

Performance of STBC-MC-CDMA system based on complex wavelet packet and turbo coding

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Abstract: On the basis of analyzing the principle of the space-time coding technique and the multi-carrier code division multiple access (MC-CDMA) technique, adopting the turbo codes as channel coding and the optimized complex wavelet packet as multi-carrier modulation, a novel space-time block coded the MC-CDMA system based on complex wavelet packet and turbo coding is proposed, and the system bit error rate (BER) performance in the Rayleigh fading channel is investigated. The system can make full use of space-time block codes' transmit diversity and turbo codes' good ability against fading channel to improve the BER performance significantly, and it can also avoid the decrease of spectrum efficiency of conventional MC-CDMA due to inserting cyclic prefix (CP) by utilizing superior characteristics of the optimized complex wavelet packet. Simulation results show that the proposed space-time block coded MC-CDMA system based on the complex wavelet packet performs better than the conventional space-time block coded MC-CDMA (STBC-MC-CDMA) system, and slightly outperforms the STBC-MC-CDMA with CP. Moreover, the application of the space-time block coding technique concatenated with turbo codes strengthens the system ability to combat various interferences in fading channel further.

Key words: multi-carrier technique; complex wavelet packet; space-time block code; code division multiple access; turbo coding

Recently, the space-time coding (STC) technique and the multi-carrier CDMA (MC-CDMA) technique have received much interest because of their high spectrum efficiency and high data transmission rate, and the effective combination of the STC technique and the multi-carrier technique will be a major study hotspot of future wideband mobile communications^[1,2]. However, conventional MC-CDMA is implemented by means of inverse discrete Fourier transform (IDFT) and discrete Fourier transform (DFT) operators. So in its frequency spectrum, the main-lobe energy cannot concentrate effectively and side-lobe attenuates slowly; and the multi-path fading or synchronization error will cause severe performance degradation due to inter-channel interference (ICI) and inter-symbol interference (ISI). For this reason, the multi-carrier system often resorts to cyclic prefix to eliminate the ISI and maintain orthogonality between neighboring sub-carriers, which decreases the efficiency of spectrum utilization considerably in some communication scenarios. To search for an efficient multi-carrier CDMA scheme, a number of improved MC-CDMA systems have been proposed. Among them, a wavelet packet based MC-CDMA^[3] has advantages of stronger ability to combat multi-path interference (MPI) and ISI than conventional MC-CDMA. Especially, the optimized complex wavelet packet^[4] not only has the good properties that the real wavelet packet possesses, such as shifting orthogonality, adaptability, time-frequency localization, etc., but also matches the complex channel frequency spectrum and suits multi-carrier communications. Motivated by the reasons above, we develop an MC-CDMA scheme based on the optimized complex wavelet packet (CWP-MC-CDMA) in Ref. [5], and it obtains better performance than the MC-CDMA based on the real wavelet packet. Considering that the space-time block coding (STBC) technique concatenated with channel codes (especially turbo codes) can significantly improve the system robustness against channel fading and increase the system capacity^[6], we apply the technique to the CWP-MC-CDMA system. On the one hand, the technique can strengthen the system ability to combat spatial fading via obtaining the space diversity gains. On the other hand, it can make use of turbo codes' soft decision information from multiple iterations decoding, random interleaver and good ability to combat the burst error of fading channel to perfect sys-

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tem performance further. Based on this, we propose a space-time coded MC-CDMA scheme based on the optimized complex wavelet packet and turbo coding, and investigate the BER performance in the Rayleigh fading channel.

