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A Joint Synchronization and Channel Estimation Algorithm for MIMO-OFDM System

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Abstract

As conventional methods deal with channel estimation and synchronization separately, they can't achieve the best performances because the unknown parameters will interfere with each other. Besides, the convergence speed is slow using expectation maximization algorithm. Thus in this paper, we propose a joint synchronization and channel estimation method in MIMO-OFDM systems. The algorithm achieves coarse estimation by using a preamble. Then by combining this algorithm with SAGE algorithm and using joint iteration technology, time-frequency synchronization and channel estimation is achieved. Consequently, the performance of the system is enhanced. Theoretical researches and simulation results show that the algorithm can obtain relatively perfect result of time-frequency synchronization and channel estimation with fewer iterative times. Meanwhile our algorithm can approach the ideal case of perfect channel estimation and synchronization with lower complexity.

Keywords: *MIMO, OFDM, Synchronization, Channel Estimation, SAGE algorithm*

1. Introduction

Orthogonal frequency division multiplexing (OFDM) [1], as a multi-carrier modulation technique for wireless high data-rate transmission, is effective for supporting high speed transmission as well as combating multipath fading in broadband wireless communications. Multiple-input multiple-output (MIMO) technology is usually divided into two main types: spatial multiplexing and diversity coding [2]. Spatial multiplexing improve the performance of the communication system by applying coding over the different transmitter branches, whereas spatial diversity is focused on improving the quality of the transmitted signal. The combination of MIMO and OFDM [3], which increases diversity gain and enhances system capacity in a multipath fading environment, is a promising wireless access key technology and it has been adopted in the fourth generation communication systems.

Two critical issues in the design of a MIMO-OFDM communication system are synchronization and channel estimation. Similar to single-input single-output OFDM systems, MIMO-OFDM systems are very sensitive to synchronization errors, especially the carrier frequency offset (CFO). Inaccurate carrier frequency offset correction results in loss of orthogonality among subcarriers and produces inter-carrier interference (ICI) as well as multiple-access interference with ensuing limitations of system performance. However, when the channel state information is not perfect such as the large delay spreading in frequency-selective fading channels, it will result in considerable performance degradation. In order to detect the transmitted signal from each transmit antenna with data detection, we should estimate an accurate channel state information (CSI) at the receiver. Therefore, synchronization and channel estimation are challenging tasks in a mobile MIMO-OFDM communication system due to the existence of multiple carrier frequency offsets and transmission channels.

A good synchronization scheme is necessary to make an MIMO-OFDM system practical. A blind technique of [4] proposed for single-input multiple-output (SIMO) antenna system can be extended to MIMO systems. But this approach requires too long a delay due to iteration over a large number of OFDM Symbols. Therefore it is not suitable for delay-sensitive

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applications. In addition, this technique requires that there are more receive antennas than transmit antennas and the received signals measured off the multiple receive antennas are subject to independent fading. High complexity and the lack of convergence analysis are also the concerns in the approach of [4]. Training-based algorithms have been proposed by either using the cyclic prefix and the orthogonal polyphase sequences can be modulated directly or using only the cyclic prefix. The method in [5] have low complexity. But the use of training sequences for carrier frequency offset recovery significantly reduces achievable data rate. As to the method in [6], its performance degrades significantly as frequency selectivity becomes more severe. Thus it is imperative to develop a fast-converging carrier frequency and phase offset estimator.

In a MIMO-OFDM system, a coherent signal detection requires a reliable estimate of the channel impulse responses between transmit antennas and receive antennas. There are three classes of channel estimation approaches. First, the training-based methods [7], such as least square (LS), maximum likelihood (ML) and maximum mean square error (MMSE) algorithms, employ known training signals to render an accurate channel estimation. Blind channel estimation algorithms, such as those proposed in [8], which exploit the second-order stationary statistics and other properties, have a better spectral efficiency. Combining the advantages of both the training-based algorithms and the blind algorithms, a semi-blind channel estimation technique has been proposed in [9]. With a small number of training pilots, problems such as slow convergence of the blind methods can be solved. Simultaneously, by using statistical information maintained high accuracy of the training-based methods, a semi-blind estimation scheme would be an appropriate choice herein.

