Layered Turbo Space-Time Coded MIMO-OFDM Systems for Time Varying Channels

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***Abstract*— A joint detection-estimation scheme is proposed for multiple-input multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) systems. The transmitter employs a layered turbo space-time code. The iterative scheme proposed for the receiver integrates in one scheme turbo decoding, co-antenna nterference suppression and channel estimation. A main observation is that the high coding gain of the turbo space-time code can compensate for the application of simple co-antenna interference suppression scheme and can also facilitate good channel estimates. The performance of the MIMO-OFDM system is validated by simulations under various channel conditions including high Doppler. Numerical results show that for a 4 transmit 4 receive antenna (4T4R) system, with a data rate of 4 Mb/s over a 1.25 MHz bandwidth, a packet error rate (PER) of 10^{-2} is achieved at a signal to noise ratio (SNR) of 16 dB for a typical urban channel with Doppler frequencies as high as 200 Hz. For a 4T6R system, the same performance can be achieved for only 6 dB SNR.**

I. Introduction

Multiple-input multiple-output (MIMO) systems have been studied and applied to improve both data rate and reliability of wireless links. Likewise, orthogonal frequency division multiplexing (OFDM) is a wideband technology designed to reduce or eliminate intersymbol interference (ISI) over the multipath fading channel by dividing the whole bandwidth into smaller parallel subbands and consequently increasing the symbol duration. The combination of MIMO and OFDM is a promising approach for high data rate wireless communications.

With MIMO systems, channel coding is applied across the transmit antennas. When the number of transmit antenna is large, the complexity of the optimal receiver increases exponentially with the number of transmit antennas. To address this problem, reduced complexity layered structure were introduced such as [1]-[3], where the signals from each transmit antenna (or subset of transmit antennas) referred to as a *layer*, are first separated and then decoded. Signal separation can be achieved through well-known signal processing techniques such as zero forcing (ZF) or minimum mean square error (MMSE) [1]-[3]. Decoding and co-antenna interference suppression in MIMO systems can be viewed as signal-to-interference-noise ratio (SINR) transformers working in tandem. The combined effect of these operations has to provide sufficient SINR processing gain for satisfactory performance. With ordinary trellis coding, it befalls the interference suppression operation to provide sufficient processing gain to guarantee good performance. This leads to high complexity adaptive filters such as MMSE. Furthermore, in practical systems, filter coefficients are derived based on channel estimates. Mismatch between the estimated and real MIMO channel gains leads to inaccurate filter parameters and to self interference. All this suggests an alternative approach, where more of the processing gain is supplied by the channel encoder. Suitable encoders for this application are powerful turbo space-time codes [4], [5].

When the number of transmit antennas is large (≥ 4), this type of coding is applied in layered form, referred to as *layered turbo space-time coding* (LTST). The ability of turbo codes in general, and LTST in particular, to provide good performance at low SINR can compensate for the presence of interference, and the turbo decoding can be initiated before the interference from other layers has been canceled. With each iteration, soft decisions are used to subtract out the interference in lieu of the more complex adaptive filtering.

To further exploit the iterative nature of turbo decoding and to address the time varying nature of the wireless channel, we have also developed a novel iterative channel estimation algorithm. This algorithm is integrated with the turbo decoder and interference canceler. In MIMO-OFDM, channel estimation is a particularly challenging problem since channels have to be estimated across antennas, frequency and time. Non-iterative channel estimation techniques for MIMO-OFDM were proposed in [6], [7]. Therein, the channel is estimated from a training block or based on the detected previous data block. Since the techniques are non-iterative, mismatches between estimated and actual channel gains cannot be compensated. We propose an algorithm that estimates the channel state information (CSI) iteratively using decisions feedback from the decoder. As a result, the whole system is more robust against time variation of the channel compared to non-iterative systems. The channel estimator is an extension of our previous work [8] to the OFDM case. The channel estimation algorithm is efficient in the sense that it avoids repeated matrix inversions and it is stable even when the channel gain matrix is ill-conditioned.

The rest of the paper is organized as follows. In section II we present the system model. The turbo decoder and interference canceler are presented in section III. Channel estimation is developed in section IV. In section V, we present the numerical results, and section VI draws the conclusions.

