Efficient Low Complexity Turbo Equalization with soft interference cancellation in MIMO system

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Abstract—**An efficient frequency domain soft interference cancellation (FD-SIC) scheme is proposed for turbo frequency domain equalization (FDE) in multiple-input multiple-output (MIMO) system. It is designed that at the receiver the inter-antenna interference (IAI) is cancelled by prior information in terms of data expectation. It can improve the input signal to interference and noise ratio (SINR) for the equalizer. Then block (data of one transmit antenna) wise frequency domain minimum mean square error (MMSE) equalization is implemented. Later a low complexity algorithm on the basis of single frequency tone (SFT) is deduced. Symbol Log-likelihood ratio (LLR) is calculated from the output of equalizer. Simulation results show that the proposed scheme has better link performance than traditional algorithms.**

Keywords-FD-SIC; MIMO; block wise Turbo FDE; SFT

I. INTRODUCTION

Nowadays, iteration theory is introduced into equalization as in [3]. In Turbo equalization, equalization and decoding have been jointly optimized by interchanging extrinsic information with each other. A detailed description of Turbo equalization is given in [4]. It discusses the Turbo equalization approach to coded data transmission over ISI channels, with an emphasis of the basic ideas and some of the practical details. The differences between optimum and suboptimum soft–in/soft–out algorithms for equalization and decoding is described in [5], in which the comparison of them in a turbo-detection scheme concerning complexity and performance for perfect and mismatched channel estimation is presented. A novel iterative receiver algorithm is proposed in [6], which designed for coded data transmission over multipath channels. The ISI is removed with a soft-in soft-out (SISO) equalizer based on linear filtering. Frequency domain Turbo equalization was introduced in [7], which derived turbo equalization both in time and frequency domain. A number of low-complexity SISO equalization algorithms based on the MMSE criterion is explored in [8]. MIMO was imported in [9], in which a new class of block turbo equalizers for single-carrier transmission over MIMO broadband wireless channels is proposed.

In this paper, an efficient frequency domain block wise Turbo MMSE equalization with frequency domain soft interference cancellation in broadband MIMO system is proposed. Obvious distinction with traditional equalization is that equalizer in Turbo equalization can get prior information from decoder. So the useful prior information can be utilized to eliminate the interference from other antennas and symbols. After interferences from other antennas are cancelled by prior information, block wise frequency domain MMSE equalization is executed. So the SINR for each antenna data has been improved compared with original received data. As in [8], the output of the MMSE filter is approximated that it undergone an equivalent Gaussian channel. As a result, equalization has to be executed several times to obtain all transmit data. In order to decrease complexity, equalization on single frequency tone which can reduce calculation complexity in high degree is proposed. This paper is organized as follows. In section II, system model is represented. Section III describes frequency domain Turbo MMSE equalization scheme. Section IV gives complexity comparison and simulation results. The conclusions are drawn in Section V.

II. SYSTEM MODEL

The system has n_T transmit antennas and n_R receive antennas, $n_R \ge n_T$, as illustrated in Fig.1. The data symbols are i.i.d. complex, zero-mean, uncorrelated and with energy σ_s^2. Data symbols are divided into n_T sub streams. Then convolution encode is adopted to combat the AWGN. Here MPSK is the modulation mode, and the modulated data symbols belong to the M possible symbols, $\{s_i \in S \mid i. =1,2,\cdots,M\}$. The frequency selective channel is assumed to be quasi-static in a block, which means the channel impulse will not vary in a block. Channel memory length is supposed to be L, and accurate channel information can be gotten at the receiver. From Fig.1, we can see that the original data can be expressed as $\mathbf{a}=[\mathbf{a}_1,\mathbf{a}_2,\cdots,\mathbf{a}_{n_T}]$, where $\mathbf{a}_i=[a_{i,1},a_{i,2},\cdots,a_{i,l_{1,i}}]$ $(i=1,2, \cdots,n_T)$. Here $l_{1,i}$ is the length of original data block on i^{th} antenna. Then the data are coded and interleaved, the result can be expressed respectively as $\mathbf{b}_i=[b_{i,1},b_{i,2}, \cdots,b_{i,l_{2,i}}]$ and $\mathbf{c}_i=[c_{i,1},c_{i,2}, \cdots,c_{i,l_{2,i}}]$. $l_{2,i}$ is the length of coded data. Different mapping and coding schemes can be adopted on different antennas on condition

