively estimates I/Q imbalance by setting some of the training subcarriers to zero. The disjoint training sequences can be obtained by utilizing subcarriers allocated on the lower side of the active subcarriers for one antenna and upper subcarriers for the other antenna. Even though I/Q imbalance is efficiently estimated and compensated using the special training pattern in [4], but the channel estimate mean squared error (MSE) is poor since for each antenna, channel is estimated by training sequences allocated on one side of the active subcarrier band while subcarriers on the other side are nulled.

This paper presents efficient preamble sequence design for the estimation of frequency-selective channels and I/Q imbalances in MIMO-OFDM system while mitigating inter-antenna interferences. We utilize cross entropy (CE) optimization techniques together with convex optimization to design training sequences that minimizes the channel MSE while suppressing the effect of I/Q imbalance. We propose a CE based algorithm for selecting position of training symbols when some active subcarriers are nulled. Unlike [4], where special training sequence is obtained by setting active subcarriers on the lower or upper side of the active band to zero, here we propose a scheme for selecting the position of preamble sequences to obtain better channel estimate MSE. Simulation results are provided to substantiate the effectiveness of the proposed designs.

II. System Model

In this section we briefly provide an overview of the baseband model that captures the effects of frequency independent (FI) and frequency dependent (FD) I/Q imbalances at both transmitter and receiver side, for a MIMO-OFDM wireless system with N_t transmit and N_r receive antennas over frequency selective channels.

The following notations are used in the description of the system. The superscripts $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^{\mathcal{H}}$ represent conjugate, transpose and the conjugate transpose (hermitian) respectively. The operator $\otimes$, denotes the convolution operation. Other operators that are used will be defined whenever used.

First, we consider single input single output (SISO) system with FI gain and phase offsets of the I and Q branches at the transmitter side denoted as $\{a_t^I, a_t^Q\}$ and $\{\theta_t^I, \theta_t^Q\}$ respectively. The corresponding impulse shaping filters, which includes amplifiers, digital to analog converters (DAC), pulse shapers and FD imbalances for the I and Q branches of the transmitter are denoted as $\{g_t^I(t), g_t^Q(t)\}$. The equivalent receiver side parameters are denoted in the same manner with subscript t replaced by r, i.e., a_r^I,a_r^Q, θ_r^I, θ_r^Q, $g_r^I(t)$and $g_r^Q(t)$. The low pass equivalent channel is denoted as $h(t)$. The system with I/Q imbalances consists of the direct and the mirror system. The impulse responses of the direct and mirror system at the transmitter and receiver are denoted as $g_T^D(t)$, $g_T^M(t)$ and $g_R^D(t)$, $g_R^M(t)$ respectively and are related to the I/Q imbalance parameters as [5], [7]

$$g_T^D(t) = \frac{1}{2}\left[a_t^I e^{j\theta_t^I} g_t^I(t) + a_t^Q e^{j\theta_t^Q} g_t^Q(t)\right], \quad (1)$$

$$g_T^M(t) = \frac{1}{2}\left[a_t^I e^{j\theta_t^I} g_t^I(t) - a_t^Q e^{j\theta_t^Q} g_t^Q(t)\right], \quad (2)$$

$$g_R^D(t) = \frac{1}{2}\left[a_r^I e^{-j\theta_r^I} g_r^I(t) + a_r^Q e^{-j\theta_r^Q} g_r^Q(t)\right], \quad (3)$$

$$g_R^M(t) = \frac{1}{2}\left[a_r^I e^{j\theta_r^I} g_r^I(t) - a_r^Q e^{j\theta_r^Q} g_r^Q(t)\right]. \quad (4)$$

In a MIMO-OFDM system, RF links for different antennas may lead to distinct I/Q imbalances. Thus, the equivalent direct and mirror channel of the ith transmit antenna and the jth receive antenna with its corresponding transmit and receive filters as well as the channel $h_{ij}(t)$ are given by

$$\xi_{ij}(t)=g_{T,i}^D(t)\otimes h_{ij}(t)\otimes g_{R,j}^D(t)+(g_{T,i}^M)^*(t)\otimes h_{ij}^*(t)\otimes g_{R,j}^M(t), \quad (5)$$

$$\chi_{ij}(t)=g_{T,i}^M(t)\otimes h_{ij}(t)\otimes g_{R,j}^D(t)+(g_{T,i}^D)^*(t)\otimes h_{ij}^*(t)\otimes g_{R,j}^M(t). \quad (6)$$

