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A MIMO SYSTEM WITH BACKWARD COMPATIBILITY FOR OFDM BASED WLANS

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ABSTRACT

Orthogonal frequency-division multiplexing (OFDM) has been selected as the basis for the new IEEE 802.11a standard for high-speed wireless local area networks (WLANs). We consider doubling the data rate of the IEEE 802.11a system by using a multi-input multi-output (MIMO) system with two transmit and two receive antennas. We propose a preamble design for this MIMO system that is backward compatible with its single-input single-output (SISO) counterpart as specified by the IEEE 802.11a standard. Based on this preamble design, we devise a sequential method for the estimation of the carrier frequency offset (CFO), symbol timing, and channel response. We also provide a simple soft-detector to obtain the soft-information for the Viterbi decoder. Both the sequential parameter estimation method and the soft-detector are ideally suited for real-time implementations. The effectiveness of our methods is demonstrated via numerical examples.

1. INTRODUCTION

Orthogonal frequency-division multiplexing (OFDM) has been selected as the basis for the IEEE 802.11a 5 GHz band high-speed wireless local area network (WLAN) standard [1]. This standard supports a data rate up to 54 Mbps by using a single-input single-output (SISO) system.

Transmission data rates higher than 54 Mbps are of particular importance for future WLANs. Deploying multiple antennas at both the transmitter and the receiver is a promising way to achieve a high transmission data rate for multipath-rich wireless channels without increasing the total transmission power or bandwidth [2]. The corresponding system is referred to as a multi-input multi-output (MIMO) wireless communication system.

Among the various popular MIMO wireless communication schemes, the BLAST (Bell-labs' LAyered Space-Time) approaches are particularly attractive (see, e.g., [3] and the references therein). BLAST attempts to achieve the potentially large channel capacity offered by the MIMO system [2]. In BLAST systems, the data stream is demultiplexed into independent sub-streams that are referred to as layers. These layers are transmitted simultaneously, i.e., one layer per transmit antenna. At the receiver, the multiple layers can be detected, for example, through successive detection via an interference cancellation and nulling algorithm (ICNA) [3].

Our focus herein is on doubling the data rate of the SISO system as specified by the IEEE 802.11a standard by using two transmit and two receive antennas (referred to as the MIMO system in the sequel) based on the BLAST scheme. We propose a preamble design for this MIMO system that is backward compatible with its SISO counterpart as specified by the IEEE 802.11a standard. That is, a SISO receiver can perform CFO, symbol timing, and channel response estimation based on the proposed preamble design and can detect up to the SIGNAL field. The SISO receiver is then informed, by using, e.g., the reserved bit in the SIGNAL field, that a transmission is a SISO or not. Our preamble design can be used with two transmit and any number of receive antennas. However, we mainly focus on the two receive antenna case herein.

Based on our preamble design, we propose a sequential method, ideally suited for real-time implementations, to estimate the CFO, symbol timing, and MIMO channel response. The convolutional code specified in the IEEE 802.11a standard will also be used in our MIMO system for channel coding. As a result, soft-information from the MIMO detector is needed by the Viterbi Algorithm (VA) to improve the decoding performance. A List Spere Decoder (LSD) algorithm [4] was recently proposed to deliver soft-information. However, LSD is too complicated to be implemented in real-time. We present herein a simple MIMO soft-detector, ideally suited for real-time implementations, based on the unstructured least-square (LS) fitting approach. This soft-detector is computationally much more efficient than LSD; yet the efficiency is achieved at a cost of a small performance degradation.

2. SYSTEM DESIGN

Our MIMO system closely resembles its SISO counterpart as specified by the IEEE 802.11a standard. We first give a brief overview of the IEEE 802.11a based SISO system before we proceed to describe our MIMO system.

