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A Space-Time Coded MIMO-OFDM Multiuser Application With Iterative MMSE-Decision Feedback Algorithm

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***Abstract*— A popular technique for high data rate wireless transmission is OFDM. To increase the diversity gain and/or to enhance the system capacity, in practice, OFDM may be used in combination with antenna arrays at the transmitter and receiver to form a multiple-input multiple-output (MIMO) system. In this paper, an iterative multiuser receiver structure for space-time block coded MIMO OFDM scheme is exploited over slow fading channels. We utilize iterative detection based on minimum mean square error updated with decision feedback (MMSE-DF). The symbols are estimated in iterative process by updating extrinsic information to develop log-likelihood ratios (LLRs). Simulation results indicate that the scheme is proposed for the considered scenario and the performance over classical linear MMSE estimator.**

I. Introduction

Recently, there has been a great and rapid growth in wireless communications and the demand for data rate of wireless transmission is becoming higher and higher, however the quality of wireless service is limited by the bandwidth and transmitting power. OFDM is a very popular and robust technique to mitigate the effect of delay spread for achieving higher spectral efficiency in high data rate wireless transmission. In practice, OFDM could be combined with a MIMO system to improve the performance of communication in terms of taking advantage of the spatial diversity, which is obtained by spatially multiple antennas in a multi-path scattering environment.

However, the performance of such MIMO OFDM systems may seriously degrade in the presence of multiple-access interference (MAI) due to the application of multiuser space-time coded systems. Moreover, the future wireless communication systems are considered that the sensitivity to both physical movement and channel variation will grow with increasing frequency bands [6], hence, we should also not ignore the influence upon quality of wireless transmission due to the presence of time-variant channel in future wireless applications. Therefore, it is of importance to design a system receiver including a suitable multiuser detection and channel equalizer.

In this paper, we extend the receiver structure of OFDM system in [3] for multiuser detection with iterative method over slow fading channels. A MIMO OFDM system with two users is implemented in this work, which utilize space-time block codes (STBC) (see [1] and [2]) and iterative MMSE multiuser detection by updating extrinsic information to develop log-likelihood ratios (LLRs), whereas the proposed scheme exploits four transmit antennas and two receive antennas to suppress the interference and decode signals. All channel knowledge is assumed known perfectly throughout.

The outline of the rest of this paper is organized as follows. In Section II, system model is given. The proposed iterative receiver scheme is described in Section III. Simulation results are discussed in Section IV, followed by our final conclusions in Section V.

Notation: Bold upper case $\mathbf{X}$ denotes a matrix and lowercase $\mathbf{x}$ denotes a vector. $\mathbf{X}^{(i)}$ and $\mathbf{x}^{(i)}$ denote the signal matrix and vector corresponding to the ith user. We use $x(k)$ to denote the kth element of the vector $\mathbf{x}$ of size N, where $k = 0,1,\dots N-1$. $\mathbf{x}_n$ denotes the nth block vector in the data stream. The matrix indexed by q and j is denoted by $\mathbf{X}_{qj}$. $\mathbf{I}_N$ is an identity matrix of size N. Complex conjugation, transposition and conjugate transposition of a matrix are respectively denoted by $(\cdot)^*$, $(\cdot)^T$ and $(\cdot)^H$.

II. System Model

Here, we exploit a multi-user MIMO OFDM uplink wireless communication system for k users. From the ith user, we suppose that a serial to parallel convertor collects a set of N bits BPSK in frequency domain $\mathbf{x}^{(i)} = [x^{(i)}(0),\cdots,x^{(i)}(N-1)]^T$ to form frequency domain signal symbols, and in practice, we will perform explicit OFDM modulating operation for each sub-channel. To avoid inter symbol interference (ISI), it have to choose sufficient cyclic prefix, which the guard interval P satisfy $P > L$, where L is the length of impulse response of each transmit-receive antennas sub-channel. Hence, the N time domain received symbols in terms of vector form can be written as

$$\mathbf{r}_t = \mathbf{H}_c^{(i)}\mathbf{s}^{(i)} + \mathbf{v}^{(i)} = \mathbf{H}_c^{(i)}\mathbf{F}^H\mathbf{x}^{(i)} + \mathbf{v}^{(i)} \quad (1)$$