1 System Model

The two transmitter antennas STBC (i. e. Alamouti code^[7]) can achieve full diversity and full rate with orthogonal design and a simple ML decoding algorithm used at the decoder^[8], and channel coding can effectively improve the system BER performance in fading channels. Especially, turbo coding technology receives great interest due to its superior performance at a lower signal to noise ratio (SNR), and the simulation results in Ref. [6] also show that the turbo coded STBC for two transmitter antennas has better performance than the other compared codes. Based on the above analysis, we develop a new CWP-MC-CDMA scheme concatenating space-time block codes with turbo coding in this section. The conventional DFT/IDFT is replaced by the complex wavelet packet transform (CWPT)/ICWPT (inverse CWPT) accordingly. The transmitter and receiver structures of the proposed system are illustrated in Fig. 1 and Fig. 2, respectively. At the transmitter, the information bits $\{b_k\}$ are first encoded by the turbo encoder, then the encoded bits $\{a_k\}$ are mapped to the QPSK constellation to form complex symbols $\{d_k\}$. The output symbols are mapped into two sub-streams by the two-antenna STBC encoder; each sub-stream is duplicated to different carriers; and the symbol over the different carriers is multiplied by a chip of the spreading code. Then the ICWPT is used to realize multi-carrier modulation. After modulation to RF, the two sub-streams signals are transmitted via two antennas, respectively. At the receiver, the RF signals are converted to baseband signals, then the CWPT is performed on the signal samples for multi-carrier demodulation. After that, the formed data streams are “demapped” by the corresponding STBC decoder, then the output signals are sent into the channel equalizer. After despreading and combining the desired signal at the different sub-carriers, the soft-decided complex values $\{\hat{d}_k\}$ are produced. Then the complex-valued symbols are mapped onto real-valued, soft-decided values $\{\hat{a}_k\}$ via the symbol demapper (i. e., the QPSK demapper), and $\{\hat{a}_k\}$ are fed into the turbo decoder for iterative decoding again. Finally, detected source bits $\{\hat{b}_k\}$ are achieved.

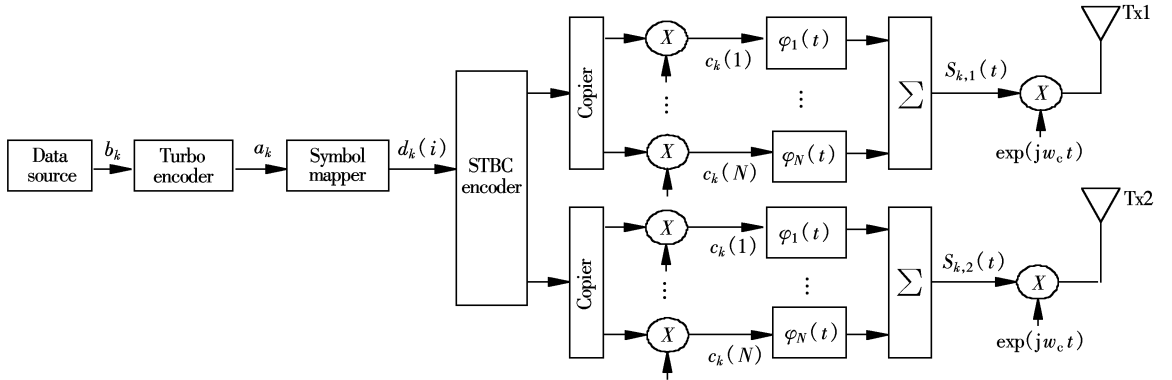


Fig. 1 Transmitter structure of the proposed system

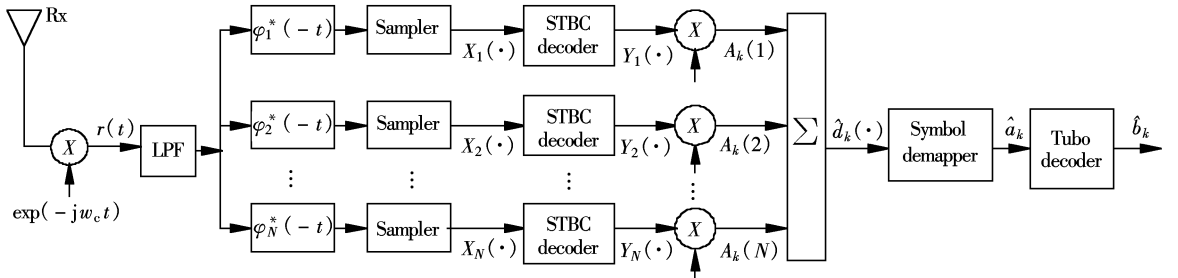


Fig. 2 Receiver structure of the proposed system (“ \cdot ” in (\cdot) denotes time unit, such as $2i+1, 2i+2$, etc.)