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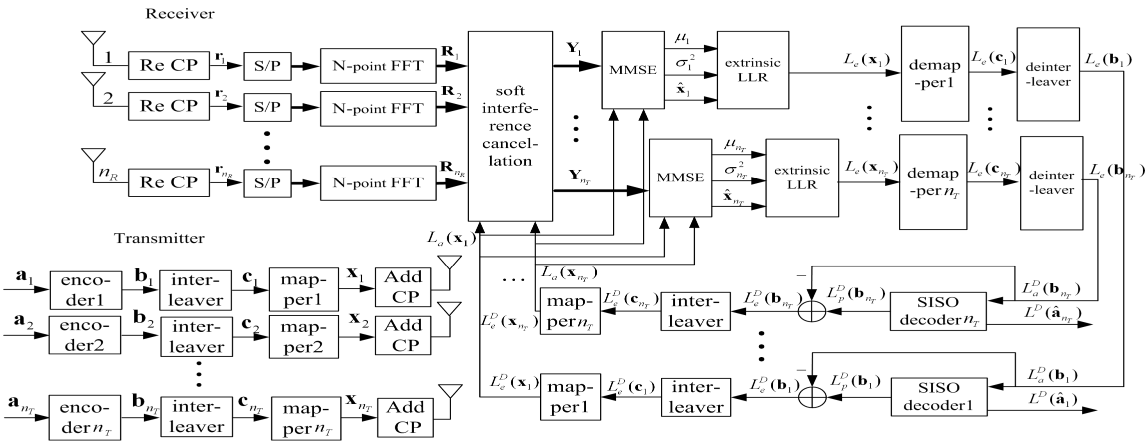

Figure 1 System model

that the size of modulated block is N. After mapping, data on i^{th} antenna can be expressed as $\mathbf{x}_i = [x_{i,1}, x_{i,2}, \cdots, x_{i,N}]^T$. At the receiver after CP is discarded, the received data on j^{th} receiving antenna can be expressed as $\mathbf{r}_j = [r_{j,1}, r_{j,2}, \cdots, r_{j,N}]^T$. System model in time domain can be described as follows,

$$\mathbf{r}_j = \sum_{i=1}^{n_T} \mathbf{h}_{j,i}\mathbf{x}_i + \mathbf{n}_j \tag{1}$$

where,

$$\mathbf{h}_{j,i} = \begin{bmatrix} h_{j,i}^1 & 0 & \cdots & \cdots & h_{j,i}^3 & h_{j,i}^2 \\ h_{j,i}^2 & h_{j,i}^1 & & & h_{j,i}^4 & h_{j,i}^3 \\ \vdots & h_{j,i}^2 & \ddots & & \vdots & \vdots \\ h_{j,i}^L & \vdots & & & h_{j,i}^{L-1} & h_{j,i}^L \\ 0 & h_{j,i}^L & & \ddots & h_{j,i}^L & 0 \\ \vdots & \vdots & & & \vdots & \vdots \\ 0 & 0 & \cdots & \cdots & h_{j,i}^2 & h_{j,i}^1 \end{bmatrix}_{(N\times N)}$$

