ITERATIVE LAYERED SPACE-TIME TRANSCEIVER FOR ISI WIRELESS CHANNELS

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ABSTRACT

In this paper, we propose a practical iterative transceiver design for multiple-input multiple-output (MIMO) frequency-selective wireless channels, which is able to realize a significant portion of the capacity promised by information theory. At the transmitter end, we exploit the idea of space-time bit-interleaved coded modulation (ST-BICM) scheme by using turbo or convolutional codes. This encoding scheme is highly effective if used in conjunction with receiver employing iterative detection and decoding. At the receiver end, we propose a novel multi-antenna detection scheme, which equalizes the intersymbol interferences (ISI) and soft estimate the transmitted symbols. These symbols are then input to a sub-optimal turbo-like receiver that performs iterative decoding of the ST-BICM codes in an iterative and, most important, simple fashion. The simulation results show that the proposed so called Turbo-MIMO transceiver error performance improves with the number of iterations of the decoding algorithm. This performance improvement will enhance the capacity and quality of the wireless link.

1. INTRODUCTION

In recent years, multiple-antenna techniques have become a pervasive idea that promises extremely high spectral efficiency for wireless communications. In [3], the authors developed a revolutionary iterative, or "turbo" receiver for decoding concatenated convolutional codes. Turbo codes are capable of approaching Shannon's limit of an additive white Gaussian noise (AWGN) channel capacity in a computationally feasible manner. Turbo-processing is an iterative exchange of soft decisions between two different stages of the receiver. In [6] and [9], we proposed a novel application of the turbo learning principle to MIMO flat fading wireless communications system, which result in new transceiver called Turbo-BLAST. This new system is based on a random layered space-time code and an iterative detection and decoding (IDD) receiver. See [7] and references therein for similar approach. For high-date-rate applications, such as high-speed downlink packet access (HSDPA), it might be necessary to utilize signals whose bandwidth exceeds the coherence bandwidth of the channel, which brings in the issue of frequency-selective channels. In this paper, we extend our Turbo-BLAST transceiver concepts to MIMO frequency-selective wireless channels.

Several receiver techniques have been proposed for MIMO frequency-selective channels, such as the decision-feedback MIMO equalizer [1] and [8], the frequency domain MIMO equalizer [5], and the maximum a posterior (MAP) MIMO equalizer [2]. Recently, the sampling based soft equalization for MIMO channels is proposed in [4]. Another approach to deal with frequency-selective channel is to use orthogonal frequency-division multiplexing (OFDM) techniques. However, in MIMO-OFDM system, a cyclic prefix is periodically inserted at the beginning of each OFDM symbol, which reduces the spectral efficiency. Moreover, high peak-to-power ratio of the OFDM signal, the sensitive to the phase noise, and the intersubcarrier interference constitute other impairments in MIMO-OFDM systems [4]. In this paper, we propose a novel multi-antenna equalizer in conjunction with turbo decoder that performs IDD. The simulation results show significant performance improvement, this will enhance the capacity and quality of the wireless link.

2. TRANSCEIVER DESIGN

The proposed Turbo-MIMO systems exploit the ideas of ST-BICM and the turbo processing principle in a space-time coding framework, as shown in Figure 1. Instead of employing dedicated space-time codes, the transmitter uses a one-dimensional forward error-correcting block code to encode the user's information bits. The channel encoder is followed by a pseudo-random interleaver, Π, and a space-time mapper. This configuration can be thought of as a serial concatenation of two constituent encoders separated by the interleaver, the inner encoder being the space-time mapper in conjunction with the channel. Denoting a block of informa-

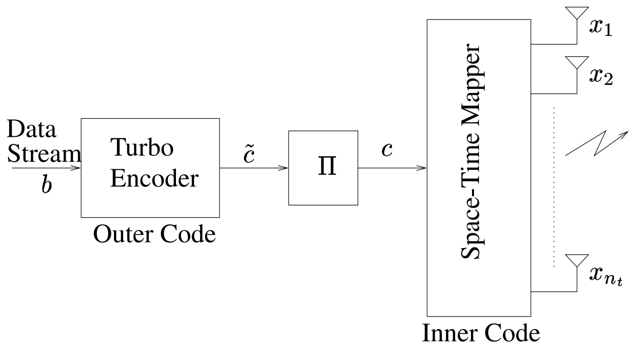

Fig. 1. Turbo-MIMO transmitter with turbo code.

