MISO JOINT SYNCHRONIZATION-PILOT DESIGN FOR OFDM SYSTEMS

John E. Kleider[†‡], *Xiaoli Ma*[‡], *Robert J. Baxley*[‡] *and G. Tong Zhou*[‡]

[†]General Dynamics C4 Systems, Scottsdale, AZ, USA
[‡]School of Electrical and Computer Engineering, Georgia Institute of Technology, Atlanta, GA, USA

ABSTRACT

In this paper we apply a new joint synchronization-pilot sequence (JSPS) optimization design technique to multiple transmitter OFDM systems. One objective in the design of the JSPS for MISO transmissions is to eliminate the requirement for the different preamble fields used for coarse and fine synchronization and channel training. In this work, independent JSPSs are designed for each transmitter in a multiple antenna system, but are transmitted simultaneously from each antenna. Consequently, JSPSs can potentially reduce the overhead of multiple antenna preamble transmissions (improving bandwidth efficiency). The objective of this paper is to determine the performance advantages of joint carrier frequency offset (CFO) and channel estimation using only the pilot portion of the JSPS. By jointly optimizing the position and power of each pilot across the transmitter antennas in Rayleigh fading channels with CFO, we show > 15 dB SNR improvement in the performance of the channel estimator.

Index Terms**—**Orthogonal frequency division multiplexing (OFDM), pilot symbol assisted modulation (PSAM), pilot design, MISO systems

1. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is an attractive technique for high data rate transmission in multipath channels [1], while multiple transmit antennas can be utilized to improve diversity performance and thus reduce the receiver demodulation bit error rate (BER) [2]. In many wireless applications it is desirable to minimize transmitter power in multiple antenna OFDM systems and thus coherent modulation is commonly utilized for this purpose (as compared to differential modulation which requires 2-3 dB additional power)[3]. Coherent OFDM, however, requires accurate synchronization and channel estimation; otherwise the power savings of coherent modulation is diminished and becomes less attractive as a preference over differential OFDM [4]. Because of coherent OFDM's sensitivity to synchronization error, information packets are typically preceded by coarse/fine synchronization and channel estimation preambles, which are commonly separated into two fields, each field with a different sequence [5]. OFDM systems have been shown to be particularly sensitive to carrier frequency offset (CFO), which causes inter-carrier interference (ICI) and if left uncorrected can result in severe BER degradation at the receiver [6]. In addition, unresolved CFO will also degrade channel estimation and equalization due to ICI and uncorrected phase rotation over time.

In this work, we develop a single sequence preamble (as opposed to two separate sequences) for preamble-based OFDM systems, where this sequence can be used for coarse synchronization (timing- and frequency-acquisition), fine synchronization (fine timing and frequency estimation) and channel estimation. We denote this sequence as a joint synchronization-pilot sequence (JSPS). JSPSs have been shown to be effective for frequency and timing acquisition [7] and sampling offset can be estimated jointly during the channel estimation phase [8]. We assume that the preambles for each antenna are transmitted simultaneously in one time slot (to minimize preamble transmission overhead) and thus will interfere with each other. The pilot part of the JSPSs used for channel training is designed to be orthogonal in frequency across all transmitter antennas. This characteristic enables robust and low complexity CFO and channel estimation at the single antenna receiver.

A number of joint detection methods have been proposed in the literature for OFDM single-input single-output (SISO) systems [9] [10] and for OFDM transmit diversity systems [11] with CFO. Similar to the idea in [11], the preamble JSPS does not assume any specific space-time code as opposed to the space-time block-code (STBC) based iterative detection method in [12]. The JSPS is designed to enable decoupling of the CFO and channel estimation from the symbol decoding, which is attractive for low receiver complexity, compared to blind and semi-blind approaches [13][6]. The scope of this paper is to show that, within each JSPS, optimized pilot positions and power can yield improved CFO and channel estimation performance in multiple OFDM transmitter systems.

This paper is organized as follows. Section 2 describes the OFDM MISO model, the multiple antenna preamble JSPS sequence is developed in Section 3, while the pilot design part of the JSPS is described in Section 4. A robust performance CFO estimation algorithm, with low complexity, is detailed in Section 5, with performance results presented in Section 6. Section 7 concludes the paper.