For an OFDM symbol with N number of subcarriers, we denote the frequency domain transmitted data from the ith transmit antenna as $\boldsymbol{X}_i = [X_i(1), \ldots, X_i(N)]$. Each OFDM symbol is passed through the IFFT operation to obtain the time domain signal given by $\boldsymbol{x}_i = \boldsymbol{F}^{\mathcal{H}}\boldsymbol{X}_i$, where

$$[F]_{m,n} = \frac{1}{\sqrt{N}} e^{\frac{-j2\pi mn}{N}}, \quad m,n = \{0, 1, \cdots N-1\}.$$

The discrete time versions of the direct and the mirror channels for the ith transmit antenna and the jth receive antenna $\xi_{ij}(t)$ and $\chi_{ij}(t)$are denoted by $\boldsymbol{\xi}_{ij}$ and $\boldsymbol{\chi}_{ij}$ respectively and consist of maximum L taps. We assume that the number of the cyclic prefix samples N_{cp} is greater than channel length L (i.e., $N_{cp} \geq L$) in order to preserve the orthogonality of the tones. For a system with I/Q imbalance, if the direct channel sees an input signal $x_i(t)$, the mirror channel sees an input $x_i^*(t)$. Note that, the DFT of the complex conjugate of a sequence is related to the DFT of the original sequence through a mirrored relation (for $1 \leq n \leq N$ and $1 \leq k \leq N$) [4], that is

$$\begin{aligned} x_i(n) &\xrightarrow{DFT} X_i(k), \\ x_i^*(n) &\xrightarrow{DFT} X_i^*(N-k+2). \end{aligned} \quad (7)$$

At the jth receive antenna, the received time-domain signal for one OFDM symbol after CP removal can be written as

$$\boldsymbol{y}_j = \sqrt{N}\sum_{i=0}^{N_t-1} \boldsymbol{F}^{\mathcal{H}}\left(D(\boldsymbol{X}_i)\boldsymbol{F}_L\boldsymbol{\xi}_{ij} + D(\boldsymbol{X}_i^{\#})\boldsymbol{F}_L^*\boldsymbol{\chi}_{ij}\right) + \boldsymbol{v}_j, \quad (8)$$

where $D(\cdot)$ denotes the diagonal matrix of the vector and $\boldsymbol{F}_L$ is an $N \times L$ sub-matrix of the DFT matrix $\boldsymbol{F}$. The elements of the mirror vector $\boldsymbol{X}_i^{\#}$ for $\boldsymbol{X}_i = \{X_i(1), X_i(2), \ldots, X_i(N)\}$ are given by $X_i^{\#} = \{X_i^*(1), X_i^*(N), \ldots, X_i^*(2)\}$.

Note that, for MIMO channels that are independent or with unknown joint statistics, the equivalent direct and mirror channels can be estimated at each receive antenna independently. Thus, it makes sense to design the preamble sequence by considering only one receive antenna. In the following, we omit the receive antenna index j.

Define $\boldsymbol{P} \triangleq \boldsymbol{F}\boldsymbol{F}^T$, then $\boldsymbol{P}\boldsymbol{P}^T = \boldsymbol{I}_N$ and $\boldsymbol{P}\boldsymbol{F}_L^* = \boldsymbol{F}_L$, then from this definitions [5],

$$\begin{aligned} D(\boldsymbol{X}_i^{\#}) &= D(\boldsymbol{P}\boldsymbol{X}_i^*) = \boldsymbol{P}D^{\mathcal{H}}(\boldsymbol{X}_i)\boldsymbol{P} \\ &= \mathrm{diag}\{X_i^*(1), X_i^*(N), \ldots, X_i^*(2)\}. \end{aligned} \quad (9)$$

Suppose that there are some null edge subcarriers, let $\mathcal{K}_a$ and $N_a = |\mathcal{K}_a|$ be a set and the number of active subcarriers respectively. Then, given $\boldsymbol{X}_{i,a}$ as a transmitted OFDM block in frequency domain at the active subcarriers, using the definition in (9) the received signal can be expressed as

$$\boldsymbol{y} = \sqrt{N}\sum_{i=0}^{N_t-1} \boldsymbol{F}_a^{\mathcal{H}}\big(D(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\boldsymbol{\xi}_i + \boldsymbol{P}_a D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\boldsymbol{\chi}_i\big) + \boldsymbol{v}, \tag{10}$$

where $\boldsymbol{F}_a$ is an $N_a \times N$ and $\boldsymbol{F}_{L,a}$ is an $N_a \times L$ sub-matrices of $\boldsymbol{F}$ and $\boldsymbol{F}_L$ respectively corresponding to N_a number of active subcarriers. Note that, $\boldsymbol{P}_a = \boldsymbol{F}_a \boldsymbol{F}_a^T$.