2.1. IEEE 802.11a Standard

The OFDM based WLAN system, as specified by the IEEE 802.11a standard, uses packet-based transmission. Fig. 1 shows the packet structure specified by the standard. The nominal bandwidth of the OFDM signal is 20 MHz and the I/Q sampling interval t_S is 50 ns. The OFDM packet preamble consists of 10 identical short OFDM training symbols $\mathrm{t}_i, i = 1, 2, \cdots, 10$, each of which contains $N_C = 16$ samples, and 2 identical long OFDM training symbols $\mathrm{T}_i, i = 1, 2$, each of which contains $N_S = 64$ samples. Between the short and long OFDM training symbols there is a long guard interval (GI2) consisting of $2N_C = 32$ data samples. GI2 is the cyclic prefix (CP) for the long OFDM training symbol T_1, i.e., it is the exact replica of the last $2N_C$ samples of T_1.

The information carrying data are encoded in the OFDM DATA field. The binary source data sequence is first scrambled and then

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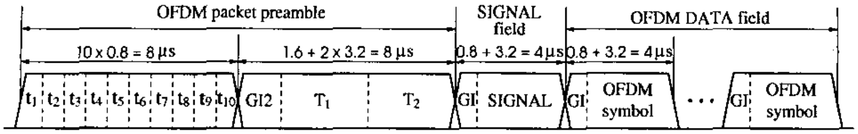

Fig. 1. Packet structure of the IEEE 802.11a standard.

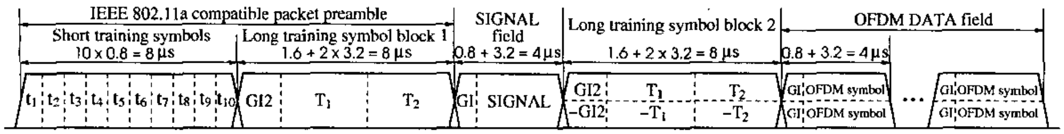

Fig. 2. Proposed MIMO preamble (and SIGNAL field) structure.

Convolutionally enCoded (CC). The CC encoded output is then punctured according to the data rate requirement and is segmented into blocks of length N_{CBPS} (number of coded bits per OFDM symbol), each of which corresponds to an data OFDM symbol. The binary data in each block is first interleaved among the subcarriers (referred to as the frequency-domain (FD) interleaving in the sequel) and then mapped (in groups of $\log_2 A$ bits) into A-QAM symbols, which are used to modulate the different data carrying subcarriers. Each data OFDM symbol in the OFDM DATA field employs $N_S = 64$ subcarriers, 48 of which are used for data symbols and 4 for pilot symbols. There are also 12 null subcarriers. The data OFDM symbols, each of which consists of $N_S = 64$ samples, are obtained via taking the IFFT of the data symbols, pilot symbols, and nulls on these N_S subcarriers. To eliminate the inter-symbol interference (ISI), each data OFDM symbol is preceded by a CP or guard interval (GI), which contains the last N_C samples of the data OFDM symbol.

The SIGNAL field contains the information including the transmission data rate and data length of the packet. The information is encoded in 16 binary bits. There is also a reserved bit (which can be used to distinguish the MIMO from SISO transmissions) and a parity check bit. These 18 bits, padded with 6 zeros, are then CC encoded and mapped via BPSK onto the data carrying subcarriers.

2.2. SISO Data Model

To establish the data model, consider first the generation of an data OFDM symbol in the OFDM DATA field. Let $\mathbf{x}_{\text{SISO}} = [x_1^1 \; x_2^1 \; \cdots \; x_{N_S}^1]^T$ be a vector of N_S data symbols, where $(\cdot)^T$ denotes the transpose, and $x_{n_S}^1$, $n_S = 1, 2, \cdots, N_S$, is the symbol modulating the n_Sth subcarrier, which is equal to 0 for null subcarriers, 1 or -1 for pilot subcarriers, and in $\mathcal{C}$ for data carrying subcarriers. Here $\mathcal{C}$ is a finite constellation, such as BPSK, QPSK, 16-QAM, or 64-QAM. Let $\mathbf{W}_{N_S} \in \mathbb{C}^{N_S \times N_S}$ be the FFT matrix. Then the data OFDM symbol $\mathbf{s}$ corresponding to $\mathbf{x}_{\text{SISO}}$ is obtained by taking the IFFT of $\mathbf{x}_{\text{SISO}}$. That is, $\mathbf{s} = \mathbf{W}_{N_S}^H \mathbf{x}_{\text{SISO}} / N_S$, where $(\cdot)^H$ denotes the conjugate transpose.