2. SYSTEM MODEL

Figure 1 illustrates the basic structure of our OFDM system. We encode the data symbols using rate-1 space-frequency block-codes (SFBC), which can be transmitted in one OFDM block duration as contrasted to a STBC OFDM system, which requires two OFDM blocks. We denote N as the total number of OFDM subcarriers. The baseband time-domain transmission signal, $\boldsymbol{x}$, is structured such that preamble, $\boldsymbol{x}_p$, precedes information symbols, $\boldsymbol{x}_d$, and thus

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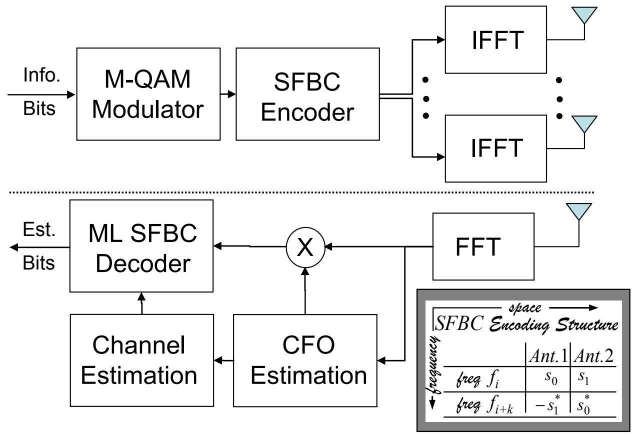

Fig. 1. Transmitter / receiver structure for a SFBC-OFDM system.

$$\boldsymbol{x} = [\boldsymbol{x}_p \;\; \boldsymbol{x}_d]^{\mathrm{T}}. \tag{1}$$

The preamble part, $\boldsymbol{x}_p$, is not a SFBC, but is designed specifically to enable course / fine synchronization and channel estimation with low complexity at the receiver. $\boldsymbol{x}_p$ is the composite baseband transmitted preamble signal from both antennas and is denoted as

$$\boldsymbol{x}_p = \sum_{i=1}^{A} \mathbf{x}_p^i \ , \tag{2}$$

where $\mathbf{x}_p^i$ is the JSPS from the ith transmit antenna and A is the number of transmit antennas. The data part, $\boldsymbol{x}_d$, is encoded according to a SFBC. For example if A = 2, the SFBC encoding of two symbols s_0 and s_1 is performed across two OFDM subcarriers f_i and f_{i+k} using the Alamouti code [2] as denoted in the lower right-hand corner of Figure 1, where $(\cdot)^*$ is the complex conjugate operator. This process is continued until all $|K_{si}|$ of the N OFDM subcarriers are assigned, where K_{si} denotes the data subcarrier indexes and $|K_{si}|$ is the cardinality of K_{si}. Since the fine CFO and channel can be estimated using the preamble part, $\boldsymbol{x}_p$, and if the channel is approximately constant over the packet burst, then $|K_{si}| = N - |K_n|$, where K_n denotes the null (unused) subcarrier indexes and $|K_n|$ is the number of null subcarriers.

The composite transmitted data signal, $\boldsymbol{x}_d$, can be written as

$$\boldsymbol{x}_d = \sum_{i=1}^{A} \mathbf{x}_d^i \ , \tag{3}$$

where $\mathbf{x}_d^i$ is the SFBC vector data sequence from the ith transmit antenna. If $\tilde{\boldsymbol{x}}$ is the same as $\boldsymbol{x}$, but with the cyclic extension removed, then prior to the cyclic extension, the baseband signal for antenna i is

$$\tilde{\mathrm{x}}^i[n] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_k e^{j2\pi nk/N} \ , \tag{4}$$

where $n \in \{0, 1, \ldots, N\text{-}1\}$, $\{X_k\}_{k\in K_{si}}$, $\{X_k\}_{k\in K_{spi}}$, and $\{X_k\}_{k\in K_n}$ are the non-zero SFBC encoded symbols during the data part, the non-zero part of the JSPS during the preamble part and a zero vector for the null subcarriers at indexes, K_n, for both the data and preamble parts.