The frequency-domain representation of the received time signals in (10) is given by

$$\boldsymbol{Y} = D(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\boldsymbol{\xi}_i + \boldsymbol{P}_a D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\boldsymbol{\chi}_i + \boldsymbol{w}, \tag{11}$$

where $\boldsymbol{w} = \boldsymbol{F}_a \boldsymbol{v}$.

For coherent detection, the direct channel $\boldsymbol{\xi}_i$ and the mirror channel $\boldsymbol{\chi}_i$ need to be estimated at the receiver. Since in practical system the statistics of the channel and the imperfections are unknown, we consider the least square channel estimators. The estimates of the direct and mirror channel impulse response vectors are given by [5]

$$\hat{\boldsymbol{\xi}}_i = \underbrace{\left(\boldsymbol{F}_{L,a}^{\mathcal{H}} D^{\mathcal{H}}(\boldsymbol{X}_{i,a}) D(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\right)^{-1} \boldsymbol{F}_{L,a}^{\mathcal{H}} D^{\mathcal{H}}(\boldsymbol{X}_{i,a})}_{\mathcal{M}}\boldsymbol{Y}, \tag{12}$$

$$\hat{\boldsymbol{\chi}}_i = \underbrace{\left(\boldsymbol{F}_{L,a}^{\mathcal{H}} D(\boldsymbol{X}_{i,a}) D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\right)^{-1} \boldsymbol{F}_{L,a}^{\mathcal{H}} D(\boldsymbol{X}_{i,a})\boldsymbol{P}_a^{\mathcal{H}}}_{\mathcal{N}}\boldsymbol{Y}. \tag{13}$$

Substitution of (11) to (12) and (13) gives

$$\hat{\boldsymbol{\xi}}_i = \boldsymbol{\xi}_i + \mathcal{M}\boldsymbol{P}_a D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\boldsymbol{\chi}_i + \mathcal{M}\boldsymbol{w}, \tag{14}$$

$$\hat{\boldsymbol{\chi}}_i = \boldsymbol{\chi}_i + \mathcal{N} D(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\boldsymbol{\xi}_i + \mathcal{N}\boldsymbol{w}. \tag{15}$$

To measure the impact of the channel noise and I/Q mismatch, the MSE caused by both effects is derived.

$$||\hat{\boldsymbol{\xi}} - \boldsymbol{\xi}||_2^2 = \underbrace{\sigma_w^2 \frac{1}{N_t}\sum_{i=0}^{N_t} \operatorname{trace}\left(\mathcal{M}\mathcal{M}^{\mathcal{H}}\right)}_{MSE_w} +$$

$$\underbrace{\sigma_\chi^2 \frac{1}{N_t}\sum_{i=0}^{N_t} \operatorname{trace}\left(\mathcal{M}\boldsymbol{P}_a D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\boldsymbol{F}_{L,a}^{\mathcal{H}} D(\boldsymbol{X}_{i,a})\boldsymbol{P}_a^{\mathcal{H}}\mathcal{M}^{\mathcal{H}}\right)}_{MSE_{\boldsymbol{\chi}}} \tag{16}$$

Let us define the matrix $\boldsymbol{\Lambda}_{i,a} = D(\boldsymbol{X}_{i,a})D^{\mathcal{H}}(\boldsymbol{X}_{i,a})$, then the MSE caused by gaussian noise can be written as

$$MSE_w = \sigma_w^2 \frac{1}{N_t}\sum_{i=0}^{N_t} \operatorname{trace}\left[\boldsymbol{F}_{L,a}\left(\boldsymbol{F}_{L,a}^{\mathcal{H}}\Lambda_{i,a}\boldsymbol{F}_{L,a}\right)^{-1}\boldsymbol{F}_{L,a}^{\mathcal{H}}\right]. \tag{17}$$

III. PREAMBLE SEQUENCE DESIGN

In the proposed design, we consider a MIMO scenario where disjoint pilot symbols are utilized for each transmit antenna. That is, if training subcarrier is allocated at the kth subcarrier for one antenna no other antenna allocates any signal in this subcarrier. Hence, inter-antenna interference can be avoided. Preamble sequences can be inserted in any locations in the active band which match this criterion.