Aiming for fast and accurate synchronization as well as channel estimation, various techniques have been proposed. In [10], Pham proposes a joint frequency offset and channel estimation method which has better performance than the methods in [11-12] and its mean square error (MSE) performance can reach the Cramer Rao Bound (CRB) if no timing error exist. But when only small timing offsets occur, its estimation performance will deteriorate. A joint small timing offset and channel estimation method is proposed in [13] without considering the frequency offsets. In [14], Minn et al. presented a joint fine timing synchronization and channel estimation scheme based on Zadoff-Chu sequences. The joint scheme relies on a threshold to identify the first significant tap within the estimated channel impulse response (CIR) and to adjust the fine timing instant accordingly. However, the threshold is determined in a trial-and-error manner. Wang and Wang [15] proposed a joint fine timing synchronization and channel estimation scheme along with optimal and suboptimal thresholds for OFDM systems. The optimal threshold reveals how the threshold should be selected in different scenarios mathematically. Furthermore, the suboptimal threshold requires no pre-simulation or prior information on the channel. However, the optimal and suboptimal thresholds derived in [15] cannot be directly applied to the MIMO-OFDM system. The Expectation maximization (EM) algorithm is known to be good for tracking the CSI and it is used for the channel estimation and synchronization in MIMO-OFDM systems [16]. The EM algorithm is an iterative method to approximate the maximum likelihood (ML) estimation when a direct computation is computationally limited. But it has a disadvantage of slow convergence with the increase in the number of transmit antennas and receive antennas. The Space-Alternating Generalized Expectation maximization (SAGE) algorithm [17] has been proposed for accelerating the convergence of the EM algorithm by updating only subset of the parameters.

This study proposes a new approach to a joint synchronization and channel estimation method for a MIMO-OFDM system. The proposed algorithm has low complexity, low latency, and good tracking ability. The synchronization and channel estimation can be done within one OFDM symbol time. The rest of the paper is organized as follows. Firstly, the system model, the channel model and preamble structure are introduced. Secondly, the proposed synchronization and channel estimation algorithm is presented in detail. Once more the simulation results are obtained. Finally, some conclusions are drawn in this section.

2. System Description

2.1. System Model

For an $Nt \times Nr$ MIMO-OFDM system with N subcarriers, the system structure is shown in Figure 1. Such a system consists of Nt Antennas at the transmitter and Nr Antennas at the receiver. At the time n, a data block is coded into Nt different symbol blocks, by an appropriate coding scheme. $x_i^n = [x_i^n(0), x_i^n(1), \cdots, x_i^n(N-1)]^T$, for $i = 1, \cdots, Nt$. The coded vector x_i^n is modulated by an inverse discrete Fourier transform (IDFT) into an OFDM symbol sequence. Then, a cyclic prefix (CP) of length Ng is inserted in the corresponding time-domain output vector to combat the dispersive channel.

At the receiver, the received signal vector first has the CP removed and is then demodulated by a discrete Fourier transform (DFT) to yield the demodulated signal vectors. The received signal after OFDM demodulation can be expressed as:

$$Y^n = X^n H^n + W^n \tag{1}$$

Where $X^n = [X_1^n, X_2^n, \cdots, X_{Nt}^n]$, X_i^n is a $N \times N$ diagonal matrix with $X_i^n(k,k) = x_i^n(k)$, $H^n = [H_1^n, H_2^n, \cdots, H_{Nr}^n]$ and $H_j^n = [H_{j,1}^n, H_{j,2}^n, \cdots, H_{j,Nt}^n]^T$. $H_{j,i}^n$ is an $N \times 1$ vector with $H_{j,i}^n(k)$ denoting the frequency response of channel at subcarrier k from the transmit antenna i to the receive antenna j, W^n is an independent and identically distributed complex-valued Gaussian matrix with zero mean. Noise sequences at different receive antennas are statistically independent.

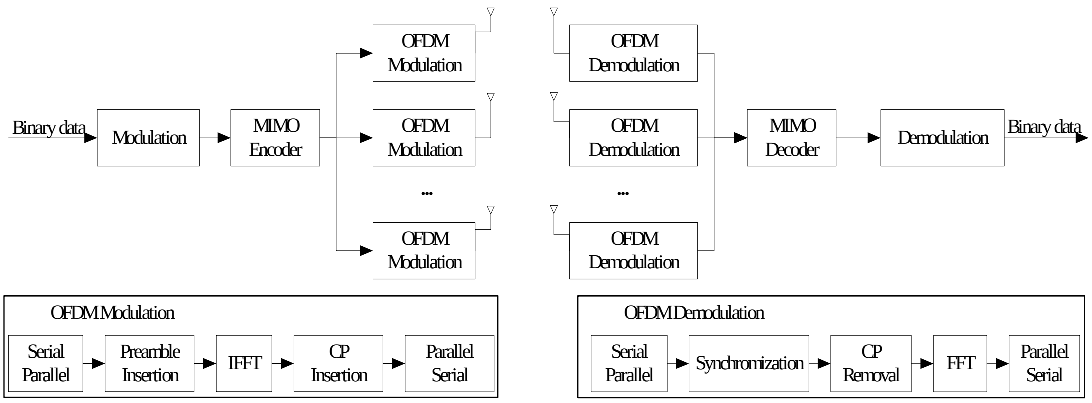

Figure 1. Block Diagram of a MIMO-OFDM System Model

2.2. Channel Model

The complex baseband representation of the mobile wireless channel impulse response (CIR) can be described by:

$$h_{j,i}^n(t,\tau) = \sum_{l=0}^{L-1} \alpha_{j,i}^l(t)\delta(\tau - \tau_l) \tag{2}$$

Where $\alpha_{j,i}^l(t)$ denotes the possibly time-variant attenuation factor for the L multipath propagation paths, which is wide-sense stationary narrow-band complex Gaussian processes. For different l, i and j, $\alpha_{j,i}^l(t)$ are independent. τ_l is the corresponding time delay. The channel frequency response for MIMO-OFDM system can be expressed as:

$$H_{j,i}^n = D \cdot h_{j,i}^n \tag{3}$$

Where $H_{j,i}^n = [H_{j,i}^n(0), \cdots, H_{j,i}^n(k), \cdots, H_{j,i}^n(K-1)]$, D is a matrix of dimension with entries $[D]_{k,l} = \exp[-j2\pi(k-1)(l-1)/N]$, for $0 \le k \le N-1, 0 \le l \le L-1$, $h_{j,i}^n = [h_{j,i}^n(0), \cdots, h_{j,i}^n(l), \cdots, h_{j,i}^n(L-1)]^T$.

2.3. Preamble Structure

In order to implement frequency synchronization and channel estimation algorithms in MIMO-OFDM systems conveniently, there are one preamble symbol and sixteen data symbols in a frame in MIMO-OFDM systems per transmitted antenna. The Constant Amplitude Zero Auto Correlation (CAZAC) sequence has been utilized to generate the preamble in MIMO-OFDM systems. These sequences are characterized by their uniform amplitudes and the autocorrelation function which is represented by a delta function, while the cross-correlation function has a small value.

Then the CAZAC sequence $C = \{C_0 C_1 \cdots, C_{L-1}\}$ is given as follows:

$$C_k = \begin{cases} \exp(j\pi k(k+1)p/L) & L \text{ is odd} \\ \exp(j\pi k^2 p/L) & L \text{ is even} \end{cases} \tag{4}$$

Where $k = 0,1,\cdots,L-1$.

3. Joint Synchronization and Channel Estimation Algorithms

3.1. SAGE-based Iterative Estimator

The SAGE algorithm procedure of the MIMO-OFDM system can be described as follows. Consider the received data Y as incomplete data, and define the complete data Z_q as:

$$Z_q = \Gamma(w_q) \cdot F^H \cdot X_q \cdot D \cdot h_q + w_q, q = 1,2,\cdots, Nt \tag{5}$$

Where $\Gamma(\omega_q) = diag\{1, \exp(j2\pi\omega_q),\cdots,\exp(j2\pi(N-1)\omega_q/N)\}$, $(\cdot)^H$ is Hermitian transposition and $Y = \sum_{q=1}^{Nt} Z_q$, $W = \sum_{q=1}^{Nt} w_q$, $h_q = [h_q(0), h_q(1),\cdots, h_q(L-1)]^T$ is the q th receive antenna's discrete-time composite channel impulse response of order L. F is the Fourier Transform matrix with entry $[F]_{i,k} = \exp[-j2\pi(i-1)(k-1)/N]$, for $0 \le i,k \le N-1$. $X_q = diag\{X_q(0), X_q(1),\cdots, X_q(N-1)\}$.

The SAGE algorithm starts with an arbitrary initial guess $h_q^{(0)}$ and $\omega_q^{(0)}$. Let $\theta^{(m)} = [\theta_1^{(m)},\cdots,\theta_q^{(m)},\cdots,\theta_{Nt}^{(m)}]$ be the estimated value of θ obtained after the $m-1$ iteration, where $\theta_q^{(m)} = [h_q^{(m)}, \omega_q^{(m)}]^T$ .As for Nt users, θ is divided into Nt groups. When updating one group of θ_q, the other groups remain fixed. The k th iteration contains Nt times the following expectation step (E-step) and maximization step (M-step).

E-step: Compute the conditional expectation of the log-likelihood function of the θ_q's hidden data space Z_q.

$$Q(\theta_q;\theta^{(m-1)}) = Q(\theta_q;\theta_q^{(m-1)},\{\theta_l^{(m-1)}\}_{l\ne m}) = E\{\log f(z_q;\theta_q,\{\theta_l^{(m-1)}\}_{l\ne q}) \big| y;\theta^{(m-1)}\} \tag{6}$$

Where:

$$f(z_q;\theta_q,\{\theta_l^{(m-1)}\}_{l\ne q}) = f(z_q;\theta_q) = \frac{1}{(\pi\sigma^2)^N}\exp\{-\frac{1}{\sigma^2}\| z_q - \Gamma(\omega_q)F^H X_q D h_q \|^2\} \tag{7}$$

Substituting Equation (7) into Equation (6), we obtain:

$$Q(\theta_k;\theta^{(m-1)}) = C - \frac{1}{\sigma^2}\| z_q^{(m)} - \Gamma(\omega_q^{(m-1)})F^H X_q D h_q^{(m-1)} \|^2 \tag{8}$$

Wwhere is C a constant, and:

$$z_q^{(m)} = E\{z_q \big| y;\theta^{(m-1)}\} = y - \sum_{l=1,l\neq q}^{Nt} \Gamma(\omega_q^{(m-1)}) F^H X_q D h_q^{(m-1)} \tag{9}$$

M-step: Find θ_q by maximizing Equation (8).

$$\theta_q^{(m)} = \arg\max_{\theta_q} Q(\theta_q;\theta^{(m-1)}) = \arg\min_{\theta_q}\{\| z_q^{(m)} - \Gamma(\omega_q^{(m-1)}) F^H X_q D h_q^{(m-1)} \|^2\} \tag{10}$$