Let $\mathbf{h}^{(t)} = [h_0^{(t)} \; h_1^{(t)} \; \cdots \; h_{L_f-1}^{(t)}]^T$ be the FIR response of the channel. Here $L_F = \lceil 10 t_r / t_S \rceil + 1$ with t_r being the root mean square (rms) delay spreading time and $\lceil x \rceil$ denoting the smallest integer not less than x. We assume, for $l_F = 0, \cdots, L_F - 1$, $h_{l_F}^{(t)} \sim \mathcal{N}(0, (1 - e^{-t_S/t_r}) e^{-l_F t_S / t_r})$. This channel model is referred to as the exponential model.

By discarding the first N_C samples at the receiver (assuming a correct symbol timing), the noise-free and CFO free received signal vector $\mathbf{z}_{\text{SISO}}^{\text{ne}} \in \mathbb{C}^{N_S \times 1}$, due to sampling the received signal, is the circular convolution of $\mathbf{h}^{(t)}$ and $\mathbf{s}$. Hence the FFT output of the received data vector $\mathbf{z}_{\text{SISO}} = \mathbf{z}_{\text{SISO}}^{\text{ne}} + \mathbf{e}_{\text{SISO}}$, where $\mathbf{e}_{\text{SISO}}$ is the additive zero-mean white circularly symmetric complex Gaussian noise, can be written as [5]

$$\mathbf{y}_{\text{SISO}} = \text{diag}\{\mathbf{h}\} \mathbf{x}_{\text{SISO}} + \mathbf{W}_{N_S} \mathbf{e}_{\text{SISO}} \in \mathbb{C}^{N_S \times 1}. \qquad (1)$$

2.3. MIMO Preamble Design

For the IEEE 802.11a based SISO system, the short training symbols can be used to detect the arrival of the packet, set up AGC, compute a coarse CFO estimate, and obtain a coarse symbol timing, whereas the long training symbols can be used to calculate a fine CFO estimate, refine the coarse symbol timing, and estimate the SISO channel.

The MIMO system considered herein has two transmit and two receive antennas. Two packets are transmitted simultaneously from the two transmit antennas. We design two preambles, one for each transmit antenna. We assume that the receiver antenna outputs suffer from the same CFO and has the same symbol timing. To be backward compatible with the SISO system, we use the same short training symbols as in the SISO preamble for both of the MIMO transmit antennas, as shown in Fig. 2.

Channel estimation for MIMO systems has attracted much research interest lately. Orthogonal training sequences tend to give the best performance (see, e.g., [6] and the references therein). We also adopt this idea of orthogonal training sequences in our preamble design. In the interest of backward compatibility, we use the same T_1 and T_2 as for the SISO system for both of the two transmit antennas before the SIGNAL field, as shown in Fig. 2. After the SIGNAL field, we use T_1 and T_2 for one transmit antenna and $-\mathrm{T}_1$ and $-\mathrm{T}_2$ for the other. This way, when the simultaneously transmitted packets are received by a single SISO receiver, the SISO receiver can successfully detect up to the SIGNAL field, which is designed to be the same for both transmit antennas. The reserved bit in the SIGNAL field can tell the SISO receiver to stop its operation whenever a MIMO transmission follows or otherwise to continue its operation. The long training symbols before and after the SIGNAL field are used in the MIMO receivers for channel estimation.

2.4. MIMO Data Model

To stay as close to the IEEE 802.11a standard as possible, we use in our MIMO system the same scrambler, convolutional encoder, puncturer, FD interleaver, symbol mapper, pilot sequence, and CP as specified in the standard. To improve diversity, we add a simple spatial interleaver to scatter every two consecutive bits across the two transmit antennas.