We utilize some criteria to design preamble sequences that minimizes the channel estimate MSE while suppressing the interference replica caused by I/Q imbalances. From equation (14), it can be seen that, the second and third terms are interferences and their contribution to channel estimation errors is reduced if they are reduced to zero. The last term is due to noise and is considered to be white. Thus we need to design preamble sequences that suppress the interference caused by the second term. Substituting $\mathcal{M}$ in (14), the second term of (14) can be represented as

$$\left(\boldsymbol{F}_{L,a}^{\mathcal{H}} D^{\mathcal{H}}(\boldsymbol{X}_{i,a}) D(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\right)^{-1}\boldsymbol{F}_{L,a}^{\mathcal{H}} D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{P}_a D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a}\boldsymbol{\chi}_{\boldsymbol{i}}$$

To suppress the interference replica, the preamble sequences should be designed in such away that

$$\boldsymbol{F}_{L,a}^{\mathcal{H}} D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{P}_a D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a} = 0. \tag{18}$$

The easiest way of meeting this condition is to set the active subcarriers in the lower or upper band of the central DC subcarrier to zero and allocate power to the subcarriers on one sideband which is not nulled as in [4]. However, this will lead to poor estimate of the channel since channel is only estimated by either upper or lower subcarriers. To ensure better MSE performance, both training allocation and power distribution need to be careful considered. In this paper we propose an algorithm for selecting training position that gives better channel estimate MSE performance while satisfying the condition in (18).

For a given energy to be utilized for channel estimation, we normalize the sum of preamble sequence power such that

$$\sum_{k\in\mathcal{K}_a}\Lambda_{i,k} = \sum_{k=1}^{N_a}\lambda_{i,k} = 1. \tag{19}$$

To minimize the channel estimate MSE, we need is to determine the optimal $\boldsymbol{\lambda}_i = [\lambda_{i,1}, \ldots, \lambda_{i,N_a}]^T$. The optimization problem under these constraints can be stated as

$$\begin{aligned}
\underset{\boldsymbol{\lambda}_i}{\text{minimize}} \quad & \operatorname{trace}\left[\boldsymbol{F}_{L,a}^{\mathcal{H}}\left(\frac{1}{\sigma_w^2}\boldsymbol{F}_{L,a}^{\mathcal{H}}\Lambda_{i,a}\boldsymbol{F}_{L,a}\right)^{-1}\boldsymbol{F}_{L,a}\right] \\
\text{subject to} \quad & [1,\ldots,1]\,\boldsymbol{\lambda}_i \le 1, \quad \boldsymbol{\lambda}_i \succeq 0, \\
& \boldsymbol{F}_{L,a}^{\mathcal{H}} D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{P}_a D^{\mathcal{H}}(\boldsymbol{X}_{i,a})\boldsymbol{F}_{L,a} = 0.
\end{aligned} \tag{20}$$

The problem in (20) above is a convex optimization problem and can be solved by existing convex packages. We adopt the method presented in [16], to optimally allocate power to the preamble sequences in order to minimize the MSE caused by the gaussian noises.

To design the special training sequences, we need to set some subcarriers to zero as proposed in [4]. Thus, once power distribution to all active subcarriers is obtained, we utilize cross entropy (CE) method to select preamble sequence that minimizes the channel estimate MSE and eliminate the interference replica.

In [4], it has been shown that, preamble sequence that utilizes some null subcarriers can effectively estimate the

I/Q imbalance. However there is no proposed algorithm for selecting optimal preamble sequences. In the following we propose an algorithm for selecting training position for a system with multiple transmit antennas.