To simplify the optimization in Eq. (10), we first use the ML estimate, then the channel vector is given by:

$$h_q^{(m)} = [(\Gamma(\omega_q^{(m-1)}) F^H X_q D)^H \cdot (\Gamma(\omega_q^{(m-1)}) F^H X_q D)]^{-1} \cdot (\Gamma(\omega_q^{(m-1)}) F^H X_q D) \cdot z_q^{(m)} \tag{11}$$

Substituting Equation (11) into Equation (10), the estimate of can be obtained as:

$$\omega_q^{(m)} = \arg\min_{\omega_q}\{\| z_q^{(m)} - \Gamma(\omega_q) F^H X_q D h_q^{(m)} \|^2\} = \arg\max_{\omega_q}\{z_q^{(m)} \Gamma(\omega_q) F^H X_q D h_q^{(m)}\} \tag{12}$$

To tackle the nonlinearity of (12), we resort to Taylor's series expansion of $e^{-j\frac{2\pi n\omega_q}{N}}$ around $\omega_q^{(k-1)}$ to the second-order term as:

$$\begin{aligned} e^{-j\frac{2\pi n w_q}{N}} &= e^{-j\frac{2\pi n w_q^{(m-1)}}{N}} + (w_q - w_q^{(m-1)}) \cdot (-j\frac{2\pi n w}{N}) \cdot e^{-j\frac{2\pi n w_q^{(m-1)}}{N}} \\ &\quad + \frac{1}{2} \cdot (w_q - w_q^{(m-1)})^2 \cdot (-j\frac{2\pi n w}{N})^2 e^{-j\frac{2\pi n w_q^{(m-1)}}{N}} \end{aligned} \tag{13}$$

After some algebraic manipulations, the update of $\omega_q^{(m)}$ in Eq. (12) is shown to be equivalent to the following equation:

$$\omega_q^{(m)} = \omega_q^{(m-1)} - \frac{\mathrm{Im}\{(z_q^{(m)})^H Q \Gamma(\omega_q^{(m-1)}) F^H X_q D h_q^{(m)}\}}{\mathrm{Re}\{(z_q^{(m)})^H Q^2 \Gamma(\omega_q^{(m-1)}) F^H X_q D h_q^{(m)}\}} \tag{14}$$

Where $Q = diag\{0,1,\cdots,N-1\}$.

The iteration stops when the difference between log-likelihood function of the two consecutive iterations is less than minimum values of the specified process or the number of cycles is greater than the set number.

3.2. Initialization Strategies

In order to begin the iteration, the initial estimates of all parameters are required for proper initial conditions increase the convergence speed of the SAGE algorithm. The initial estimate $\omega_q^{(0)}$ can be performed sequentially in the following steps.

In this section we focus on the joint estimation of ω_q and h_q, assuming that the remaining time offset has been absorbed into the channel vector except the initial time offsets estimation. CP is a replica of the data part in the OFDM symbol. It implies that CP and the corresponding data part will share their similarities that can be used for symbol time offset estimation. Time synchronization can be performed by correlating the received samples which are at a distance of the number of subcarriers N, over a length of Ng window. Time synchronization can be estimated by:

$$\tau = \arg\max_{z}\left\{\frac{\sum_{j=1}^{Nt}\sum_{n=0}^{Ng-1}\left(\left(y_j^{z+n}\right)^* \cdot y_j^{z+n+N}\right)}{\frac{1}{2}\sum_{j=1}^{Nt}\sum_{n=0}^{Ng-1}\left(\left|y_j^{z+n}\right|^2 + \left|y_j^{z+n+N}\right|^2\right)}\right\} \tag{15}$$

Carrier frequency offset can be found from the phase angle of the product of CP and the corresponding rear part of an OFDM symbol. In order to reduce the noise effect, its average can be taken over the samples in a CP interval. Fractional frequency offset can be estimated jointly as:

$$\omega_f = \frac{1}{2\pi}\arg\left\{\sum_{j=1}^{Nt}\sum_{n=0}^{Ng-1}\left(\left(y_j^n\right)^* \cdot y_j^{n+N}\right)\right\} \tag{16}$$

Where the argument operation $\arg(\bullet)$ is performed by using $\tan^{-1}(\bullet)$, the range of frequency offset estimation is $[-0.5, 0.5]$ so that $\left|\omega_f\right| < 0.5$ and consequently, integral frequency offset cannot be estimated by this technique.