Consider the n_Sth subcarrier (for notational convenience, we drop the notational dependence on n_S below). Consider the case of N receive antennas. Let $\mathbf{H} = [h_{n,m}] \in \mathbb{C}^{N\times 2}$ denote the MIMO channel matrix for the n_Sth subcarrier, where $h_{n,m}$ is the channel gain from the mth transmit antenna to the nth receive antenna for the n_Sth subcarrier. Let $\mathbf{y}$ denote a received data vector for the n_Sth subcarrier. Then it can be written as

$$\mathbf{y} = \mathbf{Hx} + \mathbf{e} \in \mathbb{C}^{N\times 1}, \tag{2}$$

where $\mathbf{x} = [x_1 \;\; x_2]^T$ is the 2×1 QAM symbol vector sent on the n_Sth subcarrier and $\mathbf{e} \sim \mathcal{N}(\mathbf{0}, \sigma^2 \mathbf{I}_N)$ is the additive white circularly symmetric complex Gaussian noise with variance σ^2. In Section 4, we will provide a soft-detector based on this model.

3. SEQUENTIAL CHANNEL ESTIMATION

In this section, we present our sequential CFO, symbol timing, and MIMO channel estimation approach based on our preamble design. The estimates are obtained in the order presented below.

3.1. Coarse CFO and Symbol Timing Estimation

Let $z_n(l) = z_n^{\text{ne}}(l) + e_n(l)$, $n = 1, \cdots, N$, denote the lth time sample of the signal received from the nth receive antenna, starting from the moment that the receiver AGC has become stationary (the receiver AGC is assumed to become stationary at least before receiving the last two short OFDM training symbols and remain stationary while receiving the remainder of the packet). In the presence of CFO, ϵ, we have [7]:

$$z_n^{\text{ne}}(l + N_C) = z_n^{\text{ne}}(l)e^{j2N_C\pi\epsilon}, \quad n = 1, \cdots, N. \tag{3}$$

For each receive antenna output, consider the correlation between two consecutive noise-free received data blocks, each of which is of length N_C. Then the sum of the correlations for all receive antennas can be written as

$$\sum_{n=1}^{N}\sum_{l=k}^{k+N_C-1} z_n^{\text{ne}}(l)(z_n^{\text{ne}}(l+N_C))^* = Pe^{-j2N_C\pi\epsilon}, \tag{4}$$

where $P \triangleq \sum_{n=1}^{N}\sum_{l=0}^{N_C-1}|z_n^{\text{ne}}(l)|^2$, $(\cdot)^*$ denotes the complex conjugate, and k is any non-negative integer such that $z_n^{\text{ne}}(k + 2N_C - 1)$ is a sample of the nth receive antenna output due to the input (transmit antenna output) being a sample of the short OFDM training symbols. Let

$$P_S = \sum_{n=1}^{N}\sum_{l=0}^{N_C-1} z_n(l)z_n^*(l+N_C) = Pe^{-j2N_C\pi\epsilon} + e_P, \tag{5}$$

where e_P is due to the presence of the noise. We calculate the coarse CFO as [8]

$$\hat{\epsilon}_C = -\frac{1}{2N_C\pi}\angle P_S, \tag{6}$$

where $\angle x$ denotes taking the argument of x.

We next correct the CFO using $\hat{\epsilon}_C$ to get the data samples $z_n^{(C)}(l)$ as $z_n^{(C)}(l) = z_n(l)e^{-j2l\pi\hat{\epsilon}_C}$, $n = 1, 2, \cdots, N$. In the sequel, we only consider the CFO corrected data given above. For notational convenience, we drop the superscript of $z_n^{(C)}(l)$.

Now we can use a correlation method similar to the one proposed in [7] to estimate the coarse symbol timing. Here the symbol timing is referred to as the starting time sample due to the input being the long OFDM training symbol T_1 (before the SIGNAL field). Once the starting time sample due to T_1 is determined, we can determine the starting time sample for every OFDM symbol thereafter. (According to the specification of the IEEE 802.11a standard and the sampling rate of 20 MHz, the true symbol timing T_0 is 193.)

