An Iterative Multiuser Receiver for Space-Time Coded MIMO-OFDM System

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Abstract

Orthogonal frequency division multiplexing (OFDM) is a popular technique to mitigate ISI caused by multipath fading effect of the radio channel. In practice, OFDM could be used in combination with space-time coding, which can be performed in a multiple-input multiple-output (MIMO) system, to increase the diversity gain and/or to enhance the system capacity over time-variant and frequency-selective channels. Otherwise, the performance wireless communications is also primarily limited by multiple-access interference (MAI) in multiuser applications. In this paper, an iterative receiver structure for space-time block coded MIMO OFDM scheme is exploited with error control coding technique. We implement multi-user environment and utilize an iterative method of equlization on receiver based on minimum mean square error (MMSE) multiuser detection followed by soft-input-soft-output (SISO) convolutional decoding. The frequency domain symbols are estimated in iterative process by updating extrinsic information in terms of log-likelihood ratios (LLRs).

1. Introduction and Model

In practice, combination of OFDM and space-time block coding (STBC) (Alamouti (1998)) is a very popular and robust technique to mitigate ISI and improve the performance of communication in terms of taking advantage of the spatial diversity in MIMO transmitting systems. it is able to provide high quality and spectral efficient transmission over a fading channel. Otherwise, the performance of such STBC MIMO-OFDM systems may also seriously degrade in the presence of multiple-access interference (MAI) due to multiuser applications. Moreover, as we expect to implement future wireless communication systems at higher frequency bands, higher levels of mobility and higher capacities, the sensitivity of systems to both physical movement and channel variation will increase proportionally. Hence, we should also not ignore the influence upon the quality of wireless transmission caused by the presence of doubly-selective channel fading.

In this paper, we extend the transmitter structure of MIMO OFDM system from S. Suthaharan, A. & Nallanathan & B. Kannan (2002) for multi-user application and design a novel iterative symbol estimation scheme with error control coding techniques. A MIMO OFDM system with two users is implemented in this work, which utilize space-time block codes (STBC) and iterative MMSE multiuser detection with convolutional channel decoding by updating extrinsic information in terms of log-likelihood ratios (LLRs), whereas the proposed scheme exploits two transmit antennas for each user terminal and two receive antennas at the receiver. The system is describe in Fig. 1. All channel knowledge is assumed known perfectly throughout.

First, the incoming data stream is encoded by a half rate convolutional encoder, and then, passed through an interleaving stage. the ith user's continuous input signal blocks $\mathbf{s}_1^{(i)}$ and $\mathbf{s}_2^{(i)}$ will be encoded as two consecutive transmitting OFDM symbol blocks $\mathbf{x}_1^{(i)} = [x_1^{(i)}(0), \cdots, x_1^{(i)}(N-1)]^T$ and $\mathbf{x}_2^{(i)} = [x_2^{(i)}(0), \cdots, x_2^{(i)}(N-1)]^T$, which are the frequency domain symbols, $i = 1, 2$.