The integral frequency offset can be estimated in frequency domain by the preamble. According to the Equation (4), CAZAC sequence has the following characteristics:

$$C_{2k} = s_k \cdot C_{2k+1}, k = 0, 1, \cdots, N/2-1 \tag{17}$$

Where s_k is the kth chip of a sequence with length of $N/2$.

After the fractional frequency offset is removed from the received sample stream by multiplying it with $e^{-j\frac{2\pi n\omega_f}{N}}$ per antenna, the received preamble signal of frequency domain is:

$$\hat{R}_j^k = FFT\{y_j^n \cdot e^{-j\frac{2\pi n\omega_f}{N}}\} \tag{18}$$

The integral frequency offset can be calculated by finding ε_i to maximize in Equation (19).

$$\omega_i = \arg\max_{\varepsilon}\frac{\sum_{j=1}^{Nt}\left\{\left|\sum_{k=1}^{N/2-1}\hat{R}_j^{\varepsilon+2k+d}\cdot\left(s_k \cdot \hat{R}_j^{\varepsilon+2k+1+d}\right)^*\right|^2\right\}}{\sum_{j=1}^{Nt}\left\{\sum_{k=0}^{N/2-1}\left|\hat{R}_j^{2k+d}\right|^2 \cdot \sum_{k=0}^{N/2-1}\left|\hat{R}_j^{2k+1+d}\right|^2\right\}} \tag{19}$$

Thus the initial estimate $\omega_q^{(0)}$ can be derived:

$$\omega_q^{(0)} = \omega_i + \omega_f \tag{20}$$

Using the LS estimate method, $h_q^{(0)}$ can be given by:

$$[h_1^{(0)}, \cdots, h_q^{(0)}, \cdots, h_{Nt}^{(0)}] = (F^H X^H X F)^{-1}\Gamma(\omega_q^{(0)}) F^H X^H Y \tag{21}$$

4. Simulation Results

The performance of the proposed synchronization and channel estimation algorithm for a MIMO-OFDM system is put into investigation by computer simulation. The system parameters are assigned as follows: 2x2, 4x4 MIMO-OFDM systems, AWGN channel and Rayleigh multipath fading channel with a delay of L=8, OFDM symbol size K=128 with each subcarriers modulated in 16QAM, preamble symbol using CAZAC sequences, CP length of

K/16, bandwidth of 20 MHz, 15 kHz subcarriers spacing, 1GHz centre frequency of modulation and 50 Hz Doppler frequency shift. The normalized CFO is independently generated with a uniform distribution within [-0.5, 0.5], and the timing offset is independently generated with a discrete uniform distribution within [0, 4]. The iteration stops when the difference between log-likelihood function of the two consecutive iterations is less than 0.001.

Then the performance of our proposed algorithm is presented. The simulations are performed both in the AWGN channel and the fading channel. Figure 2 shows the time synchronization performance of the proposed technique.It is obvious that the proposed method acquisition achieves much better performance in AWGN channel than the fading channel. Due to the deep fading, the performance declines at low SNR in the fading channel. When SNR is greater than 10dB in a MIMO-OFDM system, the correct timing probability is close to 1. It is shown that the proposed technique can achieve a good time synchronization performance in the case of different conditions from the transmit antennas.