$\mathbf{n}_j = [n_{j,1}, n_{j,2}, \cdots, n_{j,N}]^T$ is additive noise on j^{th} receiving data, whose elements are zero-mean, circularly symmetric, complex Gaussian distributed with real and imaginary part variance of $0.5\sigma^2$. Let $\mathbf{x} = [\mathbf{x}_1^T, \mathbf{x}_2^T, \cdots, \mathbf{x}_{n_T}^T]^T$ denotes all transmit data, $\mathbf{r} = [\mathbf{r}_1^T, \mathbf{r}_2^T, \cdots, \mathbf{r}_{n_R}^T]^T$ denotes all received data. Then the reception model can be expressed as follows,

$$\mathbf{r} = \mathbf{h}\mathbf{x} + \mathbf{n} \tag{2}$$

where, $\mathbf{h} = \begin{bmatrix} \mathbf{h}_{1,1} & \mathbf{h}_{1,2} & \cdots & \mathbf{h}_{1,n_T} \\ \mathbf{h}_{2,1} & \mathbf{h}_{2,2} & & \mathbf{h}_{2,n_T} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{h}_{n_R,1} & \mathbf{h}_{n_R,2} & \cdots & \mathbf{h}_{n_R,n_T} \end{bmatrix}_{(Nn_R\times Nn_T)}$

$$\mathbf{n} = [\mathbf{n}_1^T, \mathbf{n}_2^T, \cdots, \mathbf{n}_{n_R}^T]^T$$

If the expression is transformed into frequency domain, it can be written as,

$$\begin{aligned} \mathbf{R} &= \mathbf{H}\tilde{\mathbf{F}}\mathbf{x} + \mathbf{N} \\ &= \mathbf{H}\mathbf{X} + \mathbf{N} \end{aligned} \tag{3}$$

where, $\mathbf{H} = \begin{bmatrix} \mathbf{H}_{1,1} & \mathbf{H}_{1,2} & \cdots & \mathbf{H}_{1,n_T} \\ \mathbf{H}_{2,1} & \mathbf{H}_{2,2} & & \mathbf{H}_{2,n_T} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{H}_{n_R,1} & \mathbf{H}_{n_R,2} & \cdots & \mathbf{H}_{n_R,n_T} \end{bmatrix}_{(Nn_R\times Nn_T)}$

$$\tilde{\mathbf{F}} = \begin{bmatrix} \mathbf{F} & \mathbf{0} & \cdots & \mathbf{0} \\ \mathbf{0} & \mathbf{F} & \cdots & \mathbf{0} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{0} & \mathbf{0} & \cdots & \mathbf{F} \end{bmatrix}_{Nn_R\times Nn_T} \quad \begin{aligned} F_{m,l} &= \frac{1}{\sqrt{N}} e^{-j\frac{2\pi}{N}(m-1)(l-1)} \\ m &= 1, \cdots, N \\ l &= 1, \cdots, N \end{aligned}$$

$\mathbf{H}_{j,i} (j = 1, 2, \cdots, n_R; i = 1, 2, \cdots, n_T)$ is a diagonal matrix, whose diagonal elements are the FFT result of the first column of $\mathbf{h}_{j,i}$.

III. ALGORITHM DESCRIPTION

A. Frequency Domain Soft Interference Cancellations

We can make some analysis on equation (3),

$$\mathbf{R} = \mathbf{H}\tilde{\mathbf{F}}\mathbf{x} + \mathbf{N} \tag{4}$$

Considering the character of FFT, we can see that each frequency domain data symbol $X_{i,n}$ $(i = 1, 2, \cdots, n_T)$ $(n = 1, 2, \cdots, N)$ is a linear combination of all N data symbols on i^{th} transmitted antenna in time domain. So it can get frequency diversity together with multi-antenna diversity in this broadband SC-FDE MIMO system. As in Turbo equalization, prior information is fed back from SISO decoder. This prior information can be efficiently utilized to eliminate the IAI. Data block $\mathbf{X}_i$ with a factor exists in every receiving antenna. So after IAI have been eliminated using prior information $L_a(\mathbf{c}_{\hat{i}})$, the received data $\mathbf{R}$ can be used to estimate $\mathbf{X}_i$ by block wise equalization. $L_a(\mathbf{c}_{\hat{i}})$ is the prior information of transmitted data from all antennas expect antenna i. According $L_a(\mathbf{c})$, the expectation of $\mathbf{x}$ can be expressed as follow [8],

$$\bar{x}_{i,n} = E\left(x_{i,n}\right) = \sum_{s_k \in S} P(x_{i,n} = s_k) s_k \quad (5)$$