tion bits by the vector b and the transfer function of the channel code by G, the codeword at the output of the outer encoder can be written as $\tilde{c} = Gb$, and $c = \Pi(\tilde{c})$ represents the interleaved sequence of code bits. The modulation format is limited herein to be identical for all transmit antennas, and the number of bits per constellation point is denoted by M_c. The space-time mapper partitions c into L subvectors $c^{(l)} = \left[\tilde{c}_{1,1}^{(l)}, \cdots, \tilde{c}_{1,M_c}^{(l)}, \tilde{c}_{2,1}^{(l)}, \cdots, \tilde{c}_{n_t,M_c}^{(l)}\right]^T$, $l = 1, \cdots, L$ and maps each of them onto a symbol vector $x^{(l)} = \left[x_1^{(l)}, \cdots, x_{n_t}^{(l)}\right]^T$ according to a unique, predetermined bit mapping scheme. To simplify notation, the superscripts $(\cdot)^{(l)}$ will be omitted whenever possible.

The frequency-selective channel can be modeled as an FIR filter whose impulse response length is W. The sampled channel response from transmitter i to receiver j is denoted by $h_{ij} = [h_{ij}(0), h_{ij}(1), \ldots, h_{ij}(W-1)]^T$. Note that the channel response includes transmit and receive filters and no time dependence within every individual burst. If we denote the discrete-time index by k, then the signal vector received at the output can be written as

$$y(k) = \sum_{l=0}^{W-1} H(l)x(k-l) + v(k) \in \mathbb{C}^{n_r \times 1}, \qquad (1)$$

where $x(k) = [x_1(k), \ldots, x_{n_t}(k)]^T$ is the $n_t \times 1$ transmit vector sequence and $v(k) = [v_1(k), \ldots, v_{n_r}(k)]^T$ is the $n_r \times 1$ zero-mean additive Gaussian white noise, i.e., $v \sim \mathcal{C}N(0, \sigma^2 I_{n_r})$, and $H(l)$ is the $n_r \times n_t$ matrix channel impulse response, i.e,

$$H(l) = \begin{bmatrix} h_{11}(l) & \cdots & h_{n_t 1}(l) \\ \vdots & \vdots & \vdots \\ h_{1 n_r}(l) & \cdots & h_{n_t n_r}(l) \end{bmatrix}, \quad l = 0, \ldots, W-1.$$

The interleaver separating the outer and inner encoders enables near-optimal decoding at reasonable computational complexity, by exploiting the principle of iterative, or turbo processing. This approach, often referred to as IDD, is similar to the decoding of serially concatenated turbo codes. In general, the outer code of an ST-BICM MIMO system can be any type of error-correcting code that can be decoded with a soft-input soft-output decoder, for example convolutional or turbo codes. Turbo or turbo-like codes are often preferred because of their exceptionally high performance. The space-time mapper provides the flexibility to design the MIMO system such as to achieve the desired trade-off between multiplexing and diversity gain. To achieve the maximum multiplexing gain, each antenna must transmit an independent information stream, i.e., the space-time mapper must be a spatial multiplexer. On the other hand, space-time block codes can be used to achieve the maximum diversity order.

