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Simplified MMSE Detectors for Turbo Receiver in BICM MIMO Systems

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Abstract In this article, two methods adopting simplified minimum mean square error (MMSE) filter with soft parallel interference cancellation (SPIC) are discussed for turbo receivers in bit interleaved coded modulation (BICM) multiple-input multiple-output (MIMO) systems. The proposed methods are utilized in the non-first iterative process of turbo receiver to suppress residual interference and noise. By modeling the components of residual interference after SPIC plus the noise as uncorrelated Gaussian random variables, the matrix inverse for weighting vector of conventional MMSE becomes unnecessary. Thus the complexity can be greatly reduced with only slight performance deterioration. By introducing optimal ordering to SPIC, performance gap between simplified MMSE and conventional MMSE further narrows. Monte Carlo simulation results confirm that the proposed algorithms can achieve almost the same performance as the conventional MMSE SPIC in various MIMO configurations, but with much lower computational complexity.

Keywords bit interleaved coded modulation multiple-input multiple-output, ordered soft parallel interference cancellation, soft parallel interference cancellation, simplified minimum mean square error, turbo receiver

1 Introduction

With the rapid development of wireless communications in recent years[1-4], the requirement of services with high data rate and high quality of service (QoS) increases. Design of maximizing data rates with limited spectrum resources has always been a major challenge in research. Bit interleaved coded modulation (BICM)[5] is a spectrum-efficient technique which employs bit interleaving to couple channel coding with high order modulations for wireless communications over fading channels. BICM technology could increase the diversity order up to binary Hamming distance of the code[6], which makes the bit error rate (BER) curve steeper. However, due to the random modulation caused by bit interleaving, the reduction of free Euclidean distance is unavoidable in BICM. The research of Li in [7] demonstrated that its performance can be improved by using iterative receiver. Thus, most research work on blending BICM and multiple-input multiple-output (MIMO) here focuses on designing iterative receivers[8-11].

Iterative receiver, also known as turbo receiver, applies the turbo principle which was proposed by Berrou[12] in 1993. MIMO turbo receiver consists of an SISO (soft input soft output) MIMO detector and an SISO channel decoder. The methods of SISO MIMO detection can be mainly divided into three categories: MAP (maximum a posteriori), ML (maximum likelihood), and MMSE. Sphere decoding, a suboptimal ML algorithm, was introduced by Viterbo in [13], and then applied in MIMO detection by Damen in [14]. Based on the research above, a serial of MIMO detectors in iterative receivers were designed recently, including a list sphere decoding[15], iterative tree searching detector[16], expectation maximum[17], sequence Monte-Carlo[18], etc. In the domain of linear detection, algorithms combining IC (interference cancellation) and MMSE filter

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were widely investigated, especially for MIMO OFDM (orthogonal frequency division multiplexing) systems which are widely used for wideband communications[11,19-20]. The complexities of MAP and ML increase exponentially with the number of transmit antennas, the order of modulation, and the channel memory length. Even though MMSE IC algorithms have comparatively low computation complexity and acceptable performance, matrix inverse is needed for detecting each symbol in each iterative process. Thus, the complexity is still prohibitive. In order to reduce the complexity, simplified MMSE SPIC (soft parallel interference cancellation) algorithms, which consider residual interference after SPIC plus noise as approximately uncorrelated random variables are proposed in this article. By avoiding matrix inverse in MMSE filtering, the computational complexity is reduced significantly with slight performance loss. Optimal ordering is further adopted to compensate the loss. Since the simplified detector can be used in both OFDM and non-OFDM systems without any difference, this article discusses the method under a universal MIMO system.

The rest of this paper is organized as follows. In Section 2, BICM MIMO-SM (spatial multiplexing) system model is introduced. Section 3 briefly introduces two conventional MMSE IC methods for comparison, as well as part of the system model. The proposed simplified MMSE SPIC methods with and without ordering are stated detailedly in Section 4, along with the complexity analysis. Performance comparison of the proposed methods and other algorithms is shown via Monte-Carlo simulation in Section 5. Finally it comes to the conclusions in Section 6.

2 System Model

Fig.1 shows the block diagram of the transmitter and turbo receiver of BICM MIMO-SM system with N_t transmit antennas and N_r receive antennas.

At the transmitter, the long data sequence $\{a_i\}$ is firstly channel encoded and bit interleaved. The modulation module maps every $N_t \log_2 |\Omega_S|$ bits into a block of N_i symbols, where Ω_S is the set of signal constellation points. The N_t symbols are transmitted within the same frequency band and time slots[21], and can be written as $\boldsymbol{s} = (s_1, s_2, \ldots, s_{N_t})^{\mathrm{T}}$. They are transmitted over an $N_r \times N_t$ MIMO radio channel described by channel matrix $\boldsymbol{H} = (h_{ik})_{N_r \times N_t}$, and the receive signal is given by:

$$\boldsymbol{r} = \boldsymbol{H}\boldsymbol{s} + \boldsymbol{n} = \sum\nolimits_{k=1}^{N_t} \boldsymbol{h}_k s_k + \boldsymbol{n}. \tag{1}$$

In (1), $\boldsymbol{h}_k$ represents the $N_r \times 1$ channel coefficient vector of the transmit antenna k. $\boldsymbol{H}$ is known at the receiver. The wireless channel is assumed to be rich-scattering and flat-fading. The fading between each antenna pair of transmitter and receiver is assumed independent. $\boldsymbol{n} = (n_1, n_2, \ldots, n_{N_r})$ is independently and identically distributed complex Gaussian noise with mean zero and variance σ_n^2.

At the receiver, using the estimated channel state information, SISO MIMO detector derives the information of transmitted symbols from receive signal $\boldsymbol{r}$.

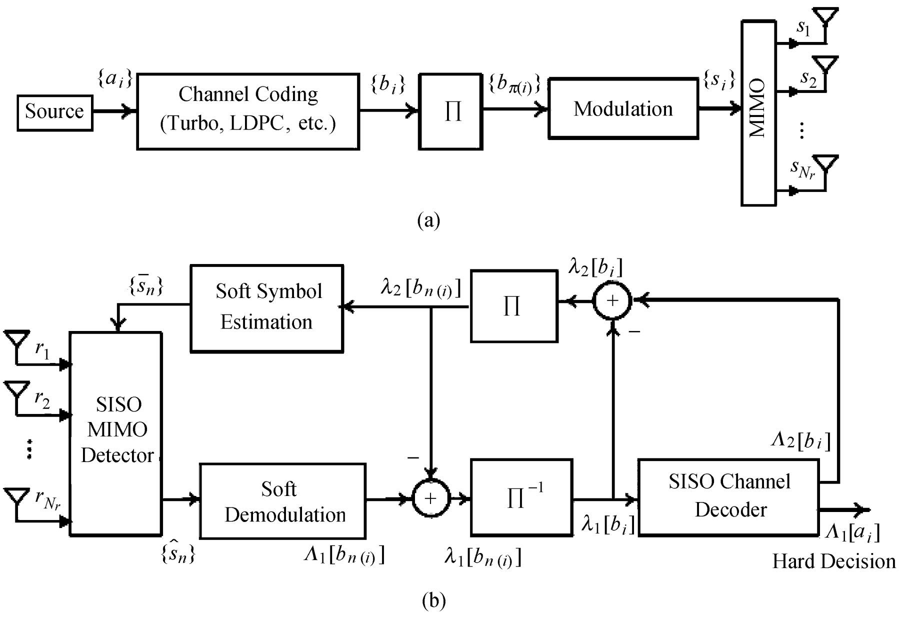

Fig.1. Block diagram of BICM MIMO-SM. (a) Transmitter. (b) Turbo receiver.

Estimated symbols $\{\hat{s_n}\}$ are fed to the demodulator and de-interleaver before the SISO channel decoder. The LLR (log likelihood ratio) produced by channel decoder is feedback to MIMO detector, to act as a priori information for the next iteration of receiving process.

3 Conventional MMSE Methods

In this section, two conventional MIMO detectors are introduced. One is optimal ordered serial IC (OSIC) with MMSE filter, the other is SPIC with conventional MMSE (CMMSE) in turbo receiver. The performance of the two methods will be discussed later as comparison of the two proposed methods. Meanwhile, the former one is also used in the first iteration of turbo receiver discussed in this article.

3.1 CMMSE with OSIC

OSIC detector does not detect the signals at one run. Instead, it starts with linear detection of only one selected sub-stream. This sub-stream was selected because it holds the best post detection signal to noise ratio (post SNR) among all the streams. Then the effect of the detected signal is subtracted from the receive signals, resulting in a modified version of receive signals. This process proceeds until all the signals are detected. The full OSIC with CMMSE filtering can be described compactly as a recursive procedure, as follows:

1) *Initialization*:

$$i \leftarrow 1,$$

$$\boldsymbol{G}_1 = (\boldsymbol{H}^{\mathrm{H}} \cdot \boldsymbol{H} + \sigma_n^2 \cdot \boldsymbol{I}_{N_r})^{-1} \cdot \boldsymbol{H}^{\mathrm{H}},$$

$$k_1 = \arg\min_j \|(\boldsymbol{G}_1)_j\|^2,$$

2) *Recursion*:

$$\boldsymbol{w}_{k_i} = [(\boldsymbol{G}_i)_{k_i}]^{\mathrm{T}},$$

$$y_{k_i} = \boldsymbol{w}_{k_i}^{\mathrm{T}} \cdot \boldsymbol{r}_i,$$

$$\hat{a}_{k_i} = Q(y_{k_i}),$$

$$\boldsymbol{r}_{i+1} = \boldsymbol{r}_i - \hat{a}_{k_r} \cdot (\boldsymbol{H})_{k_i},$$

$$\boldsymbol{G}_{i+1} = (\boldsymbol{H}_{\bar{k}}^{\mathrm{H}} \boldsymbol{H}_{\bar{k}} + \sigma_n^2 \cdot \boldsymbol{I}_{\bar{k}})^{-1} \cdot \boldsymbol{H}_{\bar{k}}^{\mathrm{H}},$$

$$k_{i+1} = \mathop{\arg\min}_{j \notin \{k_1,\ldots,k_i\}} \|(\boldsymbol{G}_{i+1})_j\|^2,$$

$$i \leftarrow i+1,$$

where $(\boldsymbol{G}_i)_j$ denotes the j-th column of $\boldsymbol{G}_i$, $\boldsymbol{H}_{\bar{k}}$ is obtained from $\boldsymbol{H}$ by setting $\boldsymbol{h}_{k_1}, \ldots, \boldsymbol{h}_{k_i}$ to zero. $Q(\cdot)$ stands for quantization operation. $\{k_1, k_2, \ldots, k_{N_t}\}$ is a permutation of the integers $1, 2, \ldots, N_t$ specifying the order in which components of the transmitted symbol vector are detected. The above ordering method is optimal in the sense of post SNR of all data streams[21].

3.2 CMMSE with SPIC

In the non-first iterative processing of turbo receiver, SPIC could be adopted in MIMO detection utilizing the symbol estimation from the last iteration[22]. The SPIC with CMMSE could be described compactly as follows:

1) *Initialization*:

$$k \leftarrow 1,$$

2) *Recursion*:

$$\boldsymbol{w}^k = \boldsymbol{h}_k^{\mathrm{H}} (\boldsymbol{H}\boldsymbol{V}^k \boldsymbol{H}^{\mathrm{H}} + \sigma_n^2 \boldsymbol{I}_{N_r})^{-1},$$

$$\boldsymbol{r}_k = \boldsymbol{r} - \sum_{j=1, j\neq k}^{N_t} \tilde{s}_k \cdot \boldsymbol{h}_j,$$

$$\hat{s_k} = \boldsymbol{w}^k \cdot \boldsymbol{r}_k,$$

$$k \leftarrow k+1.$$

4 Simplified MMSE SPIC MIMO Detector

In the non-first iterative processing, SPIC could be adopted in MIMO detector using a priori information. In this article, novel SPIC MMSE MIMO detectors with and without ordering are introduced for turbo receiver in BICM MIMO-SM systems. Both the two methods are carefully designed for implementation possibility.

4.1 Signal Modeling After SPIC

Assume that in the non-first iteration, the MIMO detector employs SPIC with MMSE filter. According to the system model demonstrated in (1), when estimating the symbol of the k-th data stream, the receive signal after SPIC can be shown as (2), if the residual interference of SPIC is considered:

$$\boldsymbol{r}^k = \boldsymbol{r} - \sum_{i\neq k}^{N_t} \boldsymbol{h}_i \tilde{s}_i = \boldsymbol{h}_k s_k + \sum_{i\neq k}^{N_t} \boldsymbol{h}_i (s_i - \tilde{s}_i) + \boldsymbol{n}, \quad (2)$$

where, $\tilde{s}_i$ is an estimated version of symbol s_i fed by the soft symbol estimation module in the last iteration. Equation (2) shows that the observed signal for s_i after SPIC consists of three parts: $\boldsymbol{h}_k s_k$ is a weighted version of the desired symbol s_k, $\sum_{i\neq k}^{N_t} \boldsymbol{h}_i (s_i - \tilde{s}_i)$ represents the residual interference of all the other transmitted symbols, and $\boldsymbol{n}$ is additive white Gaussion noise (AWGN). According to [23], the elements of $\sum_{i\neq k}^{N_t} \boldsymbol{h}_i (s_i - \tilde{s}_i)$ can be approximately modeled as independent Gaussian random variables. As the components in $\boldsymbol{n}$ are also independent Gaussian random variables, the sum of the interference and noise in the received signal can be modeled as an equivalent noise, denoted by $\boldsymbol{n}^k$:

$$\boldsymbol{r}^k = \boldsymbol{h}_k s_k + \boldsymbol{n}^k.$$

For the equivalent noise $\boldsymbol{n}^k = (n_1^k, n_2^k, \ldots, n_{N_r}^k)$, the expectation of its component $n_i^k = \sum_{j\neq k}^{N_t} h_{ij}(s_j - \tilde{s}_j) +$

n_i is 0, and its variance is deduced as:

$$\begin{aligned}\mathrm{Var}(n_i^k) &\triangleq \sigma^2_{n^k}(i) = E(n_i^k(n_i^k)^*)\\ &= E\bigg(\Big(\sum_{j\neq k}^{N_t} h_{ij}(s_j - \tilde{s}_j) + n_i\Big)\times\\ &\quad \Big(\sum_{j\neq k}^{N_t} h_{ij}(s_j - \tilde{s}_j) + n_i\Big)^*\bigg)\\ &= \Big[\sum_{j=1}^{N_t}(|h_{ij}|^2\sigma_s^2(j)) + \sigma_n^2\Big] - |h_{ik}|^2\sigma_s^2(k),\end{aligned} \tag{3}$$

where, $|\cdot|^*$ denotes the conjugation. $\sigma_s^2(k)$ stands for the variance of symbol s_k which is fed back by the soft symbol estimation module from the previous iteration. $\sigma_s^2(k)$ can be calculated as follows:

$$\begin{aligned}\sigma_s^2(k) &\triangleq \mathrm{Var}(s_k) = E(|s_k|^2) - E(s_k)^2\\ &= \sum_{S_j\in\Omega_S} |S_j|^2 \times P(\tilde{s}_k = S_j) -\\ &\quad \Big(\sum_{S_j\in\Omega_s} S_j \times P(\tilde{s}_k = S_j)\Big)^2\\ &= \sum_{S_j\in\Omega_S} |S_j|^2 \prod_{m=1}^{\log_2 M} \frac{\exp(B_m^j\lambda_2[b_m^k])}{1+\exp(B_m^j\lambda_2[b_m^k])} -\\ &\quad \Big(\sum_{S_j\in\Omega_s} S_j \prod_{m=1}^{\log_2 M} \frac{\exp(B_m^j\lambda_2[b_m^k])}{1+\exp(B_m^j\lambda_2[b_m^k])}\Big)^2,\end{aligned}$$

where, b_m^k denotes the m-th bit of the bit sequence mapping to symbol s_k, $S_j \in \Omega_S$, B_m^j is the m-th bit of the bit sequence mapping to constellation point S_j.

From (3), it can be seen that $\sum_{j=1}^{N_t}(|h_{ij}|^2\sigma_s^2(j)) + \sigma_n^2$ appears in the i-th component of equivalent noise for all the symbols. The calculation result of $\sum_{j=1}^{N_t}(|h_{ij}|^2\sigma_s^2(j)) + \sigma_n^2$ can be used in all N_t stream estimations, which significantly reduces the computational complexity.

4.2 Simplified MMSE SPIC Without Ordering

Based on the above signal model, the estimation value of the k-th transmit symbol can be derived according to (4)[24]:

$$\hat{s}_k = \mathrm{E}(s_k) + Cov(s_k, \boldsymbol{r}^k)Cov(\boldsymbol{r}^k, \boldsymbol{r}^k)^{-1}(\boldsymbol{r}^k - \mathrm{E}(\boldsymbol{r}^k)), \tag{4}$$

where,

$$\begin{aligned}Cov(s_k, \boldsymbol{r}^k) &= \mathrm{E}\{(s_k - \mathrm{E}(s_k))[\boldsymbol{r}^k - \mathrm{E}(\boldsymbol{r}^k)]^{\mathrm{H}}\}\\ &= \mathrm{E}\{(s_k - \mathrm{E}(s_k))[\boldsymbol{h}_k(s_k - \mathrm{E}(s_k))]^{\mathrm{H}}\}\\ &= (h_{1k}^*\ h_{2k}^*\ \cdots\ h_{N,k}^*)\sigma_s^2(\boldsymbol{k}) \triangleq \boldsymbol{P}.\end{aligned} \tag{5}$$

In (5), $\sigma_s^2(k) = \mathrm{Var}(s_k)$, $|\cdot|^{\mathrm{H}}$ denotes the conjugation transpose. The covariance of $\boldsymbol{r}^k$ is calculated as (6).

$$\begin{aligned}Cov(\boldsymbol{r}^k, \boldsymbol{r}^k) &= \mathrm{E}\{[\boldsymbol{r}^k - \mathrm{E}(\boldsymbol{r}^k)][\boldsymbol{r}^k - \mathrm{E}(\boldsymbol{r}^k)]^{\mathrm{H}}\} = \mathrm{E}\{[\boldsymbol{h}_k(s_k - \mathrm{E}(s_k)) + \boldsymbol{n}^k][\boldsymbol{h}_k(s_k - \mathrm{E}(s_k)) + \boldsymbol{n}^k]^{\mathrm{H}}\}\\ &= \begin{pmatrix} |h_{1k}|^2\sigma_s^2(k) + \sigma^2_{n^k}(1) & h_{1k}h_{2k}^*\sigma_s^2(k) & \cdots & h_{1k}h_{N_rk}^*\sigma_s^2(k)\\ h_{2k}h_{1k}^*\sigma_s^2(k) & |h_{2k}|^2\sigma_s^2(k) + \sigma^2_{n^k}(2) & \cdots & h_{2k}h_{N_rk}^*\sigma_s^2(k)\\ \vdots & \vdots & \ddots & \vdots\\ h_{N_rk}h_{1k}^*\sigma_s^2(k) & h_{N_rk}h_{2k}^*\sigma_s^2(k) & \cdots & |h_{N_rk}|^2\sigma_s^2(k) + \sigma^2_{n^k}(N_r)\end{pmatrix} \triangleq G.\end{aligned} \tag{6}$$

In a turbo receiver, a symbol estimate must be independent from the priori symbol information[25]. Thus, $\mathrm{E}(s_k) = 0$, $\sigma_s^2(k) = 1$ in (5) and (6). Further assuming $\boldsymbol{w} = (w_1^k, w_2^k, \ldots, w_{N_r}^k) = \boldsymbol{P}\boldsymbol{G}^{-1}$, then $\boldsymbol{wG} = \boldsymbol{P}$ can be attained by observation, that is (7):

$$\begin{aligned}&(|h_{1k}|^2 + \sigma^2_{n^k}(1))w_1^k + (h_{2k}h_{1k}^*)w_2^k + \cdots + (h_{N_rk}h_{1k}^*)w_{N_r}^k = h_{1k}^*,\\ &(h_{1k}h_{2k}^*)w_1^k + (|h_{2k}|^2 + \sigma^2_{n^k}(2))w_2^k + \cdots + (h_{N_rk}h_{2k}^*)w_{N_r}^k = h_{2k}^*,\\ &\qquad\vdots\\ &(h_{1k}h_{N_rk}^*)w_1^k + (h_{2k}h_{N_rk}^*)w_2^k + \cdots + (|h_{N_rk}|^2 + \sigma^2_{n^k}(N_r))w_{N_r}^k = h_{N_rk}^*.\end{aligned} \tag{7}$$

By multiplying h_{ik} in both sides of the i-th equation of (7) and adding all the equations together, the following result can be obtained as (8):

$$\begin{aligned}&\Big(\sum_{j=1}^{N_r}|h_{jk}|^2 + \sigma^2_{n^k}(1)\Big)h_{1k}w_1^k + \Big(\sum_{j=1}^{N_r}|h_{jk}|^2 + \sigma^2_{n^k}(2)\Big)h_{2k}w_2^k + \cdots +\\ &\Big(\sum_{j=1}^{N_r}|h_{jk}|^2 + \sigma^2_{n^k}(N_r)\Big)h_{N_rk}w_{N_r}^k = \sum_{j=1}^{N_r}|h_{jk}|^2.\end{aligned} \tag{8}$$

From (8), (9) can be derived.

$$w_i^k = \frac{h_{ik}^*}{\left(\sum_{j=1}^{N_r} |h_{jk}|^2 + \sigma_{n^k}^2(i)\right)}, \quad i = 1, 2, \ldots, N_r. \tag{9}$$

According to the calculation in (9), the weighting vector of MMSE can be obtained without a matrix inverse. Furthermore, for the weighting vector of the k-th symbol, the N_r elements hold a common part as $\sum_{j=1}^{N_r} |h_{jk}|^2$, which can also reduce the calculation of this part by $N_r - 1$ times. Based on the above derivation, and cite $\mathrm{E}(s_k) = 0$, $\sigma_s^2(k) = 1$ in (4), the MMSE estimation can be performed as (10):

$$\begin{aligned}\hat{s}_k &= \boldsymbol{w}^k(\boldsymbol{h}_k s_k + \boldsymbol{n}^k - \mathrm{E}(\boldsymbol{h}_k \boldsymbol{s}_k + \boldsymbol{n}^k)) \\ &= \boldsymbol{w}^k(\boldsymbol{h}_k s_k + \boldsymbol{n}^k) = \boldsymbol{w}^k \boldsymbol{r}^k.\end{aligned} \tag{10}$$

According to [22], the output of the MMSE filter can be modeled as the output of an AWGN channel:

$$\hat{s}_k = \mu_k s_k + \eta_k,$$

where, the expectation and variance are:

$$\begin{aligned}&\mu_k = \mathrm{E}\{\hat{s}_k s_k^*\} = \mathrm{E}(\boldsymbol{w}^k \boldsymbol{r}^k s_k^*) = \boldsymbol{w}^k \mathrm{E}(\boldsymbol{r}^k s_k^*) = \boldsymbol{w}^k \boldsymbol{h}_k, \\ &v_k^2 = \mathrm{Var}(\hat{s}_k) = \mu_k - \mu_k^2.\end{aligned}$$

Based on this Gaussian model, the bit posteriori LLR $\Lambda_1[b_j^k]$ and extrinsic LLR $\lambda_1[b_j^h]$ are derived:

$$\Lambda_1[b_j^k] = \underbrace{\log \frac{\sum\limits_{S_l \in S_j^+} \exp\left(-\frac{|\hat{s}_k - \mu_k S_l|^2}{\mathrm{Var}(\hat{s}_k)}\right) \prod\limits_{m=1, m\neq j}^{M} P(b_m^k = B_m^l)}{\sum\limits_{S_l \in S_j^-} \exp\left(-\frac{|\hat{s}_k - \mu_k S_l|^2}{\mathrm{Var}(\hat{s}_k)}\right) \prod\limits_{m=1, m\neq j}^{M} P(b_m^k = B_m^l)}}_{\lambda_1[b_j^k]} + \underbrace{\log \frac{P(b_j^k = +1)}{P(b_j^k = -1)}}_{\lambda_2[b_j^k]},$$

where, b_j^k is the j-th bit of the bit sequence mapping to symbol s_k, $S_l \in \Omega_S$, S_j^+, S_j^- denotes the constellation points whose j-th bit is $+1$ and -1, respectively. B_m^l is the m-th bit of the bit sequence mapping to constellation S_l. $\lambda_2[b_j^k]$ denotes the priori LLR.

The proposed simplified MMSE (SMMSE) SPIC without ordering detector can be described compactly as follows:

1) *Initialization*:

$$k \leftarrow 1, \tag{11}$$

$$\sigma_s^2(j) = \mathrm{Var}(\tilde{s}_j) \quad (j = 1, 2, \ldots, N_t), \tag{12}$$

2) *Recursion*:

$$\sigma_{n^k}^2(i) = \sum_{j=1}^{N_t} (|h_{ij}|^2 \sigma_s^2(j)) + \sigma_n^2 - |h_{ik}|^2 \sigma_s^2(k) \quad (i = 1, 2, \ldots, N_r), \tag{13}$$

$$\boldsymbol{w}^k = (w_1^k, w_2^k, \ldots, w_{N_r}^k),$$

$$\boldsymbol{w}_i^k = \frac{h_{ik}^*}{\left(\sum\limits_{j=1}^{N_r} |h_{jk}|^2 + \sigma_{n^k}^2(i)\right)} \quad (i = 1, 2, \ldots, N_r), \tag{14}$$

$$\boldsymbol{r}_k = \boldsymbol{r} - \sum_{j=1, j\neq k}^{N_t} \tilde{a}_j \cdot \boldsymbol{h}_j, \tag{15}$$

$$\hat{s}_k = \boldsymbol{w}^k \cdot \boldsymbol{r}_k, \tag{16}$$

$$k \leftarrow k + 1. \tag{17}$$

4.3 SMMSE SPIC with Optimal Ordering

In Subsection 3.1, OSIC with MMSE filter in conventional receiver is introduced. The detector processes only one sub-stream with the best post SNR among all the undetected streams each time. Then the effect of the detected signal is subtracted from the receive signals. In this subsection, a novel ordered SPIC (OSPIC) SMMSE detector is proposed for turbo receiver. In the non-first iteration of turbo receiver, the $\tilde{s}_i$ fed back by the soft symbol estimation module in previous iteration, are used for SPIC for the detection of every stream. In this proposed method, the estimated $\hat{s}_k$ from (10) will be updated into the vector of $\tilde{s}_k$ for the SPIC for all the undetected sub-streams. Since better detection performance of sub-stream will benefit the subsequent SPIC, detection order is a remarkable topic to be discussed in this scenario. The proposed optimal ordering SPIC detection with MMSE filter can be described compactly as follows:

1) *Initialization*:

$$d \leftarrow 1, \tag{18}$$

$$\sigma_s^2(j) = \mathrm{Var}(\tilde{s}_j) \quad (j = 1, 2, \ldots, N_t), \tag{19}$$

2) *Recursion*:

$$\sigma^2_{n^k}(i) = \Big[\sum_{j=1}^{N_t}(|h_{ij}|^2\sigma^2_s(j)) + \sigma^2_n\Big] - |h_{ik}|^2\sigma^2_s(k)$$
$$(i = 1, 2, \ldots, N_r; k = 1, 2, \ldots, N_t), \tag{20}$$

$$\boldsymbol{w}^k = (w^k_1, w^k_2, \ldots, w^k_{N_r})$$
$$(k = 1, 2, \ldots, N_t;\ k \neq k_1, k_2, \ldots, k_{d-1}), \tag{21}$$

$$w^k_i = \frac{h^*_{ik}}{\Big(\sum_{j=1}^{N_r}|h_{jk}|^2 + \sigma^2_{n^k}(i)\Big)},$$

$$k_d = \arg\min_k \|\boldsymbol{w}^k\|^2$$
$$(k = 1, 2, \ldots, N_t;\ k \neq k_1, k_2, \ldots, k_{d-1}), \tag{22}$$

$$\boldsymbol{r}_{k_d} = \boldsymbol{r} - \sum_{j=1, j\neq k_d}^{N_t} \tilde{a}_j \cdot \boldsymbol{h}_j, \tag{23}$$

$$\hat{s}_{k_d} = \boldsymbol{w}^{k_d} \cdot \boldsymbol{r}_{k_d}, \tag{24}$$

$$\tilde{s}_{k_d} \leftarrow \hat{s}_{k_d}, \tag{25}$$

$$d \leftarrow d + 1. \tag{26}$$

4.4 Complexity Analysis

The computational complexity of the proposed SPIC MMSE with and without ordering methods will be analysed and compared with conventional IC MMSE detection in this subsection. Comparison will be carried out in two groups. One pair is CMMSE SPIC and the proposed SMMSE SPIC. The other pair is the proposed two methods: SMMSE SPIC and SMMSE OSPIC, which are described in (11)∼(17) and (18)∼(26) respectively.

4.4.1 *Complexity of MMSE Weighting Vector Calculation*

For the two SPIC detectors with CMMSE filter and SMMSE filter, the IC parts of them are the same. So only the complexity of MMSE weighting vector calculation is considered here.

In CMMSE filtering, the weighting vector for each transmit symbol is given by:

$$\boldsymbol{w}^k = \boldsymbol{h}^{\mathrm{H}}_k(\boldsymbol{H}\boldsymbol{V}^k\boldsymbol{H}^{\mathrm{H}} + \sigma^2_n\boldsymbol{I}_{N_r})^{-1},$$

where, $\boldsymbol{V}_k = \mathrm{diag}\{\sigma^2_s(1), \sigma^2_s(2), \ldots, \sigma^2_s(k-1), 1, \sigma^2_{s(k+1)}, \ldots, \sigma^2_{s(N_t)}\}$. Thus, the number of calculations (only real multiplications and real additions are taken into account) for N_t sub-streams estimation can be counted as:

$$MUL_{\mathrm{CMMSE}} = 6N^3_rN_t + 4N^2_tN^2_r + 2N^2_rN_t + 2N_rN_t,$$
$$ADD_{\mathrm{CMMSE}} = 6N^3_rN_t + 4N^2_tN^2_r - 4N^2_rN_t - 2N^2_tN_r.$$

The SMMSE weighting vector for N_t sub-streams estimation are described in (13) and (14), for which the amount of real number calculations can be derived as:

$$MUL_{\mathrm{SMMSE}} = 10N_rN_t,$$
$$ADD_{\mathrm{SMMSE}} = 4N_tN_r - N_t.$$

It is obvious that the complexity order of CMMSE is $O(N^3_rN_t + N^2_rN^2_t)$, while, for the proposed algorithm, it is $O(N_rN_t)$. Table 1 shows the number of calculations for the above two algorithms with various MIMO configurations. It can be seen that compared with CMMSE, the complexity of SMMSE algorithm is reduced significantly.

4.4.2 *Complexity of IC Process*

The proposed methods in Subsections 4.2 and 4.3 mainly differ from each other in the steps of (13)∼(14) and (20)∼(22). For the method of SMMSE SPIC without ordering, as shown in (13) and (14), the variance of post detection equivalent noise and MMSE filter weighting vector for each sub-stream can be derived independently. But for the method of OSPIC with SMMSE proposed in Subsection 4.3, the variance of post detection equivalent noise and weighting vector for each sub-stream should be updated in each recursion process. And for the purpose of optimal ordering, $\|\boldsymbol{w}^k\|^2$ are also essential in the process. Based on the above analysis, complexity order of SPIC SMMSE is $O(N_rN_t)$, while, for OSPIC SMMSE, it is $O(N_rN^3_t)$. It has been shown in last subsection that the complexity of unordered SPIC with conventional MMSE filter is $O(N^3_rN_t + N^2_rN^2_t)$. For MIMO-SM systems, the relationship of $N_r \geqslant N_t$ is normally held. Thus, OSPIC SMMSE requires no more calculation operations compared with the method of SPIC CMMSE.

Table 1. Complexity Comparison of CMMSE and SMMSE

Configuration of TX & RX	CMMSE		SMMSE	
	Real Number Multiplication	Real Number Addition	Real Number Multiplication	Real Number Addition
4T8R	16 960	15 104	320	124
4T4R	2 720	2 176	160	60
2T4R	1 104	864	80	50

5 Simulation Results

In this section, the performance of the proposed SMMSE SPIC and SMMSE OSPIC will be evaluated in terms of BER with respect to the receive SNR. For comparison purpose, the BER of CMMSE SPIC and CMMSE OSIC are also simulated. For turbo receivers discussed in this section, CMMSE OSIC is employed in the first iteration. The parameter settings used for simulation are listed in Table 2.

Table 2. Simulation Parameters

Parameter	Value
Modulation	Quadrature Phase Shift Keying
Channel coding	1/2 turbo code, block length 5 112
Channel model	Independent & identically flat Rayleigh channel
Channel estimation & synchronization	Ideal channel estimation & ideal synchronization
Maximum iteration number	4

Fig.2 shows BER performance of the methods stated in Section 3 and proposed in Section 4 with antenna configurations of 2T2R. It can be seen that the three discussed turbo receivers have at least 2.5 dB gain at BER $= 10^{-3}$ with four iterations. Compared with the CMMSE SPIC, the proposed SMMSE SPIC suffers slight performance deterioration, while the SMMSE OSPIC almost has the same performance as that of CMMSE SPIC. The reason for the deterioration is that SMMSE considers the residual interference as independent Gaussian variables, and the interference caused by correlation between elements is ignored. On the other hand, as optimal ordering is introduced into SPIC, the performance of SMMSE OSPIC approaches that of the CMMSE SPIC.

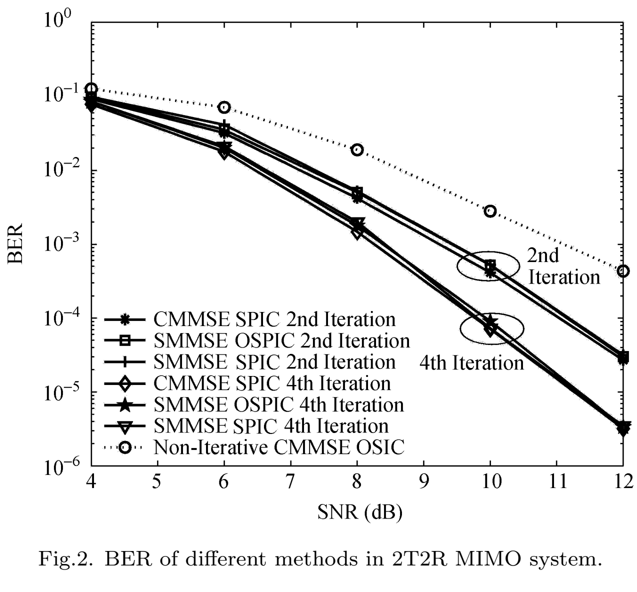

Fig.2. BER of different methods in 2T2R MIMO system.

Fig.3 discusses the performance with antenna configurations of 4T4R. It can be seen that the performance gap between SMMSE SPIC and CMMSE SPIC becomes wider, compared with the situation in 2T2R scenario. That is because the ignored interference takes a more significant effect in 4T4R systems than that with 2T2R. Furthermore, as the number of sub-stream increases, the ordering operation causes more performance gain in contrast to that of 2T2R systems.

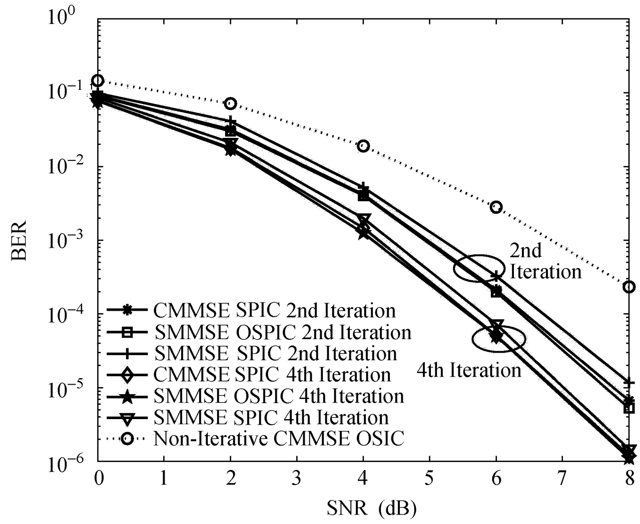

Fig.3. BER of different methods in 4T4R MIMO system.

Fig.4 shows the performance in 4T8R system. As the number of receiving antenna increases, the performance gain achieved by iterative process reduces to about 1.5 dB at BER $= 10^{-3}$. Meanwhile, the gap between SMMSE and CMMSE also becomes narrower.

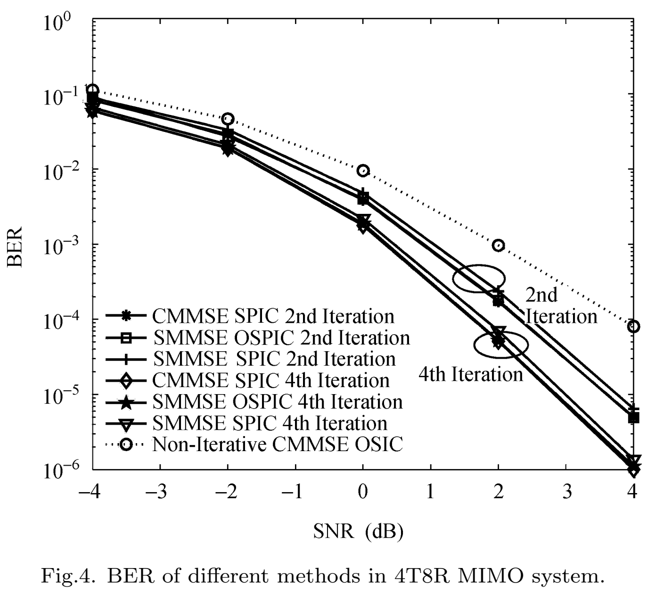

Fig.4. BER of different methods in 4T8R MIMO system.

Based on the above discussions, though there is a slight deterioration in the performance, the two proposed SMMSE methods reduce the computational complexity significantly. Thus, they would achieve a good trade-off between complexity and performance in practical system.

6 Conclusions

In this article, two low complexity MIMO detection algorithms, based on SMMSE and SPIC, were proposed for turbo receivers in BICM MIMO SM systems. By assuming residual interference after SPIC plus AWGN as independent Gaussian variables, the calculation of weighting vector for MMSE filter can be significantly simplified, which meanwhile causes slight performance deterioration. Introducing optimal ordering to SPIC can effectively compensate the performance loss. Complexity analysis and simulation results show that the performance of this simplified version does not deteriorate greatly, while, the computational complexity is decreased significantly. Thus the two proposed methods are more appropriate for implementation in real systems.

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