According to the STBC encoding principle in Refs. [7, 8], for input signals of user k at two consecutive time instants $2i+1$ and $2i+2$, i. e., $d_k(2i+1)$ and $d_k(2i+2)$, after space-time block encoding, the output signals $d_k(2i+1)$ and $-d_k^*(2i+2)$ are transmitted through the transmitter antenna 1 (Tx1) at time instants $2i+1$ and $2i+2$, re-

spectively; whereas outputs the $d_k(2i+2)$ and $d_k^*(2i+1)$ are transmitted via the transmitter antenna 2 (Tx2) at time instants $2i+1$ and $2i+2$, respectively; where superscript “*” represents complex conjugation. Considering that the present space-time block codes design is based on flat fading, we assume that the wireless channels from each transmitter antenna to receiver antenna experience independent, slow time-varying, frequency selective Rayleigh fading, while every sub-carrier channel is considered to be flat and slow fading; i. e., the fading amplitude and phase remain constant over two consecutive symbol intervals.

2 Performance Analysis

At the transmitter, the complex wavelet packet function $\varphi_n(t)$ is taken as the signature waveform, and the shifting orthogonality among $\varphi_n(t)$ ($n=1, 2, \dots, N$) can be given as $\langle \varphi_n(t), \varphi_m^*(t-iT_s) \rangle = \delta(n-m)\delta(i)^{[4]}$, where \langle, \rangle denotes the inner product and δ is the Kronecker function. $d_k(i)$ corresponding to the QPSK complex signal denotes the i -th data symbol of the k -th user, where $k=1, 2, \dots, K$, K is the active user number; and $\{d_k(i)\}$ are assumed to be independent, identically distributed (i. i. d) random variables taking value $\{\pm 2^{-1/2} \pm j2^{-1/2}\}$ with equal probability. They are from the turbo encoding and the QPSK mapping of transmitted information bits $\{b_k\}$. E_b is the mean energy of the transmitted bit. The symbol period T_s corresponds to the minimum orthogonal shifting defined in the complex wavelet packet. The k -th user spreading code $C_k = \{c_k(n), n=1, 2, \dots, N\}$ is the Walsh-Hadamard code, and the code length equals the number of sub-carriers N .

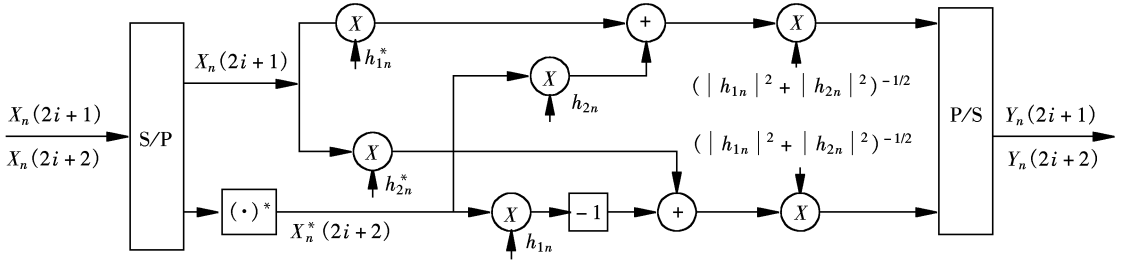


Fig. 3 Block diagram of space-time block decoder

For simplicity of analysis and calculation, channel coding (turbo coding) is not considered first, namely, the performance of the space-time block coded CWP-MC-CDMA (STBC-CWP-MC-CDMA) system is first investigated. As shown in Fig. 1, the transmitted baseband signals of user k through Tx1 and Tx2 are expressed as follows, respectively.

$$S_{k,1}(t) = \sum_{n=1}^N \sum_{i=-\infty}^{+\infty} \sqrt{\frac{2E_b}{N}} \{d_k(2i+1)c_k(n)\varphi_n[t-(2i+1)T_s] - d_k^*(2i+2)c_k(n)\varphi_n[t-(2i+2)T_s]\} \quad (1)$$

$$S_{k,2}(t) = \sum_{n=1}^N \sum_{i=-\infty}^{+\infty} \sqrt{\frac{2E_b}{N}} \{d_k(2i+2)c_k(n)\varphi_n[t-(2i+1)T_s] + d_k^*(2i+1)c_k(n)\varphi_n[t-(2i+2)T_s]\} \quad (2)$$