2.1. Iterative detection and decoding

Iterative decoding of serially concatenated codes is a practical alternative to optimal maximum-likelihood decoding. Figure 2 shows the iterative receiver of an ST-BICM MIMO system, which separates the overall decoding problem into two stages, inner decoding (equalization) and outer decoding (channel decoding), and exchanges the information learned from one stage to another iteratively until the receiver converges. The de-interleaver, Π^{-1}, is used to compensate for the interleaving operation used at the transmitter. Moreover, together with the interleaver, Π, it serves to

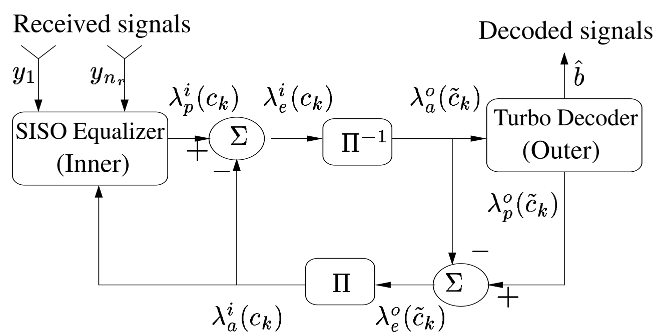

Fig. 2. Turbo-MIMO receiver.

decorrelate the output from one decoding stage before it is passed to the next. The iterative receiver updates and generally improves the soft decisions on the information bits as well as the code bits at each iteration of the information exchange process. These soft decisions are produced in the form of *a posteriori* log-likelihood ratios (LLRs), defined as

$$\lambda_p(c) = \ln \frac{P(c = +1|\cdot)}{P(c = -1|\cdot)}. \qquad (2)$$

In this equation, the probabilities are conditioned on the received signal vector y or the constraints of the channel code if the LLR is an output of the inner or outer decoder, respectively. Using Bayes' theorem, and assuming statistical independence between bits, which is a reasonable assumption because of the interleaving operation, any *a posteriori* LLR can be written as

$$\underbrace{\ln\frac{P(c=+1|y)}{P(c=-1|y)}}_{\lambda_p(c)} = \underbrace{\ln\frac{P(c=+1)}{P(c=-1)}}_{\lambda_a(c)} + \underbrace{\ln\frac{P(y|c=+1)}{P(y|c=-1)}}_{\lambda_e(c)},$$

where $\lambda_a(c)$ constitutes the *a priori* information on the bit c, and $\lambda_e(c)$ constitutes the extrinsic information. This extrinsic information is the incremental new information learned from either the received signal vector or the channel code constraints, using the available *a priori* information. The extrinsic information produced by the inner decoding stage is used as *a priori* information by the outer decoder, and vice versa.

2.2. MIMO equalizer

Here, we formulate the proposed multi-antenna detection scheme which effectively equalizes intersymbol interference due to frequency-selective MIMO channels. The received signal sample at antenna j and discrete time k can be given as

$$y_j(k) = \sum_{i=1}^{n_t}\sum_{l=0}^{W-1} h_{ij}(l)x_i(k-l) + v_j(k), \quad (3)$$

where $h_{ij}(l)$ is the channel response of the ith transmitter to the jth receiver path, $x_i(k)$ is the ith antenna transmitted symbol at k time instant, and $v_j(k)$ represents the additive Gaussian white noise. The signal model (3) can be extended to a stacked block data model by stacking $N+W-1$ received signals of $y_j(k)$'s into an $(N+W-1)\times 1$ vector $y_j = [y_j(1),\dots,y_j(N+W-1)]^T$. This stacked received signal at antenna j of the data block of N data symbols can be written as

$$y_j = H_j x + v_j \in \mathbb{C}^{(N+W-1)\times 1}, \quad (4)$$

where $x = \mathrm{vec}([x_1 \cdots x_{n_t}])$, $x_i = [x_i(1),\dots,x_i(N)]^T$, $v_j = [v_j(1),\dots,v_j(N+W-1)]^T$ and the matrix $H_j = [H_{1j}, H_{2j},\dots,H_{n_tj}] \in \mathbb{C}^{(N+W-1)\times Nn_t}$ has a block-Toeplitz structure with

$$H_{ij} = \mathrm{diag}(h_{ij},\dots,h_{ij}) \in \mathbb{C}^{(N+W-1)\times N}.$$