At the each of receive antenna, the samples corresponding to the cyclic prefix are first removed and then taking the FFT operation of received signal from (1) yields

$$\mathbf{r} = \mathbf{F}\mathbf{r}_t = \mathbf{F}\mathbf{H}_c^{(i)}\mathbf{F}^H\mathbf{x}^{(i)} + \mathbf{F}\mathbf{v}^{(i)} = \mathbf{H}^{(i)}\mathbf{x}^{(i)} + \eta^{(i)} \quad (2)$$

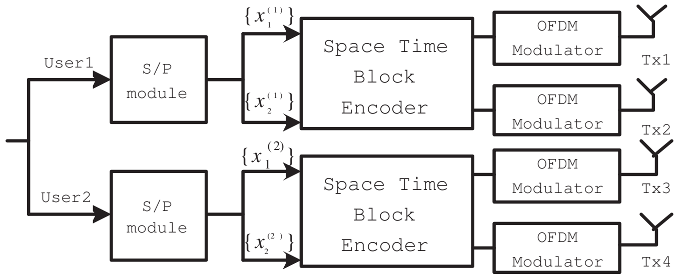

Fig. 1. Two user STBC MIMO-OFDM transmitter.

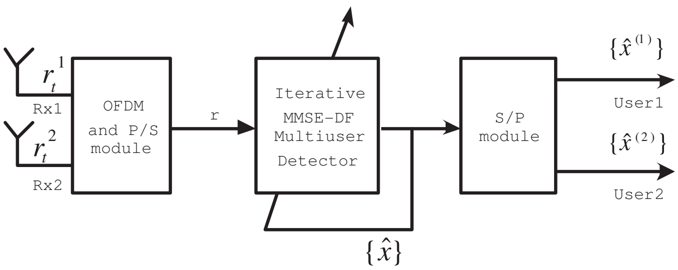

Fig. 2. Two user iterative MMSE-DF receiver

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 (3). (see also in [6]). Moreover, F denotes the $N \times N$ DFT matrix and $\mathbf{H}^{(i)}$ is defined as the *subcarrier coupling matrix*.

In this work, we adapt STBC scheme which is similar to the scheme in [2] for each MIMO OFDM system terminal. The two consecutive block signal vectors from ith user can be represented as $\mathbf{x}_1^{(i)} = [x_1^{(i)}(0), \cdots, x_1^{(i)}(N-1)]^T$ and $\mathbf{x}_2^{(i)} = [x_2^{(i)}(0), \cdots, x_2^{(i)}(N-1)]^T$. Here, we could assume that the time domain channel responses are constant during two consecutive signal block intervals, i.e. quasi-static. Hence, $[\mathbf{H}^{(i)}]$ is experienced by 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 [7]). The estimation of time invariant channels is discussed in [5]. The coded signal from two transmit antennas of ith user's terminal during two block intervals in frequency domain could be represented in matrix form as follows,

$$\begin{bmatrix} \mathbf{F}^H \mathbf{x}_1^{(i)} & \mathbf{F}^H \mathbf{x}_2^{(i)} \\ -\mathbf{F}^H \mathbf{x}_2^{(i)*} & \mathbf{F}^H \mathbf{x}_1^{(i)*} \end{bmatrix} \tag{4}$$

We now consider a MIMO-OFDM transmitter for two user terminals, which is shown in Fig 1, (see in [3] and [7]). Each user terminal is equipped with two transmit antennas for exploiting the above STBC scheme for wireless data transmission, and therefore, four antennas are required at transmitter. Two receive antennas are employed at receiver, which is shown in Fig 2.