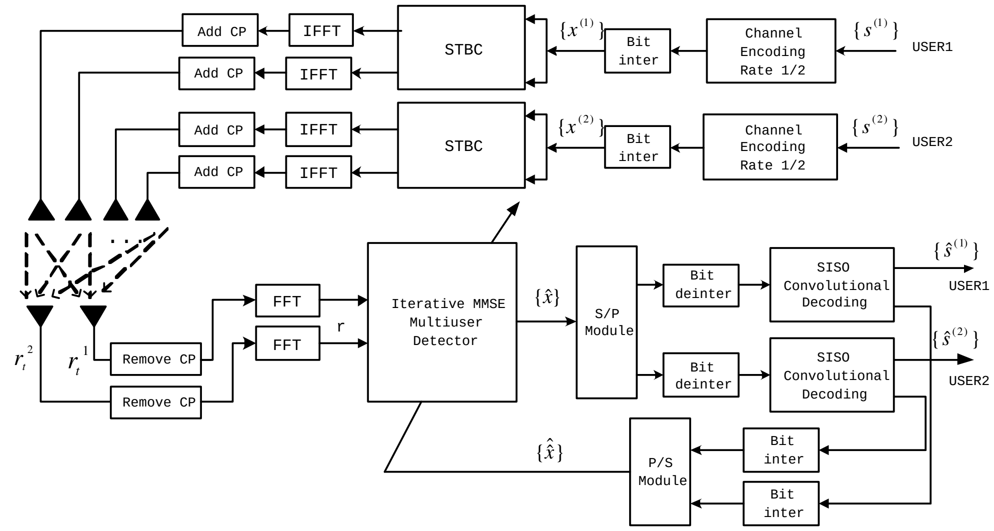

FIGURE 1. Two user convolutional coded STBC MIMO-OFDM transmission system

Then, space-time block encoding processing in implemented on the signal and the OFDM modulator is applied to the outputs of the space time encoder. The outputs of the OFDM modulators are transmitted using multiple antennas simultaneously. At the each of receive antenna, the samples corresponding to the cyclic prefix are first removed and one received OFDM block in frequency domain can be represented as

$$\mathbf{r} = \mathbf{F}\mathbf{r_t} = \mathbf{F}\mathbf{H_c^{(i)}}\mathbf{F}^H\mathbf{x}^{(i)} + \mathbf{F}\mathbf{v}^{(i)} = \mathbf{H}^{(i)}\mathbf{x}^{(i)} + \eta^{(i)} \tag{1.1}$$

where $\mathbf{H}_c^{(i)}$ is the time-domain channel (time-variant, circular) convolution matrix of size $N \times N$ passed by the *ith* user's signal, which is described in Schniter (2004). Moreover, F denotes the $N \times N$ DFT matrix and $\mathbf{H}^{(i)}$ is defined as the *subcarrier coupling matrix*. In this paper, we adapt STBC scheme which is similar to the scheme in A. F. Naguib & N. R. Canderbank (1998) for each MIMO OFDM system terminal. Here, we could assume that the time domain channel responses are constant during two consecutive OFDM block intervals, i.e. quasi-static. Hence, the signal vectors during the two sequential OFDM time-slots a and b could be obtained as $\mathbf{r}_a^1 = \mathbf{H}_{11}\begin{bmatrix}\mathbf{x}_1^{(1)}\\ \mathbf{x}_1^{(2)}\end{bmatrix} + \mathbf{H}_{21}\begin{bmatrix}\mathbf{x}_2^{(1)}\\ \mathbf{x}_2^{(2)}\end{bmatrix} + \eta_{1a}$, and $\mathbf{r}_b^{1^*} = \mathbf{H}_{21}^*\begin{bmatrix}\mathbf{x}_1^{(1)}\\ \mathbf{x}_1^{(2)}\end{bmatrix} - \mathbf{H}_{11}^*\begin{bmatrix}\mathbf{x}_2^{(1)}\\ \mathbf{x}_2^{(2)}\end{bmatrix} + \eta_{1b}^*$, where $\mathbf{H}_{qj} \triangleq [\mathbf{H}_{qj}^{(1)}, \mathbf{H}_{qj}^{(2)}]$, which $\mathbf{H}_{qj}^{(i)}$ is the frequency domain channel matrix for the *ith* user's signal transmitted from the *qth* transmit antenna to the *jth* receive antenna, and $q \in \{1,2\}$, in terms of diagonal matrix formed by the channel impulse response in frequency domain, (see in P. L. Kafle,& A. B. Sesay & J. Mcrory (2004)). For simplicity of notations, let us define signal vector $\mathbf{x} = [\mathbf{x}_1^{(1)^T}, \mathbf{x}_1^{(2)^T}, \mathbf{x}_2^{(1)^T}, \mathbf{x}_2^{(2)^T}]^T$, received vector $\mathbf{r}^1 = \begin{bmatrix}\mathbf{r}_a^1\\ \mathbf{r}_b^{1^*}\end{bmatrix}$, and $\eta_1 = \begin{bmatrix}\eta_{1a}\\ \eta_{1b}^*\end{bmatrix}$. Therefore, we could write $\mathbf{r}_a^1$ and $\mathbf{r}_b^{1^*}$ in a matrix form as $\mathbf{r}^1 = \tilde{\mathbf{H}}_1\mathbf{x} + \eta_1$. The equivalent channel matrix from transmitter to the first receive antennas could be represented as $\tilde{\mathbf{H}}_1 = \begin{bmatrix} \mathbf{H}_{11} & \mathbf{H}_{21} \\ \mathbf{H}_{21}^* & -\mathbf{H}_{11}^* \end{bmatrix}$. In a similar method, we could describe the output at the second receive antennas as $\mathbf{r}^2 = \tilde{\mathbf{H}}_2\mathbf{x} + \eta_2$. Here, the overall receive vector is represented by combining $\mathbf{r}^1$ and $\mathbf{r}^2$ as

$$\mathbf{r} = \begin{bmatrix} \mathbf{r}^1 \\ \mathbf{r}^2 \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{H}}_1 \\ \tilde{\mathbf{H}}_2 \end{bmatrix} .\mathbf{x} + \begin{bmatrix} \eta_1 \\ \eta_2 \end{bmatrix} = \tilde{\mathbf{H}}.\mathbf{x} + \eta \tag{1.2}$$

where $\tilde{\mathbf{H}}$ is the overall equivalent channel matrix between transmitter and receiver.

2. Iterative Estimation and Channel Decoding Algorithm

Through utilization of the overall channel matrix $\tilde{\mathbf{H}}$, the transmitted signal can be estimated by maximizing the posteriori probability in iterative processing. The first step is to estimate the frequency domain samples through linear MMSE equalizer. The noise in (1.2) is assumed uncorrelated and zero mean, We derive the MMSE equalizer $\mathbf{w}_n$ through minimizing the cost function $J(\mathbf{w}_n) = E\{|x(n) - \mathbf{w}_n^H \mathbf{r}|^2\}$. Hence, we obtain the MMSE equalizer coefficient vector $\mathbf{w}_n$ and the estimated value $\hat{x}(n)$s as

$$\mathbf{w}_n = (\tilde{\mathbf{H}} Cov(\mathbf{x},\mathbf{x}) \tilde{\mathbf{H}}^H + \sigma_x^2 \mathbf{I}_n)^{-1} \tilde{\mathbf{H}} Cov[\mathbf{x}, x(n)] \tag{2.1}$$

$$\hat{x}(n) = \overline{x}(n) + \mathbf{w}_n^H (\mathbf{r} - \tilde{\mathbf{H}}\overline{\mathbf{x}}) \tag{2.2}$$

Here, $\overline{x}(n) = E\{x(n)\}$ and $\overline{\mathbf{x}} = E\{\mathbf{x}\}$, where assuming that $\{\overline{x}(n) \neq 0\}$. By (2.1) and (2.2), we estimate the values of $\{\hat{x}(n)\}$ at first. Next, we need update new estimates of $\{\overline{x}(n)\}$ and $\{Cov[x(n), x(n)]\}$ based on $\hat{x}(n)$s. Here, the *extrinsic information* is utilized to find the posterior values for updating $\{\overline{x}(n)\}$ and $\{Cov[x(n), x(n)]\}$ into $\overline{\mathbf{x}}$ and $Cov[\mathbf{x}, \mathbf{x}]$ (from (2.1) and (2.2)). We set initialization by $\forall \overline{x}(n) = 0$ and $\forall Cov[x(n), x(n)] = 1$. With utilization of BPSK signals, the iterative updating processing could work through finding the log-likelihood ratios (LLR)s from $\{\hat{x}(n)\}$. Hence, We could define the difference between the posterior and prior LLRs of $x(n)$ as

$$\Delta L[x(n)] = L[x(n)|_{\hat{x}(n)}] - L[x(n)] = \ln \frac{\Pr\{x(n) = 1|_{\hat{x}(n)}\}}{\Pr\{x(n) = -1|_{\hat{x}(n)}\}} - \ln \frac{\Pr\{x(n) = 1\}}{\Pr\{x(n) = -1\}} = \ln \frac{\Pr\{\hat{x}(n)|_{x(n)=1}\}}{\Pr\{\hat{x}(n)|_{x(n)=-1}\}} \tag{2.3}$$

As the signal $x(n) = b\ \varepsilon\ \{+1, -1\}$, the conditional probability density function (PDF) of $x(n)$ is expressed as $\Pr\{\hat{x}(n)|_{x(n)=b}\} \approx \exp\left(-\frac{(\hat{x}(n) - m_n(b))(\hat{x}(n) - m_n(b))^H}{\sigma_x^2|_{x(n)=b}}\right)$, where the posterior conditional mean and covariance value of $\hat{x}(n)$ could be defined, respectively, as $m_n(b) = E\{\hat{x}(n)|_{x(n)=b}\} = \mathbf{w}_n^H \tilde{\mathbf{h}}_n b$ and $\sigma_x^2|_{x(n)=b} = Cov[\hat{x}(n), \hat{x}(n)|_{x(n)=b}] = \mathbf{w}_n^H \tilde{\mathbf{h}}_n - \mathbf{w}_n^H \tilde{\mathbf{h}}_n \tilde{\mathbf{h}}_n^H \mathbf{w}_n$, where $\tilde{\mathbf{h}}_n$ is the *nth* column of $\tilde{\mathbf{H}}$. Because of initialization, it yields $L[x(n)] = 0$ at the beginning of iteration. Hence, we could obtain $\Delta L[x(n)] = \ln \frac{\left[\exp\left(-\frac{(\hat{x}(n) - m_n(+1))^2}{\sigma_x^2|_{x(n)=+1}}\right)\right]}{\left[\exp\left(-\frac{(\hat{x}(n) - m_n(-1))^2}{\sigma_x^2|_{x(n)=-1}}\right)\right]} = \frac{4Re\{\hat{x}(n)\}}{1 - \tilde{\mathbf{h}}_n^H \mathbf{w}_n}$. Once the LLRs are obtained, the posterior values can be represented as $\overline{x}(n)_{new} = \tanh\left(\frac{L[x(n)|_{\hat{x}(n)}]}{2}\right)$ and $Cov[x(n), x(n)]_{new} = 1 - \overline{x}(n)_{new}^2$ (see in Schniter (2004)). These posterior values will be updated into (2.1) and (2.2) to find estimate values. Finally, these values are passed to the channel decoders. We use SISO convolutional decoder using Viterbi decoding. From the SISO decoder, the new error corrected estimations $\hat{\hat{\mathbf{x}}}$ will be passed to the iterative MMSE multiuser, in turn, until the specified number of iterations has elapsed.

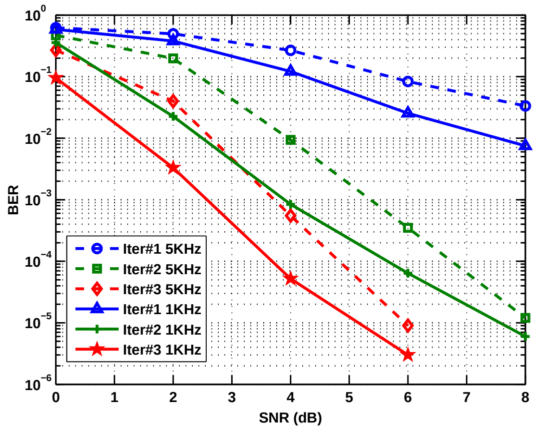

FIGURE 2. Performance comparison for iterative scheme when fading rate is 1KHz and 5kHz.

3. Numerical Results and Conclusions

With assumption of 1MHz transmitting bandwidth, which is divided into 32 sub-carriers, we exploit that each serial user data stream contains 64 symbols for two consecutive time slots. The simulation is implemented over a MIMO slow fading channel, which each sub-channel have 3 paths, generated by using the typical Jakes fading model and assumed to be uncorrelated among different transmitting antennas. Fig. 2 presents the bit error rate performance comparison of the proposed iterative receiver for different fading rate channel environment (maximum Doppler frequency $fd = 1kHz$ and $fd = 5kHz$). By comparison, we observe significant improvement of system performance by iterative processing over slow fading channel environment. Even though the performance degrades at higher SNR values with maximum Doppler frequency rising, the scheme still achieve benefit for performance within iterative processing. The simulation results indicate that the proposed scheme could obtain substantial performance improvement over multiuser slow fading channel environment only after three iterations.

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