Assume $\mathcal{K}_{p_i}$ to be a selected set of preamble sequences from the active subcarrier set $\mathcal{K}_a$. Our optimal training sequence design as a combinatorial optimization problem can be formulated as

$$\mathcal{K}_{p_i}^{\star} = \arg \min_{\mathcal{K}_p^m \in \boldsymbol{\Omega}} \mathcal{C}_{sel}(\mathcal{K}_p^m), \tag{21}$$

where

$$\mathcal{C}_{sel}(\mathcal{K}_p^m) = \text{trace}\left[\boldsymbol{F}_{L,p}\left(\boldsymbol{F}_{L,p}^{\mathcal{H}}\Lambda_p\boldsymbol{F}_{L,p}\right)^{-1}\boldsymbol{F}_{L,p}^{\mathcal{H}}\right], \tag{22}$$

represents the channel estimate MSE of the training set $\mathcal{K}_p^m$ and $\mathcal{K}_{p_i}^{\star}$ is the global optimal set of the objective function. The set $\mathcal{K}_p^m$ is defined by

$$\mathcal{K}_p^m = \mathcal{K}_a(\{I_k\}_{k=1}^{|\mathcal{K}_a|} = 1), \quad I_k \in \{0,1\}, \quad m = 1, \dots, M, \tag{23}$$

where the indicator function I_k shows whether a subcarrier at the kth position is selected. The set of all $M = \binom{|\mathcal{K}_a|}{|\mathcal{K}_p|}$ possible subsets is denoted by $\Omega = \{\mathcal{K}_p^1, \dots \mathcal{K}_p^M\}$, where $\binom{a}{b}$ denotes the possible combination set.

Applying the CE to solve (21), the first step is transforming the deterministic optimization problem (21) into a family of stochastic sampling problems. Since the considered problem is on a discrete case, a family of Bernoulli probability density functions associated with the training sequence selection vector, $\boldsymbol{\omega} = \left[\omega_1, \omega_2, \dots \omega_{|\mathcal{K}_a|}\right]$, $\omega_k \in \{0,1\}$, is given by

$$f(\boldsymbol{\omega}, \boldsymbol{p}) = \prod_{k=1}^{|\mathcal{K}_a|} p_k^{1_k(\boldsymbol{\omega})}(1 - p_k^{1-1_k(\boldsymbol{\omega})}), \tag{24}$$

where $\boldsymbol{p} = [p_1, p_2, \dots p_{|\mathcal{K}_a|}]$ is a probability vector whose p_k entry indicates the probability of selecting the kth subcarrier, and the indicator function $1_k(\boldsymbol{\omega}) \in \{0,1\}$ indicates whether the kth element of ω_k(the kth tone) is selected. If the ω_k is selected, then $1_k(\omega_k) = 1$. Each element is modeled as an independent Bernoulli random variable with probability mass function $p(\omega_k = 1) = p_k$, and $p(\omega_k = 0) = 1 - p_k$, for $k = 1, \dots |\mathcal{K}_a|$.

The problem of selecting training sequences with CE optimization is efficiently solved for single transmit antenna in [5]. In this paper we extend the scheme in [5] to accommodate multiple transmit antennas. For more details on the application of CE optimization in selection of optimal preamble set, see [5] and the references therein. Note that, not all schemes employed in SISO systems can be easily adopted in MIMO systems, some algorithms utilized in SISO systems when extended to MIMO systems may perform poorly due to inferior performance of the training symbols designed for the additional transmit antennas.

We propose an algorithm for selecting disjoint preamble sequences for N_t transmit antennas while ensuring that the interference replica is suppressed for each antenna. The following algorithm summarizes our proposed design.

Algorithm 1: Disjoint preamble sequences for N_t transmit antennas

1. Optimize power of the $\mathcal{K}_a$ active subcarriers, using convex optimization.
2. Initialize a temporary set $\mathcal{K}_t = \mathcal{K}_a$..
3. **for** $(i = 1$ to N_t) **do**
4. Apply CE algorithm to select optimal preamble sequence $\mathcal{K}_{p_i}^{\star}$ that minimizes the channel estimate MSE while suppressing the interference replica.
5. Save the selected preamble position and power of the subcarriers
6. Remove $\mathcal{K}_{p_i}^{\star}$ from $\mathcal{K}_t$, that is, update $\mathcal{K}_t = \mathcal{K}_t \setminus \mathcal{K}_{p_i}^{\star}$.
7. Optimize the power of the remaining subcarriers using SDP
8. **end for**