If we consider the output of the OFDM demodulator at the first receive antenna after cyclic prefix removal, the signal vectors during the two sequential OFDM time-slots a and b could be obtained in (5) and (6), respectively.

$$\mathbf{r}_a^1 = \mathbf{H}_{11} \begin{bmatrix} \mathbf{x}_1^{(1)} \\ \mathbf{x}_1^{(2)} \end{bmatrix} + \mathbf{H}_{21} \begin{bmatrix} \mathbf{x}_2^{(1)} \\ \mathbf{x}_2^{(2)} \end{bmatrix} + \eta_{1a} \tag{5}$$

$$\mathbf{r}_b^{1*} = \mathbf{H}_{21}^* \begin{bmatrix} \mathbf{x}_1^{(1)} \\ \mathbf{x}_1^{(2)} \end{bmatrix} - \mathbf{H}_{11}^* \begin{bmatrix} \mathbf{x}_2^{(1)} \\ \mathbf{x}_2^{(2)} \end{bmatrix} + \eta_{1b}^* \tag{6}$$

where $\mathbf{H}_{qj} \triangleq [\mathbf{H}_{qj}^{(1)}, \mathbf{H}_{qj}^{(2)}]$. 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 (5) and (6) in a matrix form as

$$\mathbf{r}^1 = \tilde{\mathbf{H}}_1 \mathbf{x} + \eta_1 \tag{7}$$

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 arrange the output at the second receive antennas as

$$\mathbf{r}^2 = \tilde{\mathbf{H}}_2 \mathbf{x} + \eta_2 \tag{8}$$

where the equivalent channel matrix $\tilde{\mathbf{H}}_2$ to the second receive antennas could be represented as $\tilde{\mathbf{H}}_2 = \begin{bmatrix} \mathbf{H}_{12} & \mathbf{H}_{22} \\ \mathbf{H}_{22}^* & -\mathbf{H}_{12}^* \end{bmatrix}$. Here, the overall receive vector is arranged by combining (7) and (8) as

$$\mathbf{r} = \begin{bmatrix} \mathbf{r}^1 \\ \mathbf{r}^2 \end{bmatrix} = \begin{bmatrix} \tilde{\mathbf{H}}_1 \\ \tilde{\mathbf{H}}_2 \end{bmatrix} .\mathbf{x} + \begin{bmatrix} \eta_1 \\ \eta_2 \end{bmatrix} = \tilde{\mathbf{H}}.\mathbf{x} + \eta \tag{9}$$

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

III. Iterative MMSE Estimation on Multiuser Receiver

We now consider an iterative MMSE multiuser receiver for joint multiuser detection as shown in Fig 2 (see also in [6]). Through utilization of the overall equivalent time domain channel matrix $\tilde{\mathbf{H}}$ in (9), we could obtain estimation of the transmitted symbol $x(n)$. The direct estimation based on linear MMSE, however, could not offer good enough performance due to presence MAI and channel variation. In this work, the transmitted signal can be estimated in an iterative detection process. We structure a MMSE equalizer to get estimated $x(n)$s at first, and then those estimated values could be used to maximize the posteriori probability in iterative processing.

The first step is to estimate the frequency domain samples through linear MMSE equalizer. The noise in (9) is assumed uncorrelated and zero mean, therefore $E\{\eta\} = 0$, $E\{\eta\eta^H\} = \sigma_n^2 \mathbf{I}_n$ and $E\{x(n)\eta\} = 0$, We could derive the MMSE equalizer $\mathbf{w}_n$ through minimizing the following cost function