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II. System Model

Consider an OFDM system with n_T transmit and n_R receive antennas (n_TTn_RR). The proposed transceiver is shown in Fig. 1. Part (a) of the figure shows the structure of the transmitter. The information data is encoded by a symbol based, parallel concatenated space-time turbo encoder as proposed in [4], [5]. The turbo encoder generates codewords represented by two-row matrices. One row of the codeword matrix contains systematic symbols and the other row contains the corresponding parity symbols. The codeword is then interleaved *pairwise* such that columns of the codeword matrix are preserved. Following interleaving, the codeword is separated into several substreams, which are modulated, transformed to time domain signals by inverse fast Fourier transform (IFFT), and transmitted from groups of two antennas. The substream of the signals transmitted from each antenna group is referred to as a *layer*. The function of the interleaver is to decorrelate adjacent codeword symbols and avoid possible burst errors due to imperfect interference suppression at the receiver.

At the receiver, following the fast Fourier transform (FFT), the signals at each subcarrier are a superposition of the signals from all transmit antennas

$$\mathbf{r}[n,k] = \mathbf{H}[n,k]\mathbf{s}[n,k] + \boldsymbol{\eta}[n,k], \quad k = 1,...,K, \;\; n = 1,2,..., \tag{1}$$

where K is the number of subcarriers; n is time index indicating the nth OFDM block; $\mathbf{s}[n,k] \triangleq \{s_j[n,k]\}$ and $\mathbf{r}[n,k] \triangleq \{r_i[n,k]\}$ are respectively, $n_T \times 1$ and $n_R \times 1$ vectors of the transmitted and received signals at the kth subcarrier and the nth OFDM block; $\mathbf{H}[n,k] \triangleq \{H_{i,j}[n,k]\}$ is a $n_R \times n_T$ matrix denoting the channel frequency response between transmit antenna j and receive antenna i. Symbols are chosen from a two-dimensional constellation such as PSK or QAM. The term $\boldsymbol{\eta}[n,k]$ is additive white Gaussian noise (AWGN). Without loss of generality, we assume that the symbol average energy is unity and that the AWGN variance is $n_T/(2\mathrm{SNR})$ per dimension.

III. Iterative Decoding and Co-Antenna Interference Cancellation

In this section we present the iterative algorithm for joint turbo decoding, co-antenna interference cancellation and channel estimation. The receiver is shown in Fig. 1-b, with two component maximum *a posteriori* (MAP) decoders that work in an iterative fashion. At each iteration, the signals received from different layers are first separated by cancelling the soft decision of the signals from other layers (regarded as co-antenna interference). Then, the separated signals of each layer are re-assembled, de-interleaved to the original order and passed to the MAP decoders to be decoded. The two decoders also exchange extrinsic information as in usual turbo decoding. The MIMO channel gain matrix used in decoding and cancellation is updated by the channel estimation algorithm. This process repeats for several iterations until some stop criteria is met.

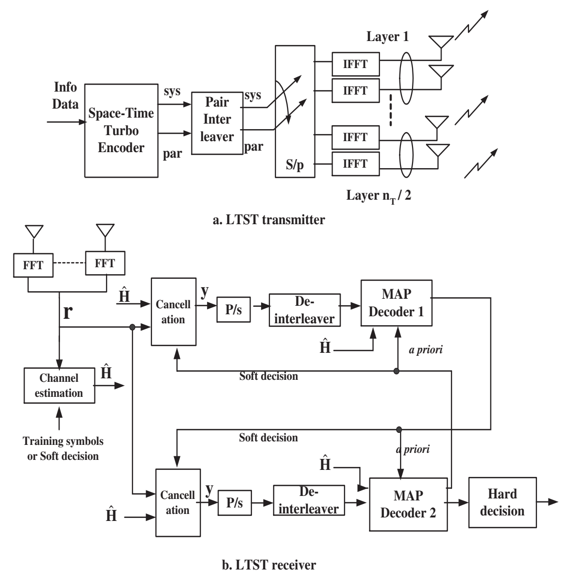

Fig. 1. MIMO-OFDM transceiver; (a) tranmistter; (b) receiver.