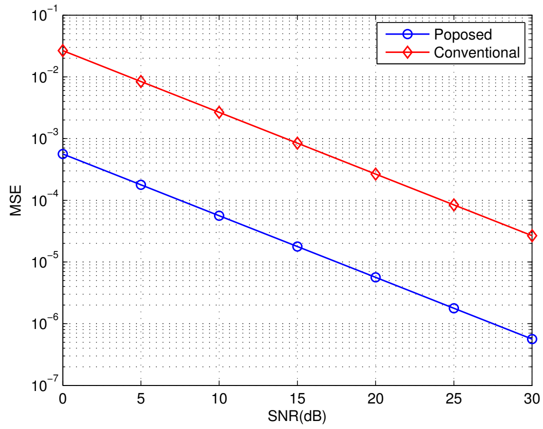

Fig. 1. Comparison of the Channel estimate MSE between the proposed preamble sequences and the equal powered trainings, $N_t = 2$

IV. Design Examples

We conduct computer simulations to demonstrate the effectiveness of our proposed schemes. The efficacy of the proposed design is evaluated by the mean squared error (MSE) as well as bit error rate (BER) performances. An OFDM transmission frame with $N = 128$ and $N_a = 100$ is considered. The remaining 28 subcarriers, 14 are null in the lower frequency guard band while 13 are nulled in the upper frequency guard band and one is the central DC null subcarrier. To estimate and compensate for the I/Q imbalances, we adopt the compensator proposed in [4] that utilizes special training pattern.

For BER simulation, frequency selective channel with L taps is considered. Each channel tap is i.i.d. complex Gaussian with zero mean and the exponential power delay profile is given by the vector $\boldsymbol{\zeta} = [\zeta_0 \dots \zeta_{L-1}]$ where $\zeta_l = \mathcal{R}e^{-l/2}$, and $\mathcal{R}$ is a constant selected so that $\sum_{l=0}^{L-1} \zeta_l = 1$.

Fig. 1 depicts the MSE performance of the two designs for different signal to noise ratio (SNR) for $N_t = 2$. From the plot it is clear that our proposed design outperforms equal powered training allocated either on the upper or lower sideband due to the fact that, for the equal powered trainings channel for each transmitter and receiver link is only estimated by upper subcarriers. This suggests that to ensure better MSE performance, both training allocation and power distribution need to be careful considered.

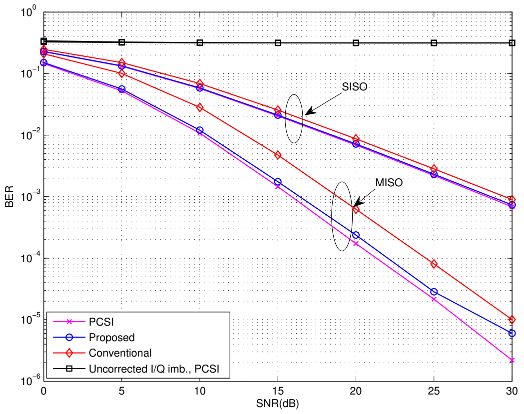

Fig. 2. BER versus SNR results for 64QAM constellation, phase imbalance of 2°, amplitude imbalance of 1dB and $L = 4$

In the following, we demonstrate the performance of our preamble sequences by considering the BER performance of both SISO and Alamouti STBC with two transmit antennas and one receive antenna (MISO). It is well known that, to obtain better BER performance, proper compensation of I/Q imbalance and accurate channel estimates are of primary important. Fig. 2 depicts the BER performance of the designed preamble sequences together with the results of the known channel state information with and without any IQ imbalance for $L = 4$. From the results it is clear that the BER performance of our proposed design is comparable to that of the known channel state information for both SISO and MISO case. This demonstrates the efficiency of our designed preamble sequences.

Although total power allocated to the preamble sequences in each transmitter is the same for the two designs, our proposed design outperforms the equal powered training symbols. This is because the proposed design optimally allocate the preamble sequences to reduce the channel estimate MSE. To obtain better BER as well as MSE performances require optimal allocation of preamble sequences.

The BER results also demonstrate the case where channel state information is known but I/Q imbalance is not compensated (see the BER curve with a legend uncorrected I/Q imb., PCSI). From the result it is clear that without I/Q imbalance compensation, the BER performance is poor even at high SNR. This suggests that, the effects of I/Q imbalances in OFDM systems cannot be corrected by increasing the signal power, thus, regardless of the channel state information (known or estimated), to obtain better BER performance calls for compensation of the I/Q imbalance.

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