$$i = 1, 2, \cdots, n_T; n = 1, 2, \cdots, N$$

$$P(x_{i,n} = s_k) = \prod_{j=1}^{M} P(c_{i,n}^j = s_{k,j})$$

$$= \prod_{j=1}^{M} \frac{1}{2}\left(1 + \tilde{s}_{k,j} \tanh(L_a(c_{i,n}^j / 2))\right)$$

$$s_{k,j} \in \{0,1\}$$

$$\tilde{s}_{k,j} \triangleq \begin{cases} +1 & s_{k,j} = 0 \\ -1 & s_{k,j} = 1 \end{cases}$$

where s_k is one mapping symbol in M, and $\log_2 M$ bits, $c_{i,n}^j$ are mapped as a symbol $x_{i,n}$ .As $\mathbf{X}$ is a linear combination of $\mathbf{x}$, expectation of $\mathbf{X}$ on the basis of $L_a(\mathbf{c})$ is,

$$\bar{\mathbf{X}} = \tilde{\mathbf{F}}\bar{\mathbf{x}} \quad (6)$$

The expected IAI for $\mathbf{X}_i$ can be written as $\boldsymbol{\alpha}_i$,

$$\boldsymbol{\alpha}_i = \mathbf{H}\ \bar{\mathbf{X}}_{\hat{i}} \quad (7)$$

Where, $\bar{\mathbf{X}}_{\hat{i}} = \left[\bar{\mathbf{X}}_1^T, \cdots, \bar{\mathbf{X}}_{i-1}^T, \mathbf{0}_{1\times N}^T, \bar{\mathbf{X}}_{i+1}^T \cdots, \bar{\mathbf{X}}_{n_T}^T\right]^T$

So the soft interference cancellation model can be expressed as follows,

$$\begin{aligned} \mathbf{Y}_i &= \mathbf{R} - \boldsymbol{\alpha}_i \\ &= \mathbf{H}\ (\mathbf{X} - \bar{\mathbf{X}}_{\hat{i}}) + \mathbf{N} \\ &= \mathbf{H}\mathbf{X}_i + \mathbf{H}(\mathbf{X}_{\hat{i}} - \bar{\mathbf{X}}_{\hat{i}}) + \mathbf{N} \end{aligned} \quad (8)$$

where, $\mathbf{X}_{\hat{i}} = \left[\mathbf{X}_1^T, \cdots, \mathbf{X}_{i-1}^T, \mathbf{0}_{1\times N}^T, \mathbf{X}_{i+1}^T \cdots, \mathbf{X}_{n_T}^T\right]^T$. In (8), the first term is the expected symbol multiplied with a factor; the second term is the residual interference from other antennas and symbols; the third term remains AWGN. Due to the iteration, the interference will be reduced as the prior information from SISO decoder is becoming accurate. Obviously SINR of vector $\mathbf{Y}_i$ has been improved compared with original received data from i^{th} transmit antenna. Then $\mathbf{Y}_i$ is sent to MMSE equalizer as its input. As a result, data on every transmit antenna are equalized on a higher SINR compared with the original received data. Then frequency domain equalization is implemented, while soft interference is ignored in iteration zero.

B. Traditional Turbo Equalization in MIMO System

Considering traditional Turbo equalization in MIMO system, estimation of data symbol $\mathbf{X}$ can be derived as,

$$\hat{\mathbf{X}} = \mathbf{C}^H \mathbf{R} + \mathbf{D} \quad (9)$$

$\mathbf{C}^H$ is the equalization coefficients, and $\mathbf{D}$ is a time-varying offset compensating for $\mathbf{X}$ on the basis of prior information. So $\mathbf{C}^H$ and $\mathbf{D}$ can be gotten with MMSE criterion,

$$\left(\mathbf{C}^H, \mathbf{D}\right) = \arg\min E\left\{\left|\hat{\mathbf{X}} - \mathbf{X}\right|^2\right\} \quad (10)$$