A. Parallel Co-antenna Interference Cancellation

To separate between desired signal and interference components, and assuming layers that consist of two transmit antennas, the channel model (1) is rewritten

$$\mathbf{r}[n,k] = \mathbf{H}_l[n,k]\mathbf{s}_l[n,k] + \sum_{m=1, m\neq l}^{n_T/2} \mathbf{H}_m[n,k]\mathbf{s}_m[n,k] + \boldsymbol{\eta}[n,k], \tag{2}$$

where $\mathbf{H}_l[n,k]$ and $\mathbf{s}_l[n,k]$, $l = 1,...,n_T/2$ are respectively, the channel gain matrix and transmitted symbols pertaining to the kth subcarrier of the lth layer (layer that represents the desired signal). Note that in the proposed scheme, each layer contains signals from two transmit antennas, one for the systematic symbol and one for its corresponding parity symbol. Hence $\mathbf{H}_l[n,k]$ is a $n_R \times 2$ matrix containing the $(2l)$th and $(2l+1)$th columns of $\mathbf{H}[n,k]$, and $\mathbf{s}_l[n,k]$ is a 2×1 vector containing the two symbols $s_{2l}[n,k]$ and $s_{2l+1}[n,k]$, transmitted from the $(2l)$th and $(2l+1)$th antenna. The second term on the right hand side of (2) represents the co-antenna interference from other layers.

Denote the soft decisions of the transmitted signals from the mth layer at the uth iteration, $\widehat{\mathbf{s}}_m^{(u)}[n,k]$, $m = 1,...,n_T/2$. Observations used by the MAP decoder of the lth layer, $\mathbf{y}_l^{(u)}[n,k]$, are obtained by subtracting out signals from other layers, followed by a spatial matched filter $\mathbf{H}_l[n,k]$, i.e.,

$$\begin{aligned}\mathbf{y}_l^{(u)}[n,k] &= \mathbf{H}_l^{\dagger}[n,k]\left(\mathbf{r}[n,k] - \sum_{m=1 m\neq l}^{n_T/2} \mathbf{H}_m[n,k]\widehat{\mathbf{s}}_m^{(u)}[n,k]\right) \\ &= \mathbf{H}_l^{\dagger}[n,k]\mathbf{H}_l[n,k]\mathbf{s}_l[n,k] + \mathbf{z}_l^{(u)}[n,k] + \boldsymbol{\zeta}_l[n,k], \end{aligned} \tag{3}$$

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where

$$\mathbf{z}_l^{(u)}[n,k] \triangleq \mathbf{H}_l^{\dagger}[n,k] \sum_{m=1,m\neq l}^{n_T/2} \mathbf{H}_m[n,k] \left(\mathbf{s}_m[n,k] - \widehat{\mathbf{s}}_m^{(u)}[n,k]\right) \tag{4}$$

is the residual interference term, and $\boldsymbol{\zeta}_l[n,k] = \mathbf{H}_l^{\dagger}[n,k]\boldsymbol{\eta}[n,k]$ is the noise term; the symbol '$\dagger$' denotes conjugate transpose. The signals $\left\{\mathbf{y}_l^{(u)}[n,k], l = 1, ..., n_T/2, k = 1, ..., K\right\}$ are rearranged to the original order, and passed to the MAP decoder (see Fig. 1(b)).

It is shown in the numerical results section that for LTST, the simple matched filter $H_l[n,k]$ suffices and that sophisticated filters such as MMSE [3] and ZF [1], [6] are not required for co-antenna interference suppression. Since these algorithms generally require matrix inversion, they have high complexity and are more sensitive to channel estimation error.

B. Decoding and Interference Variance Estimation

The MAP decoding algorithm for LTST is derived from the turbo space-time decoder in [4]. The branch metric in the reference is modified to incorporate the cancellation of the co-antenna interference as per (3). The vectors $\mathbf{y}_l^{(u)}[n,k]$ serve as observations for evaluating branch metrics in MAP decoding. Let the $\mathbf{c}$ denote the vector of codeword symbols associated with a given transition branch. The branch metric is then given by the log-likelihood

$$\begin{aligned} M\left(\mathbf{y}_l^{(u)}[n,k], \mathbf{c}\right) &= \left(\mathbf{y}_l^{(u)}[n,k] - \mathbf{H}_l^{\dagger}[n,k]\mathbf{H}_l[n,k]\mathbf{c}\right)^{\dagger} \quad (5)\\ &\quad \mathbf{K}^{-1}\left(\mathbf{y}_l^{(u)}[n,k] - \mathbf{H}_l^{\dagger}[n,k]\mathbf{H}_l[n,k]\mathbf{c}\right), \end{aligned}$$

where $\mathbf{K} = E\left[\left(\mathbf{z}_l^{(u)}[n,k] + \boldsymbol{\zeta}_l[n,k]\right)\left(\mathbf{z}_l^{(u)}[n,k] + \boldsymbol{\zeta}_l[n,k]\right)^{\dagger}\right]$ is the residual interference plus noise covariance matrix. To lower the receiver complexity and without visible penalty in performance, the following suboptimal branch metric that treats the residual interference as AWGN is used:

$$M_1\left(\mathbf{y}_l^{(u)}[n,k], \mathbf{c}\right) = \frac{1}{N_0^{(u)}[n,k]} \left\|\mathbf{y}_l - \mathbf{H}_l^{\dagger}[n,k]\mathbf{H}_l[n,k]\mathbf{c}\right\|^2, \tag{6}$$

where

$$\begin{aligned} N_0^{(u)}[n,k] &= 1/n_R \ \operatorname{tr}(\mathbf{K}) \quad (7)\\ &= 1/n_R \left(E\left[\left\|\mathbf{z}_l^{(u)}[n,k]\right\|^2\right] + E\left[\|\boldsymbol{\zeta}_l[n,k]\|^2\right]\right) \end{aligned}$$

is the variance of residual interference plus noise. The variance of the noise term $E\left[\|\boldsymbol{\zeta}_l[n,k]\|^2\right]$ can be estimated from the pilots. The variance of the residual interference $E\left[\left\|\mathbf{z}_l^{(u)}[n,k]\right\|^2\right]$ is estimated from (4):

$$E\left[\left\|\mathbf{z}_l^{(u)}[n,k]\right\|^2\right] = \left\|\mathbf{H}_l[n,k]^{\dagger} \sum_{m=1,m\neq l}^{n_T/2} \mathbf{H}_m[n,k]\right\|^2 \cdot \sigma^{(u)}, \tag{8}$$

where $\sigma^{(u)}$ denotes the variance of the soft decision error of the symbols. This variance is estimated

$$\widehat{\sigma}^{(u)} = \frac{1}{Kn_T} \sum_{j=1}^{n_T} \sum_{k=1}^{K} \left\|\widehat{s}_j^{(u)}[n,k] - \overline{s}_j^{(u)}[n,k]\right\|^2, \tag{9}$$

where $\widehat{s}_i^{(u)}[n,k]$ and $\overline{s}_i^{(u)}[n,k]$ are respectively the soft and hard decision of the transmitted symbol on the kth subcarrier from the jth antenna.

The soft decision $\widehat{s}_i[n,k]$ is computed as

$$\widehat{s}_i^{(u)}[n,k] = \sum_{q=1}^{Q} b_q P^{(u)}\left(s_i[n,k] = b_q|\text{obs}\right), \tag{10}$$

where $\{b_q, \quad q = 1, ..., Q\}$ is the modulation constellation set and $P^{(u)}\left(s_i[n,k] = b_g|\text{obs}\right)$ are *a posteriori probabilities* (APP's) computed by the decoder. The soft decision $\widehat{s}_i^{(u)}[n,k]$ is also used in (3) to cancel the co-antenna interference. At the first iteration, when no estimates from the decoder are available, $\left\{\widehat{s}_i^{(1)}[n,k], k = 1, ..., K, i = 1, ..., n_T\right\}$ are set to zero.

IV. Channel Estimation

In this section we present the decision directed channel estimation algorithm and its integration with the iterative turbo-decoder. The estimation algorithm is based on the packet transmission pattern, where each packet contains one training block followed by a number of data blocks [6]. The training block supplies the initial CSI; subsequently, the CSI is re-estimated during each data block.