The received baseband signal after multipath and CFO is

$$y[n] = (x[n] * h[n])e^{-j2\pi\varepsilon n/N} + w[n] \ , \tag{5}$$

where $h[n]$ is a length-L vector of the channel impulse response, with Rayleigh distributed amplitude and each channel tap is i.i.d. complex Gaussian with zero mean and variance such that $E[|h|^2]$ = 1. $\varepsilon = \Delta f \cdot T_s$ is the normalized CFO with respect to the subcarrier spacing $1/T_s$, where $T_s = TN$, T is the sample period, and $w[n]$ is zero-mean additive white Gaussian noise with variance σ_w^2 and "*" is the convolution operator. We note that $x[n]$ is the same as $\boldsymbol{x}$ in (1), with the index n representing the nth element of vector $\boldsymbol{x}$. In this paper, we assume coarse frame and frequency synchronization are achieved via acquisition of the JSPS [14]. Next we present algorithms to find the fine CFO and channel frequency response. Both CFO and channel estimators use the preamble JSPS, but because of the orthogonality of the transmitter pilot parts of the JSPS, the estimators are completely decoupled processes and are thus of low complexity.

3. MULTI-ANTENNA PREAMBLE JSPS DESIGN

Design of JSPS is described in [14] for SISO OFDM systems, where synchronization and pilot information are superimposed with the OFDM data. The pilot part of the JSPS in [14] uses a parametric pilot optimization process to improve symbol estimation from the least-squares (LS) channel estimates [15][16]. In this work, we extend the JSPS design introduced in [14] to multiple antennas. The pilot part of the JSPS is utilized for fine CFO estimation and also for channel estimation. In [15][16] it is shown that uniformly-spaced and constant-power (USCP) pilots are suboptimal. Accordingly, those authors propose a non-USCP solution for LS channel estimation that provides dramatic improvements over the USCP pilot schemes commonly used in OFDM. An alternative pilot design for OFDM systems with null subcarriers was proposed in [17], but was shown in [16] to perform worse than the USCP scheme for certain channel orders. Thus we choose to extend the scheme in [15][16] for the multiple antenna pilot portion of our JSPSs rather than the one proposed in [17].

In this paper, we use N = 256 and $|K_n|$ = 64, where there are 32 null subcarriers on each band edge. Given we have chosen the null subcarrier locations, the next step of the JSPS design is to determine the pilot positions and power loading for each antenna. The pilot positions and powers for each antenna, respectively, are denoted as $K_{p1}, K_{p2}, \ldots K_{pA}$ and $P_{p1}, P_{p2}, \ldots P_{pA}$, where $|K_{p1}| = |K_{p2}| = \ldots = |K_{pA}|$ and $\sum_{i=1}^{K_{p1}} P_{p1}[i] = \sum_{i=1}^{K_{p2}} P_{p2}[i] = \ldots = \sum_{i=1}^{K_{pA}} P_{pA}[i]$. Once these pilot parameters are determined for each antenna, the leftover positions are used for coarse synchronization information denoted as $K_{s1}, K_{s2}, \ldots K_{sA}$, all exhibiting a flat power profile in the frequency domain, with equal total power P_s on each $K_{s1}, K_{s2}, \ldots K_{sA}$ for each transmit antenna. We note that to provide frequency orthogonal pilots across transmitter antennas, if $K_{sp1}, K_{sp2}, \ldots, K_{spA}$ represents all possible subcarrier positions, excluding $K_{n1} = K_{n2} = \ldots = K_{nA}$, then the ith transmit antenna power,

$$P_{spi} = \begin{cases} P_{pi} \ for \ K_{pi} \\ P_s \ \ for \ K_{si} \\ 0 \ \ for \ K_{pj}, j \neq i \end{cases} \tag{6}$$

Once the pilot positions and power are determined, the method in [7] is used to fill in the remaining subcarriers for each antenna with synchronization information.

For the data part of the transmission, symbols are encoded using the SFBC method described in the first part of Section 2. If no pilots are required during data transmission, $K_{s1} = K_{s2} = \ldots = K_{sA} = K_{sp1} = K_{sp2} = \ldots = K_{spA}$ and $|K_{si}| = |K_{spi}|$. If pilots are required due to dynamic channel conditions and CFO drift, where $|K_p|$ pilots are inserted every OFDM symbol for each transmitter antenna, then $|K_{si}| = |K_{spi}| - A \cdot |K_{pi}|$. Next we describe the pilot parameter optimization for multiple antennas.