(10) equals the follow equations,

$$\frac{\partial\left(E\left\{\left|\hat{\mathbf{X}} - \mathbf{X}\right|^2\right\}\right)}{\partial\left(\mathbf{C}^H\right)} = \mathbf{0} \quad (11)$$

$$\frac{\partial\left(E\left\{\left|\hat{\mathbf{X}} - \mathbf{X}\right|^2\right\}\right)}{\partial\left(\mathbf{D}\right)} = \mathbf{0} \quad (12)$$

Expanding (11) and (12), we can get,

$$\mathbf{C}^H = \sigma_s^2 \mathbf{H}^H \left(\sigma_s^2 \mathbf{H}\mathbf{H}^H + \sigma^2 \mathbf{I}\right)^{-1} \quad (13)$$

$$\begin{aligned} &\mathbf{D}_i = \mu_i \bar{\mathbf{X}}_i - \mathbf{C}_i^H \mathbf{H}\bar{\mathbf{X}} \\ &\mathbf{D}((i-1)\times N : i\times N, 1) = \mathbf{D}_i \end{aligned} \quad (14)$$

where μ_i is the mean equivalent amplitude of the output signal, $\mu_i = \mathrm{sum}(\mathbf{C}^H\mathbf{H})_{(1:N\times n_T, (i-1)\times N+1:i\times N)} / N$ $(i = 1, 2, \cdots, n_t)$ (15)

From (9), the vector to estimated data is $\mathbf{R}$. As we analyzed previously the SINR of $\mathbf{Y} = [\mathbf{Y}_1^T, \mathbf{Y}_2^T, \cdots, \mathbf{Y}_{n_T}^T]^T$ is higher than $\mathbf{R}$. So our proposed SIC block wise Turbo equalization can efficiently use the prior information of all data symbol to get a better link performance theoretically.

C. Proposed SIC Turbo Equalization in MIMO System

Our MMSE equalization is executed for data on single transmit antenna. After SIC, prior information of all data symbol is utilized to estimate the data symbol on one transmit antenna. Estimation of data symbol $\mathbf{X}_i$ can be expressed as, $\hat{\mathbf{X}}_i = \mathbf{C}_i^H \mathbf{Y}_i + \mathbf{D}_i$ (16)

$\mathbf{C}_i^H$ and $\mathbf{D}_i$ can be gotten similarly with (11) as follows,

$$\left(\mathbf{C}_i^H, \mathbf{D}_i\right) = \arg\min E\left\{\left|\hat{\mathbf{X}}_i - \mathbf{X}_i\right|^2\right\} \quad (17)$$

$$\frac{\partial\left(E\left\{\left|\hat{\mathbf{X}}_i - \mathbf{X}_i\right|^2\right\}\right)}{\partial\left(\mathbf{C}_i^H\right)} = \mathbf{0} \quad (18)$$

$$\frac{\partial\left(E\left\{\left|\hat{\mathbf{X}} - \mathbf{X}\right|^2\right\}\right)}{\partial\left(\mathbf{D}\right)} = \mathbf{0} \quad (19)$$

Expanding (19) and (20),

$$\mathbf{C}_i^H = \sigma_s^2 \mathbf{U}_i \mathbf{H}^H \left(\sigma_s^2 \mathbf{H}\mathbf{H}^H + \sigma^2 \mathbf{I} + \boldsymbol{\alpha}_i \boldsymbol{\alpha}_i^H\right)^{-1} \quad (20)$$

$$\mathbf{D}_i = \left(\mu_i - \mathbf{C}_i^H \mathbf{H}_i\right)\bar{\mathbf{X}}_i + \mathbf{C}_i^H \boldsymbol{\alpha}_i \quad (21)$$

After all these operations, estimated data symbol can be gotten according to (17). As described in [7], expectation Λ_i^k for X_i^k can be obtained,

$$\Lambda_i^k = \sum_{m=1}^{n_R} C_{(i-1)\times N+k, (m-1)\times N+k} H_{(m-1)\times N+k, (i-1)\times N+k} \quad (22)$$

$$\mu_i = \sum_{m=1}^{N} \Lambda_i^k / N \quad (23)$$

D. *Equalization on the Basis of SFT*

According to (20), the calculation complexity is focusing on calculation the inverse of matrix $\left(\sigma_s^2\mathbf{H}\mathbf{H}^H+\sigma^2\mathbf{I}+\boldsymbol{\alpha}_i\boldsymbol{\alpha}_i^H\right)$. Considering the character of $\boldsymbol{\alpha}_i\boldsymbol{\alpha}_i^H$, when SINR is not too high, some approach can be executed as the elements on block diagonal make the primary effect. Construct a matrix to substitute $\boldsymbol{\alpha}_i\boldsymbol{\alpha}_i^H$,

$$\boldsymbol{\beta}=\begin{bmatrix}\boldsymbol{\beta}_{1,1} & \boldsymbol{\beta}_{1,2} & \cdots & \boldsymbol{\beta}_{1,n_T}\\ \boldsymbol{\beta}_{2,1} & \boldsymbol{\beta}_{2,2} & \cdots & \boldsymbol{\beta}_{2,n_T}\\ \vdots & \vdots & \ddots & \vdots\\ \boldsymbol{\beta}_{n_R,1} & \boldsymbol{\beta}_{n_R,2} & & \boldsymbol{\beta}_{n_R,n_T}\end{bmatrix}_{n_RN\times n_TN} \tag{24}$$

$$\beta_{m,n}^k=[\boldsymbol{\alpha}_i\boldsymbol{\alpha}_i^H]_{((m-1)N+k,(n-1)N+k)}$$
$$m=1,2,\ldots n_R;n=1,2,\ldots n_T;k=1,2,\ldots,N$$

where $\boldsymbol{\beta}_{m,n}$ is a block diagonal matrix. When $n_R=n_T$, some transformation can be executed on equation (20),

$$\begin{aligned}\mathbf{C}_i^H&=\sigma_s^2\mathbf{U}_i\mathbf{H}^H\left(\sigma_s^2\mathbf{H}\mathbf{H}^H+\sigma^2\mathbf{I}+\boldsymbol{\alpha}_i\boldsymbol{\alpha}_i^H\right)^{-1}\\&\approx\sigma_s^2\mathbf{U}_i\mathbf{H}^H\left(\sigma_s^2\mathbf{H}\mathbf{H}^H+\sigma^2\mathbf{I}+\boldsymbol{\beta}\right)^{-1}\\ \varphi=\frac{\sigma^2}{\sigma_s^2},\gamma=(\frac{\sigma^2}{\sigma_s^2}),&\ \sigma_s^4\mathbf{U}_i\mathbf{H}^H\left(\mathbf{H}\mathbf{H}^H+\varphi\gamma\right)^{-1}\\&=\sigma_s^4\mathbf{U}_i\mathbf{H}^H\mathbf{D}\end{aligned} \tag{25}$$

$$\mathbf{D}=\begin{bmatrix}\mathbf{D}_{1,1} & \mathbf{D}_{1,2} & \cdots & \mathbf{D}_{1,n_T}\\ \mathbf{D}_{2,1} & \mathbf{D}_{2,2} & \cdots & \mathbf{D}_{2,n_T}\\ \vdots & \vdots & \ddots & \vdots\\ \mathbf{D}_{n_R,1} & \mathbf{D}_{n_R,2} & \cdots & \mathbf{D}_{n_R,n_T}\end{bmatrix}_{n_RN\times n_TN} \tag{26}$$

As a result **D** is a block diagonal matrix too. It can be easily proved by mathematical deduced that,

$$\mathbf{D}^k=\left(\mathbf{H}^k\mathbf{H}^{k,H}+\varphi\gamma^k\right)^{-1} \tag{27}$$

$$\mathbf{H}^k=\begin{bmatrix}H_{1,1}^{k,k} & H_{1,2}^{k,k} & \cdots & H_{1,n_T}^{k,k}\\ H_{2,1}^{k,k} & H_{2,2}^{k,k} & \cdots & H_{2,n_T}^{k,k}\\ \vdots & \vdots & \ddots & \vdots\\ H_{n_R,1}^{k,k} & H_{n_R,2}^{k,k} & \cdots & H_{n_R,n_T}^{k,k}\end{bmatrix}_{n_R\times n_R}$$

$$\boldsymbol{\gamma}^k=\begin{bmatrix}\gamma_{1,1}^{k,k} & \gamma_{1,2}^{k,k} & \cdots & \gamma_{1,n_T}^{k,k}\\ \gamma_{2,1}^{k,k} & \gamma_{2,2}^{k,k} & \cdots & \gamma_{2,n_T}^{k,k}\\ \vdots & \vdots & \ddots & \vdots\\ \gamma_{n_R,1}^{k,k} & \gamma_{n_R,2}^{k,k} & \cdots & \gamma_{n_R,n_T}^{k,k}\end{bmatrix}_{n_R\times n_R}$$

$$\mathbf{D}^k=\begin{bmatrix}D_{1,1}^{k,k} & D_{1,2}^{k,k} & \cdots & D_{1,n_T}^{k,k}\\ D_{2,1}^{k,k} & D_{2,2}^{k,k} & \cdots & D_{2,n_T}^{k,k}\\ \vdots & \vdots & \ddots & \vdots\\ D_{n_R,1}^{k,k} & D_{n_R,2}^{k,k} & \cdots & D_{n_R,n_T}^{k,k}\end{bmatrix}_{n_R\times n_R}$$

In other words, the valve of matrix **D** on frequency tone k is only correlative with matrix $\left(\sigma_s^2\mathbf{H}\mathbf{H}^H+\sigma^2\mathbf{I}+\boldsymbol{\beta}\right)_s$ on the same frequency tone. Thus, the system model can be considered on single frequency tone, equalization coefficients $\mathbf{C}^{k,H}$ is expressed as,

$$\mathbf{C}^{k,H}=\sigma_s^2\mathbf{U}_i^k\mathbf{H}^{k,H}\left(\sigma_s^2\mathbf{H}^k\mathbf{H}^{k,H}+\sigma^2\mathbf{I}^k+\boldsymbol{\beta}^k\right)^{-1} \tag{28}$$

Mean equivalent amplitude μ_i can be obtained from **H** and $\mathbf{C}^H$ as,

$$\Lambda_i^k=\sum_{m=1}^{n_R}C^H_{(i-1)\times N+k,(m-1)\times N+k}H_{(m-1)\times N+k,(i-1)\times N+k}$$
(29)

The estimation of $\mathbf{X}^k$ can be obtained as following,

$$\hat{\mathbf{X}}^k=\mathbf{C}^{k,H}\mathbf{Y}^k+\mathbf{D}^k \tag{30}$$

$$\begin{aligned}&\mathbf{D}_i^k=\mu_i\bar{\mathbf{X}}_i^k-\mathbf{C}_i^{k,H}\mathbf{H}\bar{\mathbf{X}}_i^k\\&\mathbf{D}((i-1)\times N:i\times N,1)=\mathbf{D}_i\end{aligned} \tag{31}$$

$\bar{\mathbf{X}}_i^k$ is the expectation on the basis of prior information from decoder. After all these processes, $\hat{\mathbf{X}}$ have been obtained. Then data from one transmit antenna $\hat{\mathbf{X}}_i$ are considered together, and IFFT is implemented.

$$\hat{\mathbf{x}}_i=\mathbf{F}\hat{\mathbf{X}}_i \tag{32}$$

IV. Complexity Comparison and Simulation

TABLE I. CALCULATION COMPLEXITY COMPARISON

Algorithm	**Complex multiplications**
Traditional SC-FDE	$O\left((N\times n_R)^3\right)$
Traditional Turbo FDE	$O\left((N\times n_R)^3\right)$
FD-SIC Turbo FDE	$O\left((N\times n_R)^3\right)$
FD-SIC SFT Turbo FDE	$O\left((n_R)^3\right)$

Complexity comparison is given in Table I. From Table 1, we can see that the complexity for equalization is the same among three algorithms. However, the times of equalization of