First, we present the channel estimation algorithm. It utilizes the soft decision of the transmitted data sequence or the training sequence to estimate the MIMO channel gain matrix in the time domain. The frequency domain channel gain matrices $\{\mathbf{H}[n,k]\}_{k=0}^{K-1}$ used in decoding and interference cancellation are obtained by performing IFFT on the estimated time domain channel gain matrices.

Rewrite (1) to emphasize that the channel gains are unknown

$$r_i[n,k] = \sum_{j=1}^{n_T} s_j[n,k]H_{ij}[n,k] + \eta_i[n,k] \tag{11}$$

For OFDM systems with proper cyclic extension in time, it has been shown that, with tolerable leakage, the channel frequency response between the jth transmit and ith receive antennas at the kth subcarrier and nth OFDM block can be expressed as

$$H_{ij}[n,k] = \sum_{\ell=0}^{L_0-1} h_{ij}[n,\ell]e^{-\jmath 2\pi k\ell/K} = \mathbf{w}[k]^T\mathbf{h}_{ij}[n], \tag{12}$$

where $\mathbf{w}[k] = \left(e^{-\jmath 0}, e^{-\jmath 2\pi k/K}, \cdots, e^{-\jmath 2\pi k(L_0-1)/K}\right)^T$; the $L_0 \times 1$ vector $\mathbf{h}_{ij}[n] = (h_{ij}[n,0], h_{ij}[n,1], \cdots, h_{ij}[n, L_0 - 1])^T$ expresses the time domain impulse response of the channel, 'T' denotes transpose; L_0 is the number of taps which is determined by the time spread of the channels and the time resolution. Rewriting (11) in vector form with $\mathbf{r}_i[n] = (r_i[n,0], \cdots, r_i[n,K-1])^T$, we have

$$\mathbf{r}_i[n] = \sum_{j=1}^{n_T} \mathbf{S}_j[n]\mathbf{H}_{ij}[n] + \boldsymbol{\eta}[n], \quad (13)$$

where $\mathbf{S}_j[n] = \mathrm{diag}\,(s_j[n,0], s_j[n,1], \cdots, s_j[n,K-1])$, and $\mathbf{H}_{ij}[n], \boldsymbol{\eta}[n]$ are defined analogous to $\mathbf{r}_i[n]$. After a few manipulations and using (12) in (13), we obtain

$$\mathbf{r}_i[n] = \mathbf{S}[n]\mathbf{W}\mathbf{h}_i[n] + \boldsymbol{\eta}[n], \quad (14)$$

where $\mathbf{S}[n] = (\mathbf{S}_1[n], \cdots, \mathbf{S}_{n_T}[n])$, $\mathbf{W} = \mathbf{I}_{n_T} \otimes (\mathbf{w}[0], \cdots, \mathbf{w}[K-1])^T$, $\mathbf{I}_{n_T}$ is the unity matrix of dimension n_T, the symbol $\otimes$ denotes Kronecker product, and $\mathbf{h}_i[n] = \left(\mathbf{h}_{i1}[n]^T, \cdots, \mathbf{h}_{in_T}[n]^T\right)^T$. For channel estimation purposes, $\mathbf{S}[n]$ is formed either by the training sequence or the decisions on the transmitted symbol vectors.

The channel impulse response for the nth OFDM block is found by the method least-squares. For $n_T L_0 \le K$, (14) affords the following least-squares solution

$$\hat{\mathbf{h}}_i[n] = \mathbf{Q}[n]^{-1}\mathbf{p}_i[n], \quad (15)$$

where $\mathbf{Q}[n] = \mathbf{W}^\dagger\mathbf{S}^\dagger[n]\mathbf{S}[n]\mathbf{W}$, and $\mathbf{p}_i[n] = \mathbf{W}^\dagger\mathbf{S}[n]^\dagger\mathbf{r}_i[n]$.

Computing (15) involves inversion of the matrix $\mathbf{Q}[n]$, the latter being, in general, dependent on the transmitted data $\mathbf{S}[n]$. To reduce the computational complexity due to repeated matrix inversions, we propose an iterative method to carry out (15). Moreover, these iterations will be integrated with the decoding iterations. To proceed, let $\mathbf{Q}[n] = \mathbf{M} - \mathbf{R}[n]$ be any decomposition of $\mathbf{Q}[n]$ such that $\mathbf{M}$ is nonsingular, and let $\hat{\mathbf{h}}_i^{(0)}[n]$ be an arbitrary initial vector. Then the sequence $\hat{\mathbf{h}}_i^{(0)}[n], \hat{\mathbf{h}}_i^{(1)}[n], \cdots$, generated by the following iteration

$$\mathbf{M}\hat{\mathbf{h}}_i^{(v+1)}[n] = \mathbf{R}[n]\hat{\mathbf{h}}_i^{(v)}[n] + \mathbf{p}_i[n] \quad (16)$$

converges to the true solution if and only if the spectral radius of the iteration matrix $\mathbf{B}[n] = \mathbf{M}^{-1}\mathbf{R}[n]$ satisfies $\rho(\mathbf{B}) < 1$, where the spectral radius $\rho(\mathbf{B})$ is defined as the largest modulus of the eigenvalues of $\mathbf{B}$ [9]. Different decompositions of $\mathbf{Q}[n]$ result in well-known iterative methods such as Jacobi, Gauss-Seidel, and successive overrelaxation (SOR) iteration. Details on the convergence and computation aspects of such methods can be found in mathematical literature, e.g., [9]. To reduce the computational complexity, $\mathbf{M}$ is chosen as an identity matrix, i.e.,

$$\begin{aligned} \mathbf{M} &= m\mathbf{I} \\ \mathbf{R}[n] &= m\mathbf{I} - \mathbf{W}^\dagger\mathbf{S}^\dagger[n]\mathbf{S}[n]\mathbf{W}, \end{aligned} \quad (17)$$

where m is some constant. With $\mathbf{M}$ and $\mathbf{R}[n]$ defined in (17), the iteration becomes

$$\hat{\mathbf{h}}_i^{(v+1)}[n] = \frac{1}{m}\left(\left(m\mathbf{I} - \mathbf{W}^\dagger\mathbf{S}^\dagger[n]\mathbf{S}[n]\mathbf{W}\right)\hat{\mathbf{h}}_i^{(v)}[n] + \mathbf{p}_i[n]\right), \quad (18)$$

where v is the iteration index. Let λ be any eigenvalue of $\mathbf{R}[n]$, then $m - Kn_T \le \lambda < m$. It follows that $\rho(\mathbf{B}) = \rho\left(\mathbf{M}^{-1}\mathbf{R}[n]\right) < 1$, if $m > Kn_T/2$. After $\{\hat{\mathbf{h}}_i\}_{i=1}^{n_R}$ are obtained, the frequency domain channel gains $\{\widehat{\mathbf{H}}[n,k]\}_{k=1}^{K}$ are found by performing FFT on $\{\hat{\mathbf{h}}_i\}_{i=1}^{n_R}$.

Since both computing (18) and the turbo decoding are iterative, they can be implemented jointly to make the receiver more efficient. This is done by computing (18) for one more iteration after either of the two component decoders finishes one MAP decoding and outputs new soft decisions of the transmitted symbols. The updated channel state estimates are then used in the next decoding. That is, (18) is modified as

$$\hat{\mathbf{h}}_i^{(v+1)}[n] = \frac{1}{m}((m\mathbf{I} - \mathbf{W}^\dagger\widehat{\mathbf{S}}^{(v)\dagger}[n]\widehat{\mathbf{S}}^{(v)}[n]\mathbf{W})\hat{\mathbf{h}}_i^{(v)}[n] + \mathbf{p}_i[n]), \quad (19)$$

where $\mathbf{S}^{(v)}[n]$ are soft decisions of the transmitted symbols at channel estimation iteration ν.

For estimating the channel using the training block, we found that a single iteration of (18) is sufficient to obtain accurate channel estimates, if the optimum training block suggested in [7] is used.

To summarize, the iterative joint decoding and channel estimation algorithm works as follows:

(a) For the nth data block in a packet, if $n > 1$, start turbo decoding using the channel state estimates from the previous data block; for $n = 1$, start decoding using the channel state estimates from the training block.

(b) After some MAP decoding iterations, start channel estimation using (19). Updated channel state estimates are used in the next decoding iteration.

(c) Decoding iterations end when two consecutive iterations yield the same hard decisions, or a maximum number of iterations is reached.

V. Numerical Results and Discussions

Simulation results are presented for the proposed joint detection-estimation system for a typical urban channel model designated (TUx_6a) [10]. Performance was evaluated for Doppler frequencies 40, 100 and 200Hz, respectively. The system consists of 4 transmit antennas and different number of receive antennas. To facilitate comparison with the literature, simulation parameters were set to be the same as in [6]. The bandwidth was 1.25 MHz divided into 256 subcarriers. Two subcarriers on each end were used as guard tones. To ensure orthogonality of the subcarriers, the symbol duration was chosen 204.8 μs with an additional 20.2 μs guard interval, resulting in a total block length of 225 μs. The turbo space-time code was the rate 1/2, 4-PSK code introduced in [5]. Consequently, the data rate was 4 Mb/s over a 1.25 MHz bandwidth resulting in a transmission efficiency of 3.2 b/s/Hz. The signal to noise ratio (SNR) was measured as the total power from all transmit antennas to noise power ratio.

The performance measured was packet error rate (PER). Each packet contained 10 OFDM blocks. The first block is for training and the rest contain data. In the channel estimation algorithm, the number of time domain channel impulse response taps L_0 was set to 7. For systems with different number of receive antennas systems and for channels with different Doppler frequencies, the average number of iterations used was 3 to 5 at the operating point of PER=1%. The average number of iteration used for the channel estimation (19) varied from 3 to 6. Note computing (19) iteratively

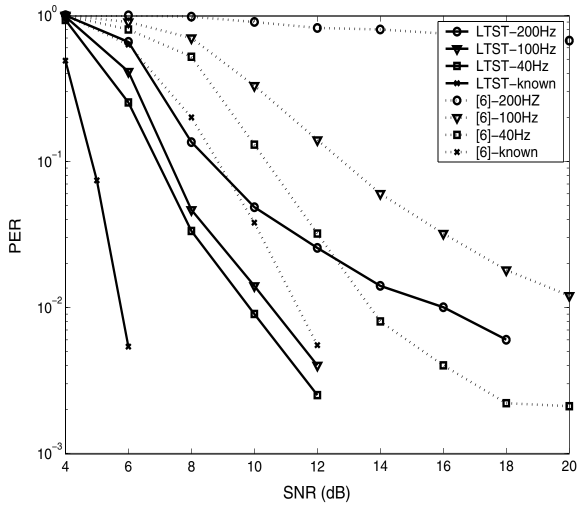

Fig. 2. PER performance of LTST 4T4R and comparison to ref. [6], for different Doppler frequences.

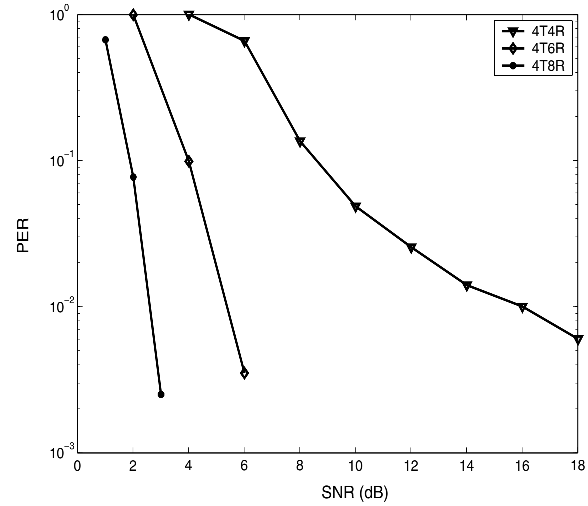

Fig. 4. PER performance comparison of LTST with 4 transmit antenna and differnt number of receive antennas, for Doppler frequence 200Hz.