From (4), we note that the correlation (after the CFO correction) is approximately the real-valued scalar P (plus a complex-valued noise). Hence we propose to use the following real-valued correlation sequence for coarse symbol timing determination. We calculate the correlation sequence in an iterative form similar to the complex-valued approach in [7] as follows:

$$P_R(k+1) = P_R(k) + \text{Re}\sum_{n=1}^{N}\left[z_n(k+N_C)z_n^*(k+2N_C) - z_n(k)z_n^*(k+N_C)\right], \tag{7}$$

where $\text{Re}(\cdot)$ denotes the real part of a complex entity. We start the iteration by using $P_R(0) = \text{Re}(P_S)$.

When some of the data samples of the sliding data blocks are taken from the received data due to the input being GI2 or the long training symbols following the short OFDM training symbol, $P_R(k)$ will drop since (3) no longer holds. This property is used to obtain the coarse symbol timing. Let T_P denote the first time sample when $P_R(k)$ drops to less than half of its peak value. Then the coarse symbol timing can be written as

$$T_C = T_P + \frac{3}{2}N_C + N_C. \tag{8}$$

Note that the second term at the right hand side of (8) is due to the fact that $P_R(k)$ will drop to approximately one half of its maximum value when the data samples of the second half of the second of the two sliding blocks are due to GI2 in the preamble; the third term is due to one half of the length of GI2, since our goal of coarse timing determination is to place the coarse timing estimate between the true timing $T_0 = 193$ and $T_0 - N_C = 177$ to make accurate fine CFO estimation possible.

3.2. Fine CFO and Symbol Timing Estimation

The fine CFO estimate can be computed as

$$\hat{\epsilon}_F = -\frac{1}{2N_S\pi}\angle\sum_{n=1}^{N}\sum_{l=0}^{N_S-1} z_n(l+T_C)z_n^*(l+T_C+N_S). \tag{9}$$

We can use $\hat{\epsilon}_F$ in the same way as $\hat{\epsilon}_C$ to correct the CFO. We assume that for the data we use below $\hat{\epsilon}_F$ has been already corrected. (The fine CFO estimation is accurate enough for the following fine symbol timing and MIMO channel response estimation. However, it can never be perfect due to the noise. Hence before data bits detection, we need to use the pilot symbols to track the CFO residual phase for each data OFDM symbol. A maximum-likelihood (ML) CFO residual tracking scheme is given in Appendix A.)

Let $\mathbf{y}_n$ denote the N_S-point FFT of the data block from the nth receive antenna, starting from T_C, and let $\mathbf{h}_{n,m}^{(t)}$ be the FIR channel in the time-domain between the mth transmit antenna and the nth receive antenna, $m = 1, 2$, $n = 1, 2, \cdots, N$. Then, by neglecting the existence of the residual CFO, $\mathbf{y}_n$ can be written as $\mathbf{y}_n = \mathbf{X}_B \mathbf{W}_{N_S} \sum_{m=1}^{2} \mathbf{h}_{n,m}^{(t)} + \mathbf{W}_{N_S} \mathbf{e}_n$, where $\mathbf{X}_B$ is a diagonal matrix with the 52 known BPSK symbols and 12 zeros, which form the T_1 in Fig. 2, on the diagonal. We get an estimate of $\mathbf{h}_n^{(t)} = \sum_{m=1}^{2} \mathbf{h}_{n,m}^{(t)}$ as $\hat{\mathbf{h}}_n^{(t)} = \mathbf{W}_{N_S}^H \mathbf{X}_B \mathbf{y}_n / N_S$. Let T_I denote the index of the first element of $\sum_{n=1}^{N} |\hat{\mathbf{h}}_n^{(t)}|$ that is above 1/3 of the maximum value of the elements of $\sum_{n=1}^{N} |\hat{\mathbf{h}}_n^{(t)}|$. Then the fine symbol timing T_F is obtained as

$$T_F = T_C + T_I - 3. \quad (10)$$

The last term above is chosen to be 3 to ensure that $T_F > T_0$ with negligible probability.