$$J(\mathbf{w}_n) = E\{\left|x(n) - \mathbf{w}_n^H \mathbf{r}\right|^2\}$$

$$
\mathbf{H}_c = \begin{bmatrix}
h_0^0 & 0 & \cdots & & 0 & & & h_0^{N_h-1} & h_0^{N_h-2} & \cdots & h_0^1 \\
h_1^1 & h_1^0 & 0 & \cdots & & 0 & & 0 & h_1^{N_h-1} & \cdots & h_1^2 \\
\vdots & & \ddots & & & \ddots & & & \ddots & & \vdots \\
h_{N_h-2}^{N_h-2} & \cdots & h_{N_h-2}^0 & 0 & \cdots & 0 & & \cdots & & 0 & h_{N_h-2}^{N_h-1} \\
h_{N_h-1}^{N_h-1} & h_{N_h-1}^{N_h-2} & \cdots & h_{N_h-1}^0 & 0 & \cdots & 0 & & \cdots & & 0 \\
0 & h_{N_h}^{N_h-1} & h_{N_h}^{N_h-2} & \cdots & h_{N_h}^0 & 0 & \cdots & 0 & & \cdots & 0 \\
\vdots & & \ddots & & & \ddots & & & \ddots & & \vdots \\
0 & \cdots & & \cdots & & 0 & h_{N-2}^{N_h-1} & h_{N-2}^{N_h-2} & \cdots & h_{N-2}^0 & 0 \\
0 & 0 & \cdots & & \cdots & & 0 & h_{N-1}^{N_h-1} & h_{N-1}^{N_h-2} & \cdots & h_{N-1}^0
\end{bmatrix}_{N\times N} \qquad (3)
$$

which obtains the MMSE equalizer coefficient vector

$$\mathbf{w}_n = (\tilde{\mathbf{H}}Cov(\mathbf{x},\mathbf{x})\tilde{\mathbf{H}}^H + \sigma_x^2\mathbf{I}_n)^{-1}\tilde{\mathbf{H}}Cov[\mathbf{x},x(n)] \qquad (10)$$

and then the estimated value $\hat{x}(n)$s could be obtained as

$$\hat{x}(n) = \bar{x}(n) + \mathbf{w}_n^H(\mathbf{r} - \tilde{\mathbf{H}}\bar{\mathbf{x}}) \qquad (11)$$

Here, $\bar{x}(n) = E\{x(n)\}$ and $\bar{\mathbf{x}} = E\{\mathbf{x}\}$, where assuming that $\{\bar{x}(n) \neq 0\}$. By (10) and (11), we estimate the values of $\{\hat{x}(n)\}$ at the first step. In order to apply an iterative algorithm, next, we need obtain new values for $\{\bar{x}(n)\}$ and $\{Cov[x(n),x(n)]\}$ based on the estimates $\hat{x}(n)$s. With these values, we choose to use only *extrinsic information* and wish to find the posterior values to update $\{\bar{x}(n)\}$ and $\{Cov[x(n),x(n)]\}$ into $\bar{\mathbf{x}}$ and $Cov[\mathbf{x},\mathbf{x}]$, which is shown in (10) and (11). Here, a block based iterative estimation approach is chosen. We set initialization by $\forall\bar{x}(n) = 0$ and $\forall Cov[x(n),x(n)] = 1$. With utilization of BPSK signals, the updating processing of iterative algorithm could work through finding the log-likelihood ratios (LLR)s from the estimated values of $\{\hat{x}(n)\}$. We could define the difference between the posterior and prior LLRs of $x(n)$ is

$$L[x(n)] = \ln\frac{\Pr\{x(n)=1\}}{\Pr\{x(n)=-1\}} \qquad (12)$$

$$L[x(n)|_{\hat{x}(n)}] = \ln\frac{\Pr\{x(n)=1|_{\hat{x}(n)}\}}{\Pr\{x(n)=-1|_{\hat{x}(n)}\}} \qquad (13)$$

$$\begin{aligned}\Delta L[x(n)] &= L[x(n)|_{\hat{x}(n)}] - L[x(n)] \\ &= \ln\frac{\Pr\{\hat{x}(n)|_{x(n)=1}\}}{\Pr\{\hat{x}(n)|_{x(n)=-1}\}}\end{aligned} \qquad (14)$$