4. MULTI-ANTENNA PILOT OPTIMIZATION

In [15][16] a parametric pilot optimization process is proposed for SISO OFDM systems. The parameterization is accomplished using a segmented optimization technique by applying boundary constraints to an un-tractable non-linear discontinuous optimization. We extend this technique to multiple-antenna OFDM systems. We will show from simulation that this extension provides near optimal performance when compared to the perfect channel knowledge case with no CFO.

Candidate K_{pi} indexes for each antenna can be formulated as

$$K_{pi} = \{\mathrm{int}(f^{-1} \circ g_i(\tau)) \mid \tau \in \{0, 1, 2, \ldots, |K_{pi}| - 1\}, \tag{7}$$

until the minimizing set is found for each transmit antenna. The domain of f is restricted to $[1, N]$, where f is a one-to-one mapping such that $K_{pi} = f^{-1}(S)$ and S is a circularly shifted version of K_{pi}. Following the procedure in [16], a cubic function is used to parameterize the pilot subcarrier positions according to

$$g_i(\tau) = a_{3i}\tau^3 + a_{2i}\tau^2 + a_{1i}\tau + a_{0i}. \tag{8}$$

Given that the number of non-null subcarriers for the ith antenna is denoted as N_i and denoting the middle of the non-null subcarriers as ½, the constraint equations become

$$g_i\left(\frac{|K_{pi}|-1}{2}\right) = 1/2 \tag{9}$$

$$g_i(0) = -(N_i - 1)/2 + \delta_i \tag{10}$$

$$g_i(|K_{pi}|-1) = (N_i + 1)/2 - \delta_i \tag{11}$$

$$g_i'(0) > 0, \tag{12}$$

where δ_i represents the distance the edge pilots are from the signal band edges. Using the constraint equations in (9)-(12) and the fact that the edge pilot placement should be no further from the signal band edge than the average pilot spacing, three of the five variables can be eliminated and thus the optimization problem in (7) is greatly simplified. We note that once K_{pi} is found starting with antenna i, then K_{pi} are excluded from the set for the next antenna pilot positions and so on until K_{pi} are determined for all antennas. The composite pilot sequence from all antennas can be positioned in any non-null subcarriers and are placed symmetrically about the center of the signal band. Without loss of generality, the pilots are placed sequentially from left to right giving $g_i(\tau)$ a positive slope. If $\hat{K}_{pi}$ represents the ith antenna candidate set of pilot subcarrier positions and extending the single antenna optimization [16], the pilot powers are found by solving

$$\arg\min_{\mathbf{u}} \quad \|\mathbf{A}\mathbf{u}\|_\infty, \text{ subject to } \sum_{k=1}^{|K_{pi}|} \frac{1}{[\mathbf{u}]_k} = E_p, \tag{13}$$
$$\hat{K}_{pi} = K_{pi}, \ [\mathbf{u}]_k > 0 \forall k,$$

where E_p is the power allocated to the pilots, $\mathbf{A}$ is the element-wise magnitude square of the matrix $\mathbf{Q}_{si}\mathbf{Q}_{pi}^{+}$, where $\mathbf{Q}_{pi} = [\mathbf{Q}]_{K_{pi},\{1,2,\ldots,L\}}$, $\mathbf{Q}_{si} = [\mathbf{Q}]_{K_{si},\{1,2,\ldots,L\}}$, $\mathbf{Q}_{k,n} = N^{-1/2}\exp(-j2\pi(n-1)(k-1)/N), 1 \le k, n \le N$, and the LS channel estimator is $\hat{H}_i = \mathbf{Q}_{si}\mathbf{Q}_{pi}^{+} Y_{ki}^{p} [X_{ki}^{p}]^{-1}$ (CFO = 0). The LS estimator data subcarrier MSE of the channel estimate is approximated by $z_i \approx \mathrm{diag}\left\{\frac{\sigma_w^2}{E_p}\mathbf{Q}_{si}\mathbf{Q}_{pi}^{+}\mathbf{D}_{|\mathbf{x}_p^i|^{-2}}\mathbf{Q}_{pi}^{H^+}\mathbf{Q}_{si}^{H}\right\}$, where $\mathbf{x}_p^i$ are the ith antenna modulated pilot values, $\mathbf{D}_\mathbf{u}$ is a diagonal matrix with diagonal elements from vector $\mathbf{u}$, $\mathbf{Q}^+$ and $\mathbf{Q}^H$ denote the pseudo inverse and Hermitian transpose of $\mathbf{Q}$, respectively. The transmitter pilot optimization does not require channel knowledge. At the receiver, CFO estimation is performed prior to channel estimation.

5. CFO ESTIMATION

The CFO is estimated prior to channel estimation. The preamble part for each antenna is comprised of one JSPS followed by another JSPS (i.e., [JSPS JSPS]) [6]. The CFO is estimated by measuring the phase change from JSPS-to-JSPS along each pilot subcarrier k. We can write the received version of preamble JSPS part of (1) as $\boldsymbol{y}_p$, which can be separated into two parts such that $\boldsymbol{y}_p = [\boldsymbol{y}_{p1}\ \boldsymbol{y}_{p2}]^T$. The frequency domain pilot part of the $\boldsymbol{y}_{p1}$ and $\boldsymbol{y}_{p2}$ signals for the ith transmit antenna (ignoring receiver noise, assuming L < cyclic prefix length and the CFO ICI is part of the channel frequency response, H_i) can be approximated as $Y_{ki1}^{p} = X_{ki1}^{p} H_{ki1}^{p} e^{-j2\pi\varepsilon_1'/N}$ and $Y_{ki2}^{p} = X_{ki2}^{p} H_{ki2}^{p} e^{-j2\pi\varepsilon_2'/N}$, respectively. If the phases ϕ_{ki1} and ϕ_{ki2}, are computed as $\angle Y_{ki1}^{p}$ and $\angle Y_{ki2}^{p}$, respectively, then the CFO is estimated according to

$$CFO_i = \frac{1}{2\pi T_s} \sum_{k=0}^{|K_{pi}-1|} (\phi_{ki1} - \phi_{ki2}) \cdot v_{ki}, \tag{14}$$

where v_{ki} is a weighting vector of size $|K_{pi}|$ based on the received pilot power. If $P_{ki1} = Y_{ki1}^{p} \cdot (Y_{ki1}^{p})^{*}$, then v_k can be written as

$$v_{ki} = \frac{(P_{ki1} + P_{ki2})/2}{\sum_{k=0}^{|K_{pi}-1|} (P_{ki1} + P_{ki2})/2}. \tag{15}$$

This CFO estimator has low complexity and is decoupled between transmitters. This *decoupling* between each antenna means that we can exploit both estimates for reduced estimation variance.

6. PERFORMANCE RESULTS

Unless otherwise specified, the SNR is defined as E_s/N_0, where E_s is the energy in a single constellation symbol s and $N_0 = 2\sigma_w^2$ is the noise spectral density. The left plot in Figure 2 shows the MSE profile for antenna 1 of a 2-transmitter system using the optimization procedure outlined in Section 4 for two values of a_{3i}. The USCP pilot design occurs when a_{3i} = 0.0 and a_{3i} = -0.057 provides a flat maximum MSE profile, which we denote as the "*optimized*" pilot design in subsequent plots. The MSE profile for antenna 2 is nearly identical to that of antenna 1, so it is not shown. The right plot in Figure 2 compares the channel estimate maximum MSE in an L = 16 tap multipath channel as described in Section 2. The optimized scheme performs greater than 15 dB better than the USCP scheme.

Simulated MSE performance of CFO_i in (14) for $|K_{pi}|$=L=16 tap Rayleigh channel and CFO, $\varepsilon_1 = \varepsilon_2 = 0.25$, is shown in Figure 3 for both pilot schemes. SNR in Figure 3 is defined as the JSPS, $\boldsymbol{x}_p$,

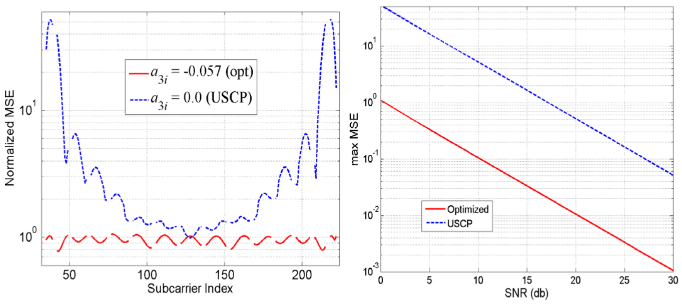

Fig. 2. Normalized MSE profile (left), LS $\hat{H}_i$ maximum MSE (right).

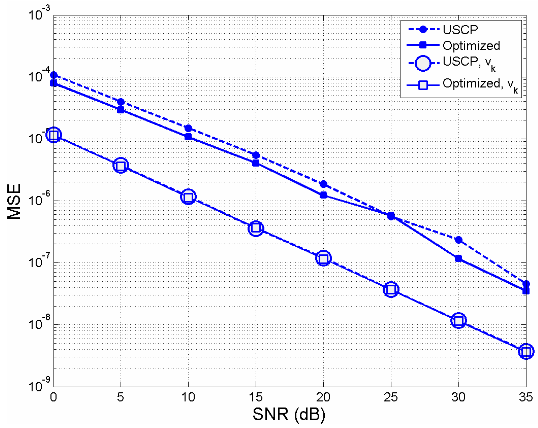

Fig. 3. CFO_i MSE for USCP and optimized pilots.

signal-to-noise power, $\sigma_{x_p}^2 / \sigma_w^2$. The MSE is measured according to $E[|CFO_i - \Delta f|^2]$. For CFO estimation, the optimized pilot design works slightly better than USCP pilots without weighting. Once weighting is utilized according to (15), a factor of 10 reduction in MSE performance is found. In the weighting case both pilot schemes perform equally well.

Demodulated BER performance is measured for A=2 transmit antennas and one receiver antenna SFBC-OFDM system. BER performance is compared in Figure 4 for channel estimation with and without CFO correction. Without CFO correction, the induced CFO is, $\varepsilon_1 = \varepsilon_2 = 0.05$ (0.2 kHz), while for the case with CFO correction, $\varepsilon_1 = \varepsilon_2 = 0.25$ (1 kHz). A maximum likelihood decoder is utilized, as defined in [2]. When the CFO is not corrected, both the USCP and optimized pilot schemes degrade severely. When the CFO is estimated and corrected (denoted as the "w/C" curves) prior to channel estimation, the optimized design performs nearly 4 dB better than USCP and within ~0.5 dB of the perfect channel state information (PCSI) curve.

7. CONCLUSIONS

A new parametric pilot optimization scheme was developed for multi-antenna OFDM transmission systems. The optimized scheme provides >15 dB improvement in maximum MSE LS channel estimation performance compared to uniformly-spaced constant-power (USCP) pilots. The optimized pilot scheme MSE performance was shown to be equal or better than USCP pilots for carrier frequency offset (CFO) estimation. For demodulation BER performance of SFBC-OFDM, the optimized pilot scheme performs at least 3 dB better than USCP.

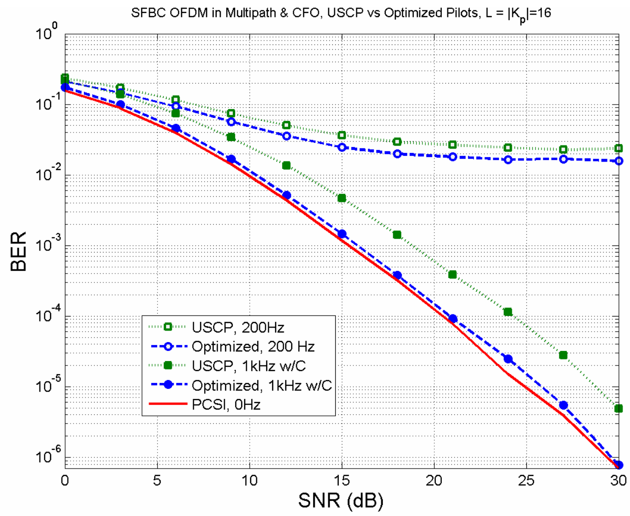

Fig. 4. BER for SFBC-OFDM, USCP and optimized pilots in multipath and CFO (with and without CFO correction).

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