3.3. MIMO Channel Estimation

After we obtained T_F, we can now estimate the MIMO channel response. Let $\mathbf{y}_{n,1}$ denote the N_S-point FFT of the average of the two consecutive blocks, each of which is of length N_S, associated with the two long training symbols before the SIGNAL field, from the nth receive antenna. Let $\mathbf{y}_{n,2}$ denote the counterpart of $\mathbf{y}_{n,1}$ after the SIGNAL field. Then, for the n_Sth subcarrier, we have

$$y_{n,1} \approx x_B \left(h_{n,1} + h_{n,2}\right), \quad y_{n,2} \approx x_B \left(h_{n,1} - h_{n,2}\right), \quad (11)$$

where x_B denotes the n_Sth diagonal element of $\mathbf{X}_B$ and $y_{n,i}$ denotes the n_Sth element of $\mathbf{y}_{n,i}$, $i = 1, 2$. Solving (11) yields

$$\hat{h}_{n,1} = x_B(y_{n,1} + y_{n,2})/2, \quad (12)$$

$$\hat{h}_{n,2} = x_B(y_{n,1} - y_{n,2})/2. \quad (13)$$

4. A SIMPLE MIMO SOFT-DETECTOR

With the CFO, symbol timing, and MIMO channel determined and accounted for, we can proceed to detect the data bits contained in each BLAST layer and subcarrier of the data OFDM symbols in the OFDM DATA field. In the sequel we present a very simple soft-detector for the MIMO system. Note that this soft-detector can be used in a general setting of the BLAST system and hence we present it in a general framework based on the data model of (2), where $\mathbf{H}$ is assumed to be $N \times M$ and $\mathbf{x}$ to be $M \times 1$. (We use $\hat{\mathbf{H}}$ to replace $\mathbf{H}$ in some of our simulations.)

Consider first the ML hard detector of the BLAST system. For the data model of (2), the ML hard detector is given by:

$$\hat{\mathbf{x}} = \arg \min_{\mathbf{x} \in \mathcal{C}^{M \times 1}} \|\mathbf{y} - \mathbf{H}\mathbf{x}\|^2, \quad (14)$$

where $\|\cdot\|^2$ denotes the Euclidean norm. Let $\mathbf{H}^\dagger = (\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H$. Then we can obtain an unstructured LS estimate $\hat{\mathbf{x}}_{\text{us}}$ of $\mathbf{x}$ as:

$$\hat{\mathbf{x}}_{\text{us}} = \mathbf{H}^\dagger \mathbf{y} = \mathbf{x} + \mathbf{H}^\dagger \mathbf{e} \triangleq \mathbf{x} + \mathbf{c}. \quad (15)$$

Note that $\hat{\mathbf{x}}_{\text{us}}$ is the soft-decision statistic that we are interested in. We refer to this simple scheme of obtaining a soft-decision statistic as the MIMO soft-detection scheme. We remark that a necessary condition for $\mathbf{H}^H\mathbf{H}$ to be nonsingular is $N \geq M$ and $\mathbf{c}$ is still Gaussian with zero-mean and covariance matrix

$$\mathrm{E}\left[\mathbf{c}\mathbf{c}^H\right] = \sigma^2 \mathbf{H}^\dagger (\mathbf{H}^\dagger)^H = \sigma^2 (\mathbf{H}^H\mathbf{H})^{-1}. \quad (16)$$

Due to the use of the interleaver and deinterleaver, the data bits contained in $\mathbf{x}$ are independent of each other. By ignoring the dependence among the elements of $\mathbf{c}$, we can consider only the marginal probability density function (pdf) for the elements $\hat{x}_{\text{us}}(m)$, $m = 1, 2, \cdots, M$, of $\hat{\mathbf{x}}_{\text{us}}$. Let $\mathbf{H}^\dagger = [\breve{\mathbf{h}}_1 \ \cdots \ \breve{\mathbf{h}}_M]^T \in \mathbb{C}^{M \times N}$. Then the mth element of $\mathbf{c}$, $m = 1, 2, \cdots, M$, can be written as $c_m = \breve{\mathbf{h}}_m^T \mathbf{e}$. Obviously, c_m is still Gaussian with zero-mean and variance

$$\sigma_m^2 = \mathrm{E}\left[|c_m|^2\right] = \|\breve{\mathbf{h}}_m\|^2 \sigma^2. \quad (17)$$

(The estimate of the above noise variance σ^2 can be easily obtained via the difference of the two consecutive blocks of the nth receive antenna, from which we got $\mathbf{y}_{n,1}$ [cf. (11)].) σ_m^2 and $\hat{x}_{\text{us}}^{(m)}$ provide the soft-information for the mth, $m = 1, 2, \cdots, M$, symbol in $\hat{\mathbf{x}}_{\text{us}}$, needed by the VA. Note that the noises corresponding to different layers have different variances which means that the symbols corresponding to different layers have different quality. This unbalanced layer quality is the reason why we want to use a spatial interleaver.

5. NUMERICAL EXAMPLES

In this section, we provide numerical examples to demonstrate the effectiveness and performance of our sequential parameter estimation method as well as the simple MIMO soft-detector.

We consider doubling the maximum 54 Mbps data rate by using two transmit and two receive antennas, i.e., $M = N = 2$. In our simulations, each of the $MN = 4$ time-domain MIMO channels is generated according to the exponential model; the 4 channels are independent of each other.

Due to the fact that 52 out of 64 subcarriers are used in the OFDM WLAN system, the SNR for the SISO system used herein is defined as $52/(64\sigma^2)$ for the constellations whose average energies are normalized to 1. Whereas for the MIMO system, the SNR is defined as $52/(128\sigma^2)$ (i.e., we use the same total transmission power for the MIMO system as for its SISO counterpart).

We first provide a simulation example for symbol timing estimation. Two curves in Fig. 3 show the 10^4 Monte-Carlo simulation results of the coarse symbol timing estimates for the exponential channels with $t_r = 25$ and 50 ns, respectively, when SNR = 10 dB. The other two curves in Fig. 3 show the 10^4 Monte-Carlo simulation results of the fine symbol timing estimates based on their corresponding coarse timing estimates. Note that our simple fine symbol timing approach gives highly accurate timing estimates.

We then provide a simulation example to show the effectiveness of the MIMO channel estimator and the PER (packet error rate) performance of the MIMO soft-detector. (One packet consists of 1000 bytes.) In Fig. 4, we show the 10^4 Monte-Carlo simulation results of the PER performance of the MIMO soft-detector as a function of the SNR for the MIMO system, with t_r being 50 ns for the exponential channels, when the data rate is 108 Mbps. We consider two cases: the case of perfect knowledge for CFO, symbol timing, and MIMO channel and the case of estimated channel parameters. (For the second case, besides the parameters obtained with our sequential approach, we correct the CFO residual phase

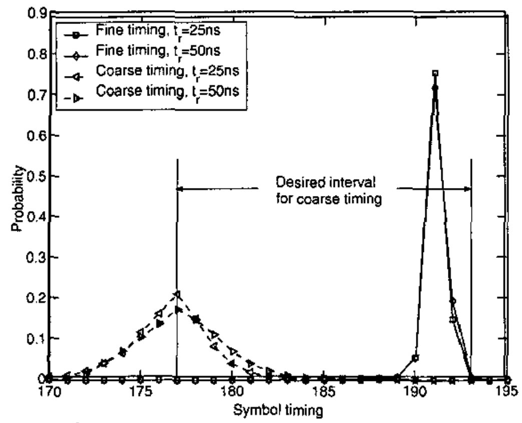

Fig. 3. Coarse and fine symbol timing estimates.