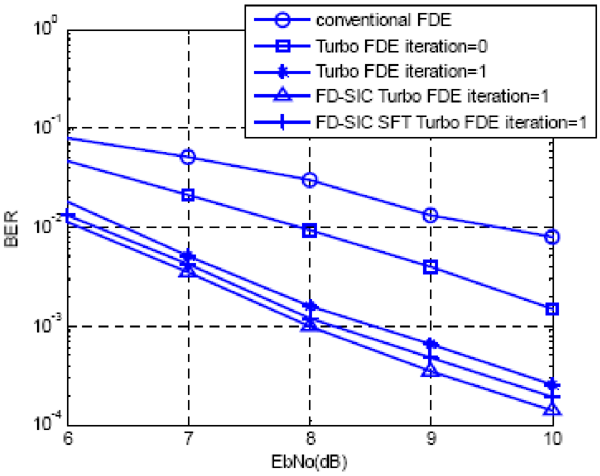

Figure 2 BER performance comparison ($\lambda = 0$)

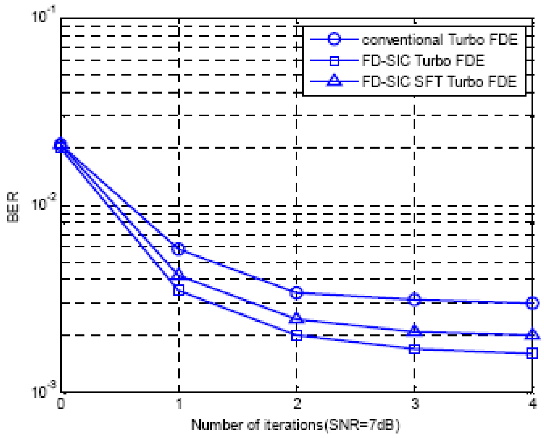

Figure 3 BER performance versus the number of iterations

proposed Turbo FDE is n_T, and extra process as SIC is executed. According to the simulation result, the extra complexity compensates the link performance. SFT Turbo FDE has a little link performance losing compared with FD-SIC Turbo FDE.

In the simulation, modulation mode is quadrature phase shift keying (QPSK) and a 64-point FFT and IFFT are used for the frequency domain processing. Convolution code rate is 1/2. Let the length of CP be 16. These 80 symbols occupy 2Mbaud, and are modulated to 2GHz. It is assumed that the channel is quasi-static for each data block. The frequency selective channel is set to be 7 paths with a mobile velocity of 30km/h. Aaccurate channel information are available at the receiver. In the simulation, independent coded data packets are transmitted with frequency selective MIMO channel. Also the different performance of different correlation coefficient can be seen from the simulation result. E_b/N_o is defined as the average transmit energy per bit over the received AWGN single-side power spectral density at the input of the receive filter per antenna.

Fig.2 gives the BER performance comparison between four algorithms under the correlation coefficient $\lambda = 0$. We can see that the performance of tradition SC-FDE without iteration is the worst. The proposed SIC Turbo FDE is better than traditional Turbo FDE, because they do obtain the link gain from SIC. The performance of FD-SIC SFT Turbo FDE is a little worse than FD-SIC Turbo FDE as the approximately process.

Fig.3 shows the BER performance of different algorithms versus the number of iterations for SNR=7dB with Ḱ Ḱ 0 , respectively. It is clear that the performance floor for all algorithms is existent, and the iterative gains become comparatively small after three iterations. So the best time of iteration should be two, as more iteration times would bring more delay. This will decrease executive efficiency. Complexity Comparison and Simulation

In this paper, a frequency domain block wise Turbo equalization with frequency domain SIC is proposed. This algorithm utilizes the prior information to eliminate the IAI and improve the SINR for data from every transmit antenna. Then MMSE equalization is conducted according to prior information as traditional turbo equalization. So from theory and simulation, it can reach an obvious link level performance gain. The simplified scheme executes equalization on single frequency tone to decrease calculation of matrix inverse complexity in high degree with a little performance losing.

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