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 as $m_n(b) = E\{\hat{x}(n)|_{x(n)=b}\}$ and $\sigma_x^2|_{x(n)=b} = Cov[\hat{x}(n),\hat{x}(n)|_{x(n)=b}]$, respectively. They could be determined by (10) and (11) as

$$E\{\hat{x}(n)|_{x(n)=b}\} = \mathbf{w}_n^H\tilde{\mathbf{h}}_n b \qquad (15)$$

$$\begin{aligned}\sigma_x^2|_{x(n)=b} &= E\{\hat{x}(n)\hat{x}^H(n)|_{x(n)=b}\} - m_n(b)m_n(b)^H \\ &= \mathbf{w}_n^H\tilde{\mathbf{h}}_n - \mathbf{w}_n^H\tilde{\mathbf{h}}_n\tilde{\mathbf{h}}_n^H\mathbf{w}_n\end{aligned} \qquad (16)$$

where $\tilde{\mathbf{h}}_n$ is the *nth* column of $\tilde{\mathbf{H}}$. Moreover, because we set $\bar{x}(n) = 0$ and $Cov[x(n),x(n)] = 1$ initially for finding the posterior LLR of $x(n)$ by using only *extrinsic* information, therefore, yields $L[x(n)] = 0$ at the beginning of iteration. Now, we could obtain

$$\begin{aligned}\Delta L[x(n)] &= L[x(n)|_{\hat{x}(n)}] - L[x(n)] \\ &= \ln\frac{\left[\exp\left(-\frac{(\hat{x}(n)-m_n(+1))^2}{\sigma_x^2|_{x(n)=+1}}\right)\right]}{\left[\exp\left(-\frac{(\hat{x}(n)-m_n(-1))^2}{\sigma_x^2|_{x(n)=-1}}\right)\right]} \\ &= \frac{4Re\{\hat{x}(n)\}}{1-\tilde{\mathbf{h}}_n^H\mathbf{w}_n}\end{aligned} \qquad (17)$$

Once the LLRs are obtained, the posterior values for $\bar{x}(n)$ and $Cov[x(n),x(n)]$ can be updated as

$$\begin{aligned}\bar{x}(n)_{new} &= \Pr\{x(n)=+1|_{\hat{x}(n)}\} - \Pr\{x(n)=-1|_{\hat{x}(n)}\} \\ &= \tanh\left(\frac{L[x(n)|_{\hat{x}(n)}]}{2}\right)\end{aligned} \qquad (18)$$

$$\begin{aligned}&Cov[x(n),x(n)]_{new} \\ &= \sum_{b\varepsilon\{+1,-1\}} (b - E\{x(n)|_{\hat{x}(n)}\} \cdot \Pr\{x(n)=b|_{\hat{x}(n)}\} \\ &= 1 - \bar{x}(n)^2_{new}\end{aligned} \qquad (19)$$

Then, the equalizer (11) receive soft decision feedback as the posterior values $\bar{x}(n)_{new}$ obtained by (18), and we also can obtain $Cov[x(n),x(n)]_{new}$ by (19) to update relative value in diagonal matrix $Cov(\mathbf{x},\mathbf{x})$ from (10). These updating coefficient is used to estimate new value $x(n+1)$ by repeating the step (10) though (19). The iterative updating process does not stop until the specified number of iterations has elapsed.

IV. SIMULATION

To illustrate the performance of the proposed iterative receiver structure, a two user STB coded MIMO OFDM system case using four transmit and two receive antennas is simulated in this section. With assumption of 1MHz transmitting bandwidth,i.e. the OFDM symbol duration is $1\mu s$, which is divided into 64 sub-carriers by OFDM operation, we exploit a BPSK signal constellation in this work for high speed computation with log-likelihood ratios. Each serial user data stream contains 128 symbols, which is coded into two data block by STBC operation. Therefore, two user terminals could transmit two stream in parallel, which are 256 bit information symbols in one transmitting signal stream totally.