At the receiver, after the RF signals are converted to baseband, the received signal can be written as

$$r(t) = \sum_{k=1}^K \sum_{n=1}^N \sum_{i=-\infty}^{+\infty} \sqrt{\frac{2E_b}{N}} \{[d_k(2i+1)\varphi_n(t-(2i+1)T_s) - d_k^*(2i+2)\varphi_n(t-(2i+2)T_s)]c_k(n)h_{1,n} + [d_k(2i+2)\varphi_n(t-(2i+1)T_s) + d_k^*(2i+1)\varphi_n(t-(2i+2)T_s)]c_k(n)h_{2,n}\} + W(t) \quad (3)$$

where $h_{l,n}$ ($l=1, 2; n=1, 2, \dots, N$) is the fading coefficient of the n -th sub-carrier from the transmitter antenna l to the receiver antenna, for different l, n , the coefficients $\{h_{l,n}\}$ are modeled as i. i. d complex-valued Gaussian random variables with zero mean and variance σ^2 . And $h_{l,n}$ can be expressed as $h_{l,n} = \alpha_{l,n} \exp(j\varphi_{l,n})$, where the fading amplitude $\{\alpha_{l,n}\}$ are i. i. d Rayleigh variables with second-moment σ^2 and phase $\{\varphi_{l,n}\}$ are i. i. d uniform variables in the interval $[0, 2\pi]$ for different l, n . $W(t)$ is the AWGN for receiver antenna; it is a Gaussian random variable with zero mean and power spectrum density $N_0/2$ per dimension.

After the received signals pass through the low-pass filter (LPF) and the corresponding complex wavelet packet matched filter in sub-channel n , the output signals at the $(2i+1)T_s, (2i+2)T_s$ sampling interval are given by

$$X_n(2i+1) = \int_0^{+\infty} r(t)\varphi_n^*[t-(2i+1)T_s]dt = \sqrt{\frac{2E_b}{N}} \sum_{k=1}^K [d_k(2i+1)h_{1,n} + d_k(2i+2)h_{2,n}]c_k(n) + w_n(2i+1) \quad (4)$$

$$X_n(2i+2) = \int_0^{+\infty} r(t) \varphi_n^* [t - (2i+2)T_s] dt = \sqrt{\frac{2E_b}{N}} \sum_{k=1}^K [-d_k^*(2i+2)h_{1,n} + d_k^*(2i+1)h_{2,n}] c_k(n) + w_n(2i+2) \quad (5)$$

where $w_n(2i+1) = \int_0^{+\infty} W(t) \varphi_n^* [t - (2i+1)T_s] dt$, $w_n(2i+2) = \int_0^{+\infty} W(t) \varphi_n^* [t - (2i+2)T_s] dt$.

The above two noise components (i. e. $w_n(2i+1)$ and $w_n(2i+2)$) are i. i. d with zero mean and variance N_0 . By the use of space-time block decoding methods as shown in Fig. 3, the output signals from the space-time decoder are written as

$$Y_n(2i+1) = \sum_{k=1}^K \sqrt{\frac{2E_b}{N}} d_k(2i+1) c_k(n) \sqrt{|h_{1,n}|^2 + |h_{2,n}|^2} + \frac{[h_{1,n}^* w_n(2i+1) + h_{2,n} w_n^*(2i+2)]}{\sqrt{|h_{1,n}|^2 + |h_{2,n}|^2}} \quad (6)$$

$$Y_n(2i+2) = \sum_{k=1}^K \sqrt{\frac{2E_b}{N}} d_k(2i+2) c_k(n) \sqrt{|h_{1,n}|^2 + |h_{2,n}|^2} + \frac{[h_{2,n}^* w_n(2i+1) - h_{1,n} w_n^*(2i+2)]}{\sqrt{|h_{1,n}|^2 + |h_{2,n}|^2}} \quad (7)$$

In this paper, equal-gain combining (EGC) is considered for sub-channel equalization, so the equalization gain of the n -th sub-carrier for the k -th user $A_k(n) = c_k(n)$ ($1 \leq n \leq N$). Without loss of generality, let user 1 be the desired user, then the decision variable at the $(2i+1)T_s$ interval of user 1 can be given by

$$\hat{d}_1(2i+1) = \sum_{n=1}^N Y_n(2i+1) A_1(n) = \sum_{k=1}^K \sum_{n=1}^N \sqrt{\frac{2E_b}{N}} d_k(2i+1) c_k(n) c_1(n) \sqrt{\alpha_{1,n}^2 + \alpha_{2,n}^2} = D + I + \eta \quad (8)$$

where $D = