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An iterative reaching-diversity bound detector for MIMO SC-FDE systems

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Abstract: To improve the detection reliability and reduce the computational complexity of multiple-input-multiple-output (MIMO) single carrier-frequency-domain equalization (SC-FDE) systems, a novel iterative signal detection algorithm, called iterative interference cancellation (IIC), is proposed. Moreover, a receive diversity lower bound (RDLB) is derived as a benchmark to evaluate the bit error rate performance. The proposed IIC utilizes the initial value provided by the basic linear equalization algorithm, progressively removes the inter-layer interference by the iterative process, and gradually improves the reliability of the detected results. The results indicate that the proposed IIC approaches to the RDLB at the E_b/N_0 of 9 dB and 16 dB for 4-ary and 16-ary quadrature amplitude modulations, respectively. The computational complexity of an iterative loop is linear to the number of transmit antennas, and the overall complexity is lower than that of a popular vertically Bell laboratory layered space-time detector when the antenna size is larger than 3×3. Therefore, the proposed algorithm remarkably enhances the reliability of MIMO SC-FDE systems with an antenna size of not less than 3×3 and is computationally efficient and practical even in large-dimension MIMO systems for 5th-generation (5G) or beyond 5G communications.

Key words: receiving-diversity bound; spatially multiplexed systems; multiple-input-multiple-output (MIMO) single carrier-frequency-domain equalization (SC-FDE); iterative detection algorithm

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Multiple-input-multiple-output (MIMO) single-carrier-frequency-domain equalization (MIMO SC-FDE)[1] has received great attention as it possesses a similar capacity and spectral efficiency to MIMO orthogonal frequency division multiplexing (OFDM) with a lower peak-to-average power ratio and less sensitivity to carrier frequency offset. Consequently, it has been employed as one of the backbone technologies in the standards of long-term evolution (LTE)[2] and is widely concerned by the research community and industry[3-4].

Similar to MIMO OFDM systems, MIMO SC-FDE systems can be considered as multiple parallel independent narrow band MIMO systems in the frequency domain, which are called sub-systems. Both equalization and decision are completed in the frequency domain for MIMO OFDM systems[5], and the probability density function (PDF) of each decision statistic relates only to its own sub-system. Unlike MIMO OFDM systems, MIMO SC-FDE systems reframe the output matrix for IDFT transform after frequency-domain equalization and then make decisions and realize the detection process in the time domain. The PDF of each decision statistic is related to all sub-systems. Hence, the complexity of its maximum likelihood (ML) class algorithm is much higher than that of MIMO OFDM systems, and the ML class algorithm is difficult to implement due to its prohibitive complexity.

Linear algorithms[5], including the minimum mean-square error (MMSE) criterion and zero-forcing (ZF) criterion-based algorithms[6], and enhanced linear algorithms[7] have polynomial complexity. However, they suffer from noise and residual inter-layer interference (ILI), and their bit error rate (BER) performance requires improvement. Therefore, iterative receivers are applied to further reduce the complexity of matrix inversion in the linear detection algorithm for large MIMO systems[8-10].

Various iterative equalizers have been proposed to mitigate the noise. Among them, a promising scheme is frequency-domain equalization with noise prediction (FDE-NP), which predicts the noise of the current time slot by the previous ones[4]. Meanwhile, a block-iterative decision feedback equalizer with noise prediction (NP-BI) predicts the noise by that of all remaining time slots in the previous frame[11]. Clearly, the BER performance of NP-BI is slightly better than that of FDE-NP due to its full-usage information. Guvensen summarized a general framework for both time- and frequency-domain iterative block-wise equalization[12].

However, as the number of antennas increases, the ILI

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has been becoming the main bottleneck of the receivers of MIMO SC-FDE systems. To overcome this deficiency, state-of-art successive interference cancellation algorithms, such as vertically Bell laboratory layered space-time (V-BLAST), have been proposed[13]. Information streams are detected per layer in V-BLAST, and the upper layer will be canceled as interference when it detects the lower layer. Unfortunately, as the upper layer has less diversity gain and causes more errors in V-BLAST algorithms, it decreases the entire BER performance.

An iterative structure is beneficial to the improvement of the overall diversity gain and BER performance[13]. An efficient algorithm is developed in the following context: each layer exhibits a comparable diversity gain due to its iterative interference cancellation process.

1 System Model and Problem Statement

1.1 System model

A MIMO SC-FDE system with N_T transmit antennas and N_R receive antennas is considered, which is illustrated in Fig. 1. The wireless propagation environment is characterized by slow multipath Rayleigh fading, and the channel impulse response keeps invariant during a block transmission cycle and may vary per block[2]. The maximum delay of the channel impulse response is no longer than N_{max} taps. The frequency-selective fading channels between the N_T transmit and N_R receive antennas are uncorrelated to each other. Excellent channel state information (CSI) is assumed to be available at the receiver, but no available CSI is imposed on the transmitter. N_T-independent parallel data streams are launched from N_T transmit antennas, and each stream is called a layer. Each layer is launched in frames, whose length is set to N_C. A cyclic prefix (CP), a copy of the N_{CP} symbol of the end part of a frame, is inserted in front of this frame; $N_{CP} \geqslant N_{max} - 1$ is required to remove the inter-symbol interference (ISI). Further, $N_R \geqslant N_T$ is assumed to separate N_T-independent data streams[10]. A frame of the l-th layer is denoted by $\boldsymbol{x}_l = \{x_l^1, x_l^1, \dots, x_l^{N_C}\}$, $l = 1, 2, \dots, N_T$, where symbol $x_l^k (k = 1, 2, \dots, N_C)$ comes from the Gray-encoded M-ary quadrature amplitude modulation (QAM) constellation with bits of information, and its average power is $\sigma_x^2 = 2(M-1)/3$.

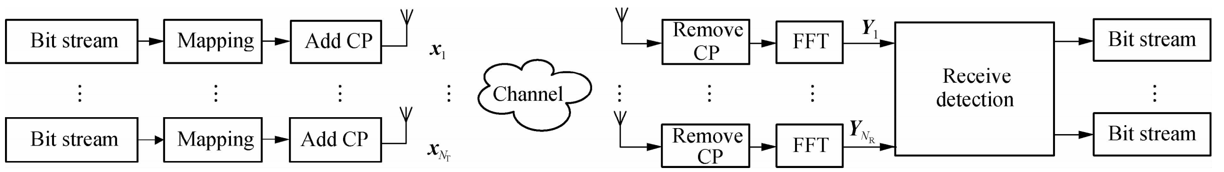


Fig. 1 Block diagram of MIMO SC-FDE systems

The original received frequency-domain baseband signal matrix is expressed as

$$\boldsymbol{Y} = [\boldsymbol{Y}^1 \quad \boldsymbol{Y}^2 \quad \dots \quad \boldsymbol{Y}^k \quad \dots \quad \boldsymbol{Y}^{N_C}] \tag{1}$$

where $\boldsymbol{Y}^k = \{Y_1^k, Y_2^k, \dots, Y_{N_R}^k\}^T$ for $k = 1, 2, \dots, N_C$ denotes the received signal vector of the k-th sub-channel and superscripts in $(\cdot)^T$ represents transpose.

The input-output relation at the k-th sub-channel is given by

$$\boldsymbol{Y}^k = \sum_{l=1}^{N_T} \boldsymbol{H}_l^k \boldsymbol{X}_l^k + \boldsymbol{W}^k = \boldsymbol{H}^k \boldsymbol{X}^k + \boldsymbol{W}^k \qquad k = 1, 2, \dots, N_C \tag{2}$$

where the additive noise vector $\boldsymbol{W}^k = \{W_1^k, W_2^k, \dots, W_{N_R}^k\}^T$ is assumed to be i. i. d. circularly symmetric complex Gaussian with zero mean and variance σ_ω^2. The channel frequency response of thek-thsub-channel is denoted by $\boldsymbol{H}^k \in \boldsymbol{C}^{N_R \times N_T}$ as $\boldsymbol{C}$ is the complex matrix (or vector), $\boldsymbol{Y}^k \in \boldsymbol{C}^{N_R \times 1}$ is the output vector, and $\boldsymbol{X}^k \in \boldsymbol{C}^{N_T \times 1}$ indicates the input vector. The input matrix $\boldsymbol{X} = [\boldsymbol{X}^1 \quad \boldsymbol{X}^2 \quad \dots \quad \boldsymbol{X}^{N_C}]$ can also be expressed as $[\boldsymbol{X}_1^T \quad \boldsymbol{X}_2^T \quad \dots \quad \boldsymbol{X}_{N_T}^T]$, where $\boldsymbol{X}_l = \mathrm{DFT}(\boldsymbol{x}_l) = \boldsymbol{F}_{N_C}(\boldsymbol{x}_l)$, $l = 1, 2, \dots, N_T$. Here $\boldsymbol{F}_{N_C}$ denotes the DFT matrix with N_C dimensions.

1.2 Problem statement

The detection algorithm aims to restore the original data block $\boldsymbol{x}$, which includes two basic procedures: equalization and decision. The frequency-domain equalization is to invert the effect of the channel, that is, to solve the statistical equation, where the input vector $\boldsymbol{X}^k$ is estimated by the output signal vector $\boldsymbol{Y}^k$ and sub-channel matrix $\boldsymbol{H}^k$.

$$\hat{\boldsymbol{X}}^k = f(\boldsymbol{Y}^k \mid \boldsymbol{H}^k) \qquad k = 1, 2, \dots, N_C \tag{3}$$

When all of the estimated signal vectors $\hat{\boldsymbol{X}}^k (k = 1, 2, \dots, N_C)$ are obtained, they are reorganized into N_T data frames.

$$\hat{\boldsymbol{X}}_l = \{\hat{X}_l^1, \hat{X}_l^2, \dots, \hat{X}_l^{N_C}\} \qquad l = 1, 2, \dots, N_T \tag{4}$$

These N_T frames are converted into time domain statistics by IDFT (or IFFT) transformation; that is,

$$\hat{\boldsymbol{s}}_l = \boldsymbol{F}_{N_C}^{-1}(\hat{\boldsymbol{X}}_l) \qquad l = 1, 2, \dots, N_T \tag{5}$$

where $\boldsymbol{F}_{N_C}^{-1}$ denotes the IDFT matrix with N_C dimensions. The decision procedure is to slice the elements of statistic vector $\hat{\boldsymbol{s}}_l$ as coordinates on the M-ary QAM constellation map, according to the ML criterion; that is,

$$\hat{\boldsymbol{x}}_l = D(\hat{\boldsymbol{s}}_l) \qquad l = 1, 2, \dots, N_T \tag{6}$$

where $D(\cdot)$ denotes the decision operation. Ideally, $\hat{\boldsymbol{x}}_l = \boldsymbol{x}_l$; that is, the launched frame is completely restored, and the BER is 0. However, in general, equalization and decision processes cannot completely eliminate interference and noise. Therefore, $\hat{\boldsymbol{x}}_l$ is close, but not equal to $\boldsymbol{x}_l$ ($l = 1, 2, \ldots, N_T$).

2 RDLB and IIC Algorithms for MIMO SC-FDE Detection

Regarding spatially multiplexed systems, interference sources may come from two aspects: ISI and ILI. If sub-channels are orthogonal to each other, the ISI, which is caused by frequency-selective fading while experienced in multipath propagation environments, can be effectively eliminated using appropriate signal processing methods. Meanwhile, the ILI results from the superposition on the receiving ends of N_T bunches of independent transmission data streams. To date, there is no effective means to completely cancel this mutual interference but to suppress it to the extent possible. Clearly, the more thoroughly the interference is suppressed, the better the possible BER performance.

It is easy to understand that a genie-aided system with attainable RDLB implies that it is ILI-free, which is theoretically equivalent to N_T parallel independent single-input multiple-output (SIMO) systems. In SIMO systems, a data stream from a transmitter is simultaneously received by the N_R receiving ends, and N_R-received copies are obtained. The possibility of multiple copies undergoing deep fading simultaneously is greatly reduced. Therefore, the system receiving reliability can be greatly improved by combining based on the ML criterion. Consider the RDLB of this ideal case in the following aspects.

2.1 Receiving-diversity bound

Without loss of generality, the output vector of the l-th layer at the k-th sub-channel is

$$\boldsymbol{P}_l^k = \boldsymbol{H}_l^k \boldsymbol{X}_l^k + \boldsymbol{W}^k \qquad l = 1, 2, \ldots, N_T;\ k = 1, 2, \ldots, N_C \quad (7)$$

If the maximum ratio combination (MRC) criterion is followed[2], then the estimated value of the input frequency-domain component in the 1-th layer is obtained; that is

$$\breve{X}_l^k = X_l^k + (\boldsymbol{H}_l^k)^+ \boldsymbol{W}^k \qquad k = 1, 2, \ldots, N_C \qquad (8)$$

where $(\boldsymbol{H}_l^k)^+ = [(\boldsymbol{H}_l^k)^H (\boldsymbol{H}_l^k)]^{-1} (\boldsymbol{H}_l^k)^{\mathrm{H}}$. The values are organized into a frame after estimating all N_C input frequency-domain components.

$$\begin{aligned}\breve{\boldsymbol{X}}_l = [\breve{X}_l^1 \quad \breve{X}_l^2 \quad \cdots \quad \breve{X}_l^k \quad \cdots \quad \breve{X}_l^{N_C}] = \\ [X_l^1 \quad X_l^2 \quad \cdots \quad X_l^{N_C}]^{\mathrm{T}} + \breve{\boldsymbol{W}}_l = \boldsymbol{X}_l + \breve{\boldsymbol{W}}_l \qquad (9)\end{aligned}$$

where $\boldsymbol{W}_l^1 = \{(\boldsymbol{H}_l^1)^+ \boldsymbol{W}^1, (\boldsymbol{H}_l^2)^+ \boldsymbol{W}^2, \ldots, (\boldsymbol{H}_l^{N_C})^+ \boldsymbol{W}^{N_C}\}^{\mathrm{T}}$. After the IDFT (or IFFT) transformation, $\breve{\boldsymbol{X}}_l$ is converted to statistic vector $\hat{s}_l$

$$\hat{\boldsymbol{s}}_l = \boldsymbol{F}_{N_C}^{-1}(\breve{\boldsymbol{X}}_l) = \boldsymbol{x}_l + \boldsymbol{F}_{N_C}^{-1} \breve{\boldsymbol{W}}_l \qquad (10)$$

Recovered symbols $\hat{\boldsymbol{X}}_l$ are obtained by judging the statistic vector $\hat{\boldsymbol{s}}_l$; that is, $\hat{\boldsymbol{x}}_l = D(\hat{\boldsymbol{s}}_l)$. At this point, the BER of $\hat{\boldsymbol{x}}_l$ is further analyzed. From Eq. (10), the pending statistic vector $\hat{\boldsymbol{s}}_l$ comprises two parts: the originally launched data frame $\boldsymbol{x}_l$, whose components come from the coordinates of the M-ary QAM constellation, and the zero mean Gaussian noise vector $\boldsymbol{F}_{N_C}^{-1} \breve{\boldsymbol{W}}_l$. Thus, $\hat{\boldsymbol{s}}_l$ is a complex Gaussian vector, whose expectation is vector $\boldsymbol{x}_l$, and the correlation matrix can be expressed as

$$\boldsymbol{B}_l = \boldsymbol{F}_{N_C}^{-1} E[\breve{\boldsymbol{W}}_l \breve{\boldsymbol{W}}_l^{\mathrm{H}}] \boldsymbol{F}_{N_C} \qquad l = 1, 2, \ldots, N_T \qquad (11)$$

$$E[\breve{\boldsymbol{W}}_l \breve{\boldsymbol{W}}_l^{\mathrm{H}}] = \mathrm{diag}[\boldsymbol{g}]$$

$$\boldsymbol{g} = \{g_1, g_2, \ldots, g_{N-1}\}^{\mathrm{T}} = \frac{\sigma_\omega^2}{N_C}\{\|(\boldsymbol{H}_l^1)^+\|^2, \|(\boldsymbol{H}_l^1)^+\|^2, \ldots, \|(\boldsymbol{H}_l^{N_C})^+\|^2\}^{\mathrm{T}} \qquad (12)$$

where $E[\cdot]$ denotes the expectation function. The covariance matrix $\boldsymbol{B}_l = [b_{ij}^l]$ ($i, j = 1, 2, \ldots, N_C$), whose diagonal entries b_{kk}^l ($k = 1, 2, \ldots, N_C$) are the covariance of the statistics and

$$b_{kk}^l = \frac{\sigma_\omega^2}{N_C} \sum_{k=1}^{N_C} \| H_l^k \|^{-2} \qquad (13)$$

For the rectangular M-ary QAM constellation with Gray coding, the virtual part is independent of the real part for the components of the statistic vector. Each component is decided independently based on the minimum distance criterion. Thus,

$$P_b \approx 4 \frac{\sqrt{M} - 1}{\sqrt{M} \log_2 M} Q(2 b_{kk}^{l-1/2}) \qquad (14)$$

where $Q(\cdot)$ is the Gaussian tail function. The total BER of the system is the average BER of all N_T layers; that is

$$P_b \approx \frac{4(\sqrt{M} - 1)}{\sqrt{M} N_T \log_2 M} \sum_{l=1}^{N_T} Q(2 b_{kk}^{l-1/2}) \qquad (15)$$

2.2 IIC algorithm for MIMO SC-FDE detection

The proposed IIC algorithm treats all transmitted signals as interference except for the desired stream from the target antenna. Further, the ILI is reconstructed based on the intermediate detection results. Therefore, the reconstructed ILI are subtracted from the received and remaining signals with the reduced interference achieving a larger diversity order. The IIC algorithm (see Fig. 2) is described in detail in the following. First, linear equalization is performed for each sub-channel of the MIMO SC-FDE system, and the initial equalized output vector $\hat{\boldsymbol{X}}_0^k$, $k = 1, 2, \ldots, N_C$ is obtained; that is

$$\hat{\boldsymbol{X}}_0^k = \boldsymbol{G}^k \boldsymbol{Y}^k \qquad k = 1, 2, \ldots, N_C \qquad (16)$$

Subscript 0 represents the initial value, and subscript N_{LP} indicates the N_{LP}-loop iteration, the same as below. If the MMSE is adopted, the equalization matrix $\boldsymbol{G}^k =$

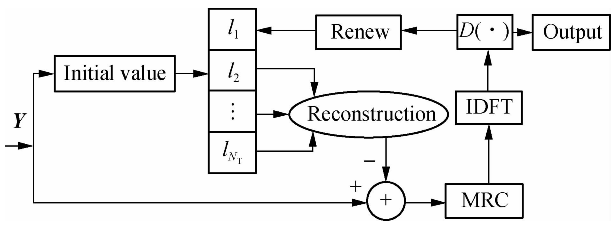


Fig. 2 Proposed IIC for MIMO SC-FDE/MA systems

$\left[(\boldsymbol{H}^k)^H\boldsymbol{H}^k+\frac{\sigma_\omega^2}{\sigma_x^2}\boldsymbol{I}_{N_T}\right]^{-1}(\boldsymbol{H}^k)^H$, where $\boldsymbol{G}^k\in\boldsymbol{C}^{N_T\times N_R}$, and if ZF is adopted, the equalization matrix $\boldsymbol{G}^k=[(\boldsymbol{H}^k)^H\boldsymbol{H}^k]^{-1}(\boldsymbol{H}^k)^H$. Here, superscripts in $(\cdot)^H$ represent the conjugate transpose. All the initial equalized output vectors can be written as matrix $\hat{\boldsymbol{X}}_0$.

$$\begin{aligned}\hat{\boldsymbol{X}}_0&=[\hat{\boldsymbol{X}}_0^1\quad \hat{\boldsymbol{X}}_0^2\quad \dots\quad \hat{\boldsymbol{X}}_0^k\quad \dots\quad \hat{\boldsymbol{X}}_0^{N_C}]=\\&[\hat{\boldsymbol{X}}_{0,1}\quad \hat{\boldsymbol{X}}_{0,2}\quad \dots\quad \hat{\boldsymbol{X}}_{0,l}\quad \dots\quad \hat{\boldsymbol{X}}_{0,N_T}]^T\end{aligned}\tag{17}$$

where $\hat{\boldsymbol{X}}_{0,l}$ represents the first l-th layer initial equalized output vector and its dimension is N_C, $l=1,2,\dots,N_T$. Further, $\hat{\boldsymbol{X}}_{0l}$ denotes the launched symbol of the l-th layer, according to Eqs. (3) to (6)

$$\hat{\boldsymbol{x}}_{0,l}=D(\boldsymbol{F}_{N_C}^{-1}(\hat{\boldsymbol{X}}_{0,l}))\qquad l=1,2,\dots,N_T\tag{18}$$

Then, take the l-th layer update as an example to describe the iterative update process, and transform $\hat{\boldsymbol{X}}_{0m}$ by DFT to obtain the estimated value in the frequency domain of the corresponding m layer signal; that is

$$\begin{aligned}\overline{\boldsymbol{X}}_{0m}&=\boldsymbol{F}_{N_C}(\hat{\boldsymbol{x}}_{0m})\qquad m=1,2,\dots,N_T;\ m\neq l\\ \overline{\boldsymbol{X}}_{0m}&=[\overline{\boldsymbol{X}}_{0m}^1\quad \overline{\boldsymbol{X}}_{0m}^2\quad \dots\quad \overline{\boldsymbol{X}}_{0m}^{N_C}]\end{aligned}\tag{19}$$

The k-th component $\overline{\boldsymbol{X}}_{0m}^k$ represents the estimated value of the k-th sub-channel of the m-th layer. According to $\overline{\boldsymbol{X}}_{0m}$, the reconstruction symbol in the frequency domain of the m-th layer is expressed as

$$\overline{\boldsymbol{E}}_m^k=\overline{\boldsymbol{X}}_{0m}^k\boldsymbol{H}_m^k\qquad k=1,2,\dots,N_C\tag{20}$$

Ignoring the noise, $\overline{\boldsymbol{E}}_m^k$ denotes the receiving contribution of the m-th component to the receiving $\boldsymbol{Y}^k$, and $\boldsymbol{H}_m^k$ is the m-th column of the k-th sub-channel. Finally, the interference cancellation of the l-th layer is expressed as

$$\widetilde{\boldsymbol{E}}_l^k=\boldsymbol{Y}^k-\sum_{m=1,m\neq l}^{N_T}\overline{\boldsymbol{E}}_m^k\tag{21}$$

where $\widetilde{\boldsymbol{E}}_l^k$ is a vector with $N_R\times1$ dimension, which can be regarded as the N_R receiving copies of symbol X_l^k adding the noise. The new estimated value $\hat{X}_l^k$ of X_l^k is achieved when MRC is adopted.

$$\hat{X}_{1,l}^k=(\boldsymbol{H}_l^k)^+\widetilde{\boldsymbol{E}}_l^k\qquad k=1,2,\dots,N_C\tag{22}$$

$\hat{\boldsymbol{X}}_{1,l}=\{\hat{X}_{1,l}^1,\hat{X}_{1,l}^2,\dots,\hat{X}_{1,l}^{N_C}\}$ is reorganized into a frame. According to Eqs. (3) to (6), the new detection result of the l-th layer is achieved after the DFT transformation and decision.

$$\hat{\boldsymbol{x}}_{1,l}=D(\boldsymbol{F}_{N_C}^{-1}(\hat{\boldsymbol{X}}_{1,l}))\qquad l=1,2,\dots,N_T\tag{23}$$

If the initial detection $\hat{\boldsymbol{x}}_{0l}$ is not too bad, the error propagation (EP) caused by $\hat{\boldsymbol{x}}_{0l}$ can be offset by the diversity gain (see Eq. (22)). Compared with the initial estimate value $\hat{\boldsymbol{x}}_{0l}$, $\hat{\boldsymbol{x}}_{1,l}$ possesses a slightly higher accuracy rate. Then, the reconstruction signal (see Eq. (20)) of the l-th layer is replaced by

$$\overline{\boldsymbol{E}}_m^k=\overline{\boldsymbol{X}}_{1,m}^k\boldsymbol{H}_m^k\qquad k=1,2,\dots,N_C\tag{24}$$

A slightly better result is achieved, and the BER decreases after the iterative update process. The above process is performed per layer until the updated result of all N_T layers is obtained. If all the N_T layer estimation results are updated once, this is called a 1-loop iteration. To achieve a balance between performance and complexity, several parameters should be considered when selecting iteration loops, such as modulation order M, antenna dimension $N_T\times N_R$, and frame length N_C.

3 Computational Complexity and Simulations Analysis

3.1 Computational complexity analysis

In this study, the computational complexity is measured by the number of complex multiplications and complex additions per block. The computational complexity calculation is divided into two parts: coefficients and equalization. According to NP-BI[11], the forward frequency-domain MMSE equalization, which includes one-time FFT, N times $\boldsymbol{C}^{N_T\times N_R}\times\boldsymbol{C}^{N_R\times1}$ matrix multiplications, and one-time IFFT, requires $N[N_TN_R+0.5(N_T+N_R)\log_2N]$ complex multiplications and $N[N_TN_R+(N_T+N_R)\log_2N]$ complex additions. Then, the forward frequency-domain MMSE equalizer $G^k(k=1,2,\dots,N)$ costs $N(N_R^3+N_T^3N_R+N_T^4)$ complex multiplications and $N(N_R^3+N_T^3N_R+N_T^4)$ complex additions.

The complexity for a loop of the IIC equalization is one-time FFT, one-time IFFT, one-time $\hat{\boldsymbol{E}}_l^k(k=1,2,\dots,N_C;\ l=1,2,\dots,N_T)$ calculation, one-time MRC, and interference elimination. Hence, N_{LP} loop iteration requires $N_{LP}(N_TN\log_2N+NN_TN_R+NN_TN_R)$ complex multiplications and $N_{LP}(2N_TN\log_2N+NN_TN_R+2NN_TN_R)$ complex additions. The complexity of the IIC coefficient calculations includes $\hat{\boldsymbol{E}}_l^k(k=1,2,\dots,N_C;\ l=l_2,l_3,\dots,l_{N_T})$, MRC coefficients $(\boldsymbol{H}_l^k)^+$, and one-time FFT. Thus, it also requires $N(0.5N_T\log_2N+3N_RN_T-N_R)$ complex multiplications and $N(N_T\log_2N+2N_RN_T-N_R)$ complex additions. Further, the total complexity of the proposed IIC requires $N[(N_T+0.5N_R)\log_2N+4N_TN_R-N_R+N_R^3+N_T^2N_R+N_T^3]+N_{LP}NN_T(\log_2N+2N_R)$ complex multiplications and $N[(2N_T+N_R)\log_2N+3N_TN_R-N_R+N_R^3+N_T^2N_R+N_T^3]+N_{LP}NN_T(2\log_2N+3N_R)$ complex additions.

Fig. 3 shows the variation trend of computational complexity with an increase in the antenna scale. Here, N_{LP}

=4 and N_C = 1 024. It can be summarized that the linear algorithm MMSE has the smallest amount of computation, while NP-BI and FDE-NP and the proposed IIC have similar calculation complexity. With an increase in the number of transmitting and receiving antennas, the operational complexity of the V-BLAST algorithm increases significantly and exceeds that of NP-BI, FDE-NP, and IIC.

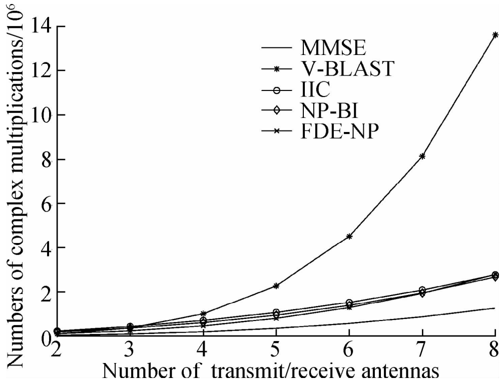


Fig. 3 Calculation complexity comparison

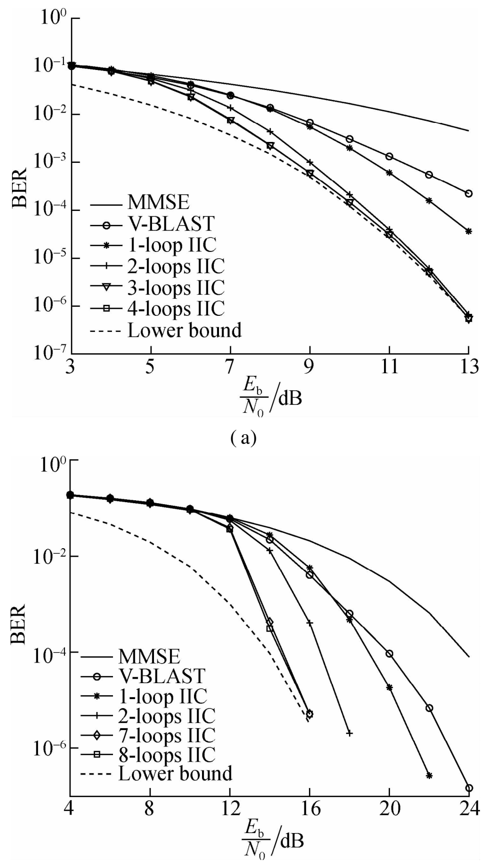
In addition, the proposed IIC has a parallel scheme in which multiple layers can renew simultaneously in a loop (at most, half of all layers). Therefore, the practical computational efficiency can be further enhanced, especially in large antenna scale systems.

3.2 Simulations analysis

The performance of the proposed IIC algorithm is evaluated through simulations. Typical urban channel model COST 259 is employed in simulations. The sampling rate is 2×10^7 samples per second, the frame length N_C = 1 024, CP length is 64, modulation modes are 4-ary quadrature amplitude modulation (4QAM), 16-ary quadrature amplitude modulation (16QAM), and 64-ary quadrature amplitude modulation (64QAM). Iterative loops N_{LP} for 64QAM is 8, for 16QAM is 6, and for 4QAM is 4.

Fig. 4 shows the influence of iterative loops on the performance of IIC algorithms, Fig. 4 (a) is for 4QAM, and Fig. 4(b) is for 16QAM. The compared algorithms are MMSE, V-BLAST[13], and the IIC algorithm of 1-loop, 2-loops, (N_{LP} − 1)-loops, and N_{LP}-loops, where V-BLASTs are based on the MMSE. The simulations show clearly that the BER performance of the IIC algorithm enhances with an increase in the loops of iteration, especially in the initial few loops.

The performance enhancement comes from the efficient mitigation of the EP, which is introduced by incorrect detection results and hinders the acquisition of maximal diversity gain. However, as the loop of iteration increases, IIC improvements tend to saturate. At a higher SNR, the curves of the proposed IIC are close, and then at 9 dB for 4QAM and 16 dB for 16QAM, they approximately



(a)

(b)

Fig. 4 Uncoded BER performance for MIMO SC-FDE systems. (a) 4QAM; (b) 16QAM

approach their RDLBs.

The performances of several detection algorithms are compared in Fig. 5, where $N_T = N_R = 4$, and the modulation mode is 64QAM. Evidently, the proposed IIC is robust against high modulation order and performs much better than other algorithms.

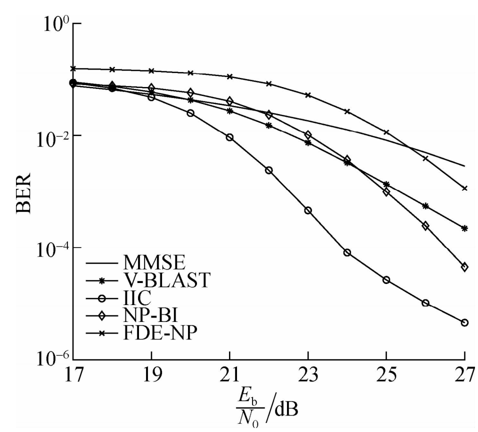


Fig. 5 Uncoded BER performance for MIMO SC-FDE systems

4 Conclusions

1) IIC, an iteration algorithm, was proposed for the detection of the MIMO SC-FDE system and compared with NP-BI and FDE-NP. Simulation results revealed that

the proposed IIC algorithm has a high performance, especially in systems with large-scale antennas and/or high modulation order.

2) Then, the RDLB of the BER performance for MIMO SC-FDE systems has been derived in the typical broadband wireless propagation environments characterized by multipath Rayleigh fading, which is facilitated to evaluate and compare the BER performance of detection algorithms since ML class algorithms are impractical due to their prohibitive complexity.

3) The computational complexity of the proposed IIC has been further analyzed. The computational complexity of an iterative loop is linear to the number of transmit antennas. Hence, IIC is computationally efficient and practical even in large-dimension MIMO systems.

4) The proposed IIC algorithm and derived RDLB are also suitable for MIMO SC-FDMA systems.

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用于 MIMO SC-FDE 系统逼近分集接收界的迭代检测算法

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摘要：为了提高 MIMO SC-FDE 系统的检测可靠性、降低系统复杂度，提出一种新的迭代干扰消除检测算法，并推导出分集接收界作为评价算法性能优劣的依据. 该检测算法采用基本的线性均衡算法提供初始值，在迭代中尽量消除层间干扰，利用空间分集增益逐步提高检测可靠性. 结果表明，该算法在 4QAM 调制方式下，E_b/N_0 为 9 dB 时逼近分集接收界；在 16QAM 调制方式下，E_b/N_0 为 16 dB 时逼近分集接收界. 该算法一轮迭代的计算复杂度与发射天线的数目成线性关系，天线规模为 3 × 3 以上时算法复杂度低于 V-BLAST. 因此，所提算法适用于天线规模数大于等于 3、调制进制数大于等于 4 的系统中，在倾向于采用 large MIMO 系统的 5G 及后 5G 时代具有高效性和实用性.

关键词：分集接收界；空间复用系统；MIMO SC-FDE；迭代检测算法

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