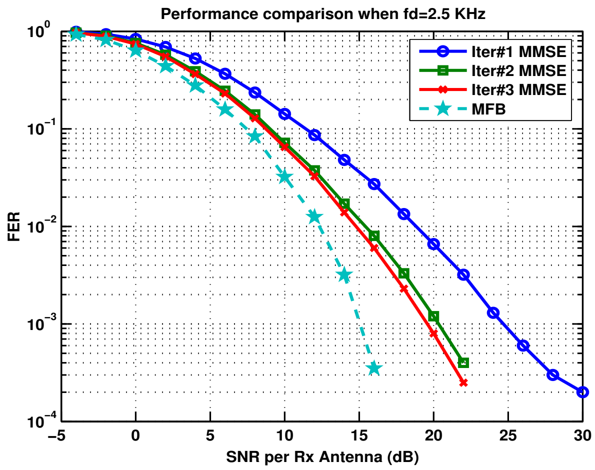

Fig. 3. Performance comparison with different number of iteration when maximum Doppler frequency is 2.5 KHz.

The simulation of data transmitting is implemented over MIMO slow fading channels, generated by using the typical Jakes fading model (see [4]). The channel impulse responses of each transmit-receive antennas of each user have 3 taps. Here, we assume $\sum_{l=0}^{L-1}\sigma_l^2 = 1$, where σ_l^2 is the variance of the lth path, and the channel fading is assumed to be uncorrelated among different transmitting antennas of different users. We assume perfect knowledge of the channel state at the receiver at any time.

Fig. 3 presents the frame error rate performance of the proposed iterative receiver over time-variant slow fading channel with maximum Doppler frequency $fd = 2.5kHz$. As a benchmark, we also evaluate performance considering the match filter bound (MFB) obtained from the model given in (9) by assuming the symbols $\{x(u)|_{u\neq n}\}$ are perfectly known. By comparison with [3], we observe significant improvement of system performance by iterative processing over slow fading channel environment after three iterations.

Fig. 4 shows the comparison of system performance over different fading rate channel environment when the receiver has three iteration process. The best performance is given in quasi-static channel, and the performance degrades at higher SNR values with maximum Doppler frequency rising. However, we could still further enhance performance by concatenating coding technique into iterative processing, for example, turbo coding.

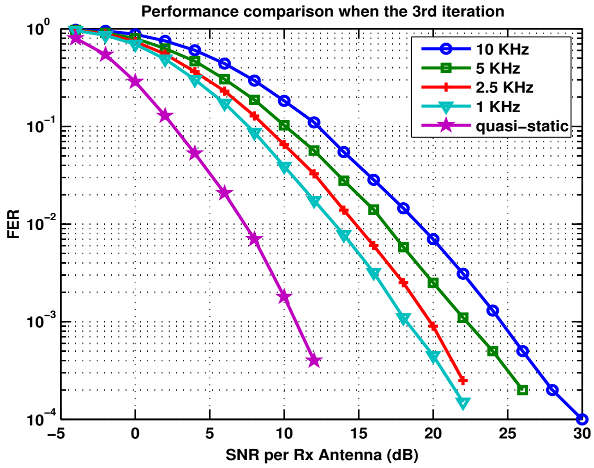

Fig. 4. Performance comparison with different fading rate when the receiver is at the 3rd iteration.

V. CONCLUSIONS

In this paper, we address the design of a iterative receiver for a MIMO OFDM wireless communication system over doubly-selective slow fading channels. The receiver is based on the minimum mean square error iterative algorithm by updating with extrinsic information. We exploit the structure of STBC techniques at transmitter, concatenated with OFDM operation to suppress interference. The simulation results indicate that the proposed scheme could obtain substantial performance improvement over slow fading channel environment. In future work, we could be able to develop this structure by introducing channel coding technology into feedback process.

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