Figure 3 shows the mean square error (MSE) of the frequency offset estimator in the above-mentioned simulation parameter channels as a function of average SNR per receiver antenna. Simulation results of the 2×2 system in the AWGN channel are also shown as a comparison. It is shown that MIMO-OFDM systems still perform much better than the SISO-OFDM system. Besides, for the 2×2 system, the MSE of the estimate in the simulation channel is almost identical to the MSE of the estimate in the AWGN channel. Combining the results, we can conclude that the MSE of the frequency offset estimator decreases as the number of receive antennas increases. The MSE for 2×2 system has an about 3dB gain over the SISO system, while the 4×4 system has an about 3dB gain over the 2×2 system. It should be noted that the performance improvement is due to more observation points which compensate for more number of channel coefficients and give some gains.

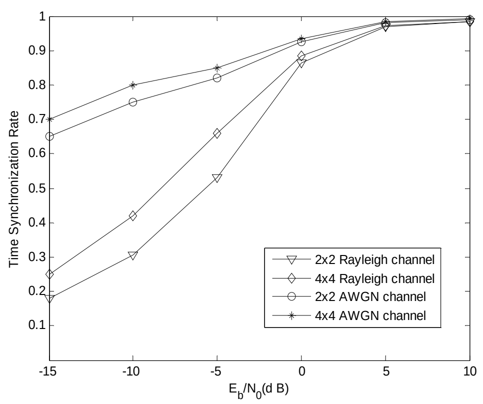

Figure 2. Time Synchronization Performance

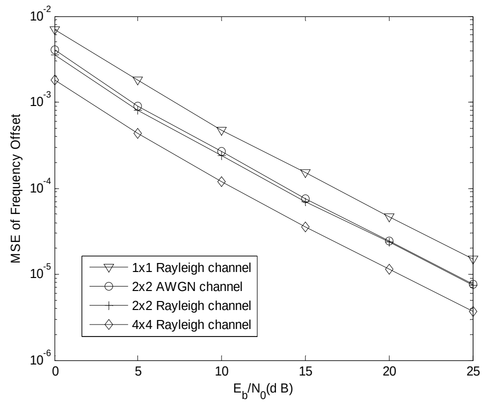

Figure 3. Frequency Offset Estimation Performance

Figure 4 shows the bit error rate (BER) performance of the proposed method with different numbers of iterations used in the joint channel estimation and synchronization for 4×4 MIMO-OFDM system in the Rayleigh fading channel. For reference, the ideal BER performance in the case of perfect channel estimation and synchronization is also plotted. The results show that the performance of the method is improved iteratively and comparable with that of the case of perfect channel estimation and synchronization after five iterations. This is mainly owing to the superior convergence property of SAGE algorithm, which virtually decouples the parameter updates by using a separate hidden data space for each parameter.

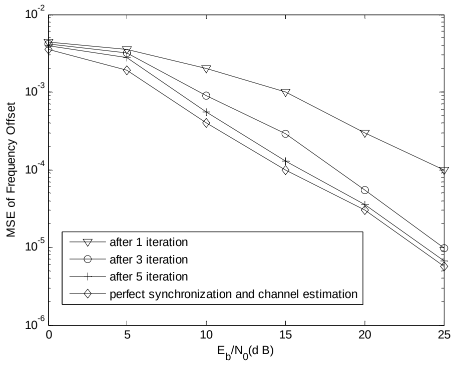

Figure 4. BER Performance of the Proposed Method

5. Conclusion

In this paper, we propose a joint channel, timing and frequency offsets estimation algorithm for MIMO-OFDM system. The synchronization algorithm only needs one preamble transmitted in all transmit antennas in the same OFDM time instant. The preamble can also be cooperated for channel estimation reducing the overall overhead. Each receiver recovers the transmitted data in each iterating making use of suitable channel and frequency offsets estimation as obtained from the previous iteration. The results of theoretical analysis and simulation show that the proposed algorithm of frequency synchronization and channel estimation has excellent performance for the MIMO-OFDM systems in Rayleigh fading channel. Furthermore, it achieves fast convergence within a finite number of iterations.

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