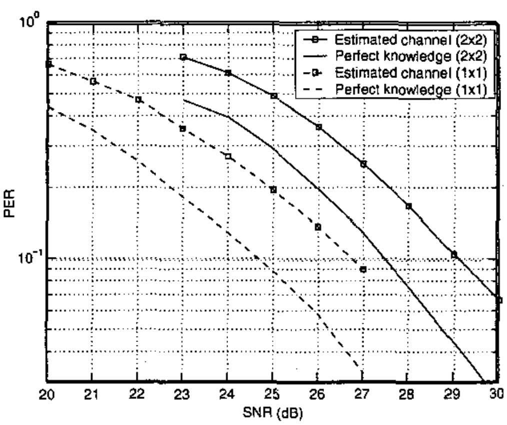

Fig. 4. PER versus SNR at the 108 Mbps data rate for the exponential channels with $t_r = 50$ ns.

as well.) As a reference, we also give the PER curves of the soft-detector for the SISO system (with the data rate being 54 Mbps). We see, from the PER curves, that both the preamble design and the sequential channel parameter estimation algorithm are effective in that the gap between the PER curves corresponding to the perfect channel knowledge case and the estimated channel parameter case for the MIMO system is no more than that of the SISO system. We also see that the MIMO soft-detector is effective in that the MIMO system needs only 2 to 3 dB extra total transmission power to keep the same PER (we are mostly interested in PERs being 0.1, according to the IEEE 802.11a standard) as its SISO counterpart, but with the data rate doubled.

Note that the MIMO soft-detector is outperformed by LSD in terms of PER. However, the MIMO soft-detector is orders of magnitude more efficient than LSD, and is ideally suited for real-time implementations.

6. CONCLUDING REMARKS

We have proposed a preamble design for the MIMO system with two transmit and two receive antennas, which is backward compatible with its SISO counterpart as specified by the IEEE 802.11a standard. Based on this preamble design, we have devised a sequential method for the estimation of CFO, symbol timing, and MIMO channel response. We have also provided a simple MIMO soft-detector. Both the sequential parameter estimation method and the soft-detector are ideally suited for real-time implementations.

Appendix A. Phase Correction using Pilot Symbols

As we mentioned earlier, each data OFDM symbol contains four known pilot symbols. We denote these pilot symbols by a 4×1 vector $\mathbf{p}$. The pilot symbols can be used to correct the CFO residual phase for each data OFDM symbol after CFO correction using $\hat{\epsilon}_F$. Let $\mathbf{y}_n^{(\mathrm{p})}$ be the vector containing the corresponding four elements of the FFT output of an data OFDM symbol in the OFDM DATA field received from the nth antenna, $n = 1, 2, \cdots, N$. Let $\hat{\mathbf{h}}_{n,m}^{(\mathrm{p})}$ be the 4×1 estimated channel vector from transmit antenna m to receive antenna n for the four corresponding subcarriers. Let $\mathbf{P} = \mathrm{diag}\{\mathbf{p}\}$. We have $\mathbf{y}_n^{(\mathrm{p})} = e^{j\phi}\mathbf{P}\sum_{m=1}^{2} \hat{\mathbf{h}}_{n,m}^{(\mathrm{p})} + \mathbf{e}_n^{(\mathrm{p})}, n = 1, 2, \cdots, N$, where ϕ is the CFO residual phase and $\{\mathbf{e}_n^{(\mathrm{p})}\}_{n=1}^{N}$ are zero-mean white circularly symmetric complex Gaussian noise vectors. Then the ML criterion leads to

$$\hat{\phi}_{\mathrm{ML}} = \arg\min_{\phi} \sum_{n=1}^{N} \left\| \mathbf{y}_n^{(\mathrm{p})} - e^{j\phi}\mathbf{P} \sum_{m=1}^{2} \hat{\mathbf{h}}_{n,m}^{(\mathrm{p})} \right\|^2 = \angle \left\{ \sum_{n=1}^{N} \left[\sum_{m=1}^{2} \left(\hat{\mathbf{h}}_{n,m}^{(\mathrm{p})} \right)^H \right] \mathbf{P}^H \mathbf{y}_n^{(\mathrm{p})} \right\}. \quad (18)$$

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