Note that, for flat fading channel, H_j is an $N\times Nn_t$ matrix. The n_r antenna received signals are combined as

$$\begin{aligned} y = \sum_{j=1}^{n_r}[H_j]^H y_j &= \sum_{j=1}^{n_r}[H_j]^H[H_j]x + [H_j]^H v_j \\ &= Hx + v \in \mathbb{C}^{Nn_t\times 1}, \end{aligned} \quad (5)$$

where $H = \sum_{j=1}^{n_r}[H_j]^H[H_j]$ and $v = \sum_{j=1}^{n_r}[H_j]^H v_j$ represents the additive noise with covariance matrix given by

$$\begin{aligned} R_{vv} &= \mathcal{E}_v\left\{\left(\sum_{j=1}^{n_r}[H_j]^H v_j\right)\left(\sum_{j=1}^{n_r}[H_j]^H v_j\right)^H\right\} \\ &= \sigma^2\sum_{j=1}^{n_r}[H_j]^H[H_j] = \sigma^2 H. \end{aligned} \quad (6)$$

In deriving the equation (6) we assumed that the channel matrices H_j are constant and expectation with respect to noise statistics. Also we use the following fact

$$\mathcal{E}_v\left\{[v_j][v_i]^H\right\} = \begin{cases} \sigma^2 I & i=j \\ 0 & i\neq j \end{cases}.$$

From the model (5), we can use the MMSE and ZF estimator for transmitted symbols, which will be input to iterative turbo receiver, as shown in Figure 2. The ZF soft estimation of transmitted signals are given by

$$\begin{aligned} \hat{x}_{\mathrm{ZF}} &= \arg\min_x\{(y-Hx)^H R_{vv}^{-1}(y-Hx)\} \\ &= H^{-1}y \in \mathbb{C}^{Nn_t\times 1}. \end{aligned} \quad (7)$$

Let $x \sim \mathcal{C}N(0, R_{xx})$ and $R_{xx} = I$. Then the MMSE soft estimation of the transmitted signals are given by

$$\begin{aligned} \hat{x}_{\mathrm{MMSE}} &= \arg\min_x \mathcal{E}\{(\hat{x}-x)^H(\hat{x}-x)\} \\ &= (H+\sigma^2 I)^{-1}y \in \mathbb{C}^{Nn_t\times 1}. \end{aligned} \quad (8)$$

These ZF and MMSE estimators can be efficiently implemented using FFT. We omitted the details here for brevity. Next, we study the soft interference cancellation technique for iterative detection. Let $x_i \in \mathbb{C}^{N\times 1}$ be the desired signal. The channel matrix for the interference signal $x_{\mathrm{int}} = \mathrm{vec}([x_1\cdots x_{i-1}x_{i+1}\cdots x_{n_t}])$ is given by $H_{\mathrm{int},j} = [H_{1j},\dots,H_{i-1j},H_{i+1j},\dots,H_{n_tj}]$. The stacked received signal at antenna j of the data block of N data symbols can be written as $y_j = H_{ij}x_i + H_{\mathrm{int},j}x_{\mathrm{int}} + v_j \in \mathbb{C}^{(N+W-1)\times 1}$ The n_r antenna received signals are combined as

$$y = \sum_{j=1}^{n_r}[H_{ij}]^H y_j = A_i x_i + B_i x_{\mathrm{int}} + u_i \in \mathbb{C}^{N\times 1}, \quad (9)$$

where $A_i = \sum_{j=1}^{n_r}[H_{ij}]^H[H_{ij}] \in \mathbb{C}^{N\times N}$, $B_i = \sum_{j=1}^{n_r}[H_{ij}]^H H_{\mathrm{int},j} \in \mathbb{C}^{N\times N(n_t-1)}$ and $u_i = \sum_{j=1}^{n_r}[H_{ij}]^H v_j$. Let W_i be an $N\times N$ weight matrix which used to estimate the desired signal x_i from (9). Then we have

$$\hat{x}_i = W_i y = W_i A_i x_i + W_i B_i x_{\mathrm{int}} + W_i u_i \quad (10)$$

Let $b_i = W_i B_i x_{\text{int}}$ and $a_i = W_i y - b_i = W_i(y - B_i x_{\text{int}})$. To estimate the desired signal x_i, we need to find a solution to following optimization problem

$$(\hat{W}_i, \hat{b}_i) = \arg \min_{(W_i, b_i)} \mathcal{E}\{\|a_i - x_i\|^2\}. \tag{11}$$