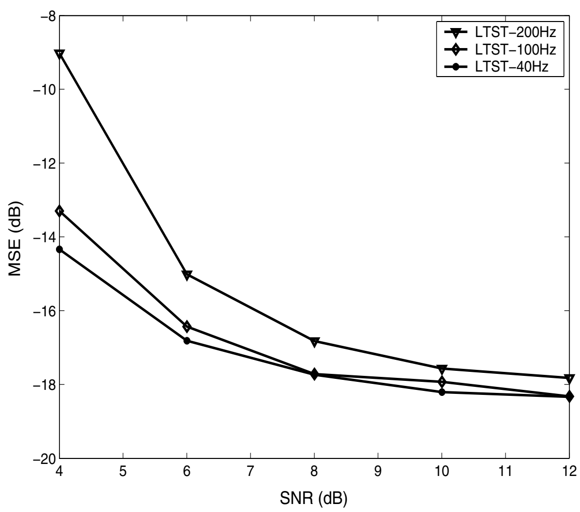

Fig. 3. MSE of the channel estimation of LTST 4T4R, with different Doppler frequences.

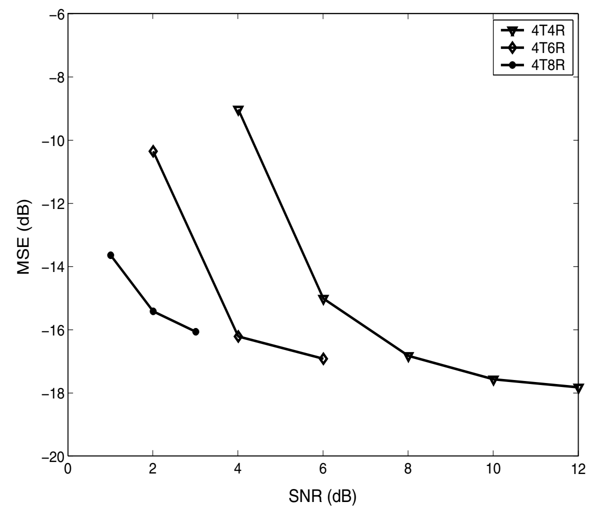

Fig. 5. MSE comparison of channel estimation algorithm used in LTST with 4 transmit antenna and differnt number of receive antennas, for Doppler frequence 200Hz.

avoids matrix inversion and reduces complexity compared to algorithms using matrix inversion.

The PER performance of LTST is presented in Fig. 2 with comparison to the non-iterative scheme proposed in [6], where a 16 state, rate 1/2 space-time code is used with 4-PSK. It is observed that for ideal (known) channel, LTST achieves PER=0.01 at an SNR of 6 dB, an improvement of 6 dB over scheme in [6]. For Doppler frequencies of 100 and 40 Hz, LTST achieves PER=0.01 at respectively, 11 and 9.5 dB, with corresponding improvements of 9 and 4.5 dB over the reference. For a Doppler frequency of 200 Hz, the scheme in [6] is virtually unusable, while LTST achieves PER=0.01 at 16 dB. The conclusion is that LTST is more robust with respect to channel Doppler and the SNR improvement is significant. This justifies the moderate complexity increase mainly due to turbo decoding. The MSE of the channel estimation for different Doppler frequencies is presented in Fig 3.

In Fig. 4, we show the effect of increased number of receive antennas on the performance of LTST, when the channel Doppler frequency is 200 Hz. It is observed from Fig. 4 that the SNR needed to achieve PER=0.01 improves by 10 dB if 6 receive antenna are used instead of 4. When 8 receive antennas are used, the improvement increases to 13 dB. The SNR improvement is also observed in Fig. 5, which shows the MSE of the channel estimation error for different number of receive antennas. From these two figures, we conclude that a moderate increase in the number of receive antennas will significantly enhance the system performance for high Doppler frequencies.

VI. Conclusions

In this paper we proposed a joint detection-estimation scheme for MIMO-OFDM systems. The scheme consists of linked iterations of a turbo-decoder and an efficient channel estimator. The joint-detector estimator demonstrates good performance over time varying channels even at relatively high Doppler frequencies. It is shown by numerical simulation that for a 4T4R configuration with a data rate of 4 Mb/s over a 1.25 MHz channel and over a channel with Doppler as high as 200 Hz, 1% PER performance is achieved at an SNR of 16 dB. When the number of receive antenna is increased to 6, the same performance can be achieved at an SNR of only 6 dB. While there is moderate complexity increase due to turbo decoding, the proposed scheme has the advantage of avoiding high complexity interference suppression techniques. Meanwhile, the moderate complexity increase is justified by significant performance improvement over previously proposed schemes.

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