Using standard optimization technique, the MMSE solution to (11) is given by $\hat{W}_i = (A_i A_i^H + B_i \mathcal{E}\{x_{\text{int}} x_{\text{int}}^H\} B_i^H + \sigma^2 A_i)^{-1} A_i$ and $\hat{b}_i = W_i B_i \mathcal{E}\{x_{\text{int}}\}$. If we ignoring the term $B_i \mathcal{E}\{x_{\text{int}} x_{\text{int}}^H\} B_i^H$, then the MMSE soft estimation of the desired signal is given by $\hat{x}_{i,\text{MMSE}} = \left(\sum_{j=1}^{n_r} [H_{ij}]^H [H_{ij}] + \sigma^2 I\right)^{-1} y \in \mathbb{C}^{N \times 1}, i = 1, \ldots, n_t$.

3. SIMULATION RESULTS

For the simulation we consider $n_t = n_r = 4$ and two equal power symbol space rays (multipaths). We use turbo and convolutional codes with rate $R = 1/2$ and QPSK and 16-QAM modulations. Figures 4 and 3 show the BER versus the SNR performance of convolutional and turbo codes, respectively. Note that we assume that the noise variance is equal to one and hence the SNR at the jth receive antenna is defined by $\text{SNR}_j = \frac{\rho}{n_t} \sum_{i=1}^{n_t} \mathcal{E}|h_{ij}|^2$, where ρ is the total transmitted power. Because of the local stationarity, the SNRs at all the receiver antennas are same. We also assume perfect channel knowledge at the receiver. It is clear from the figures that huge improvements in the BER are archived by proposed transceiver. This performance improvement can significantly improve the capacity and quality of the wireless link.

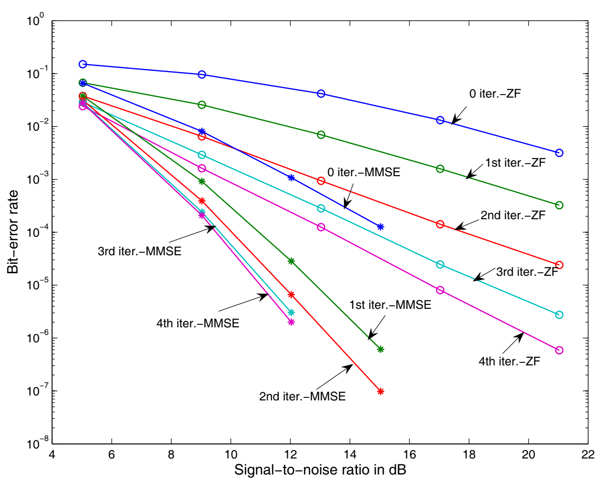

Fig. 3. BER versus SNR for various iterations with MMSE and ZF equalizers. We use convolutional code with block size of 400 bits, rate $R = 1/2$, constraint length 3 and QPSK modulation.

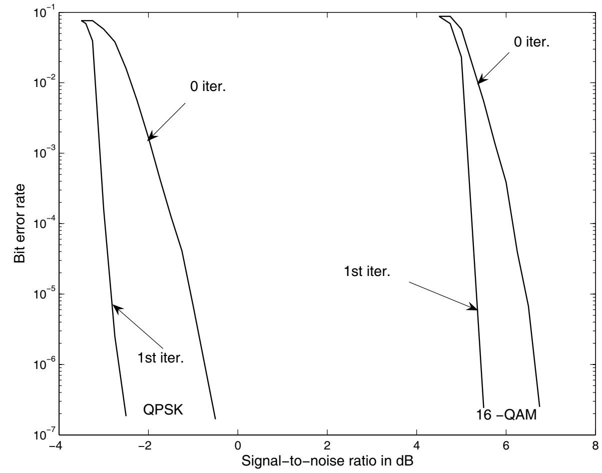

Fig. 4. BER versus SNR with MMSE equalizer. We use turbo code with block size of 16000 bits, rate $R = 1/2$, and QPSK and 16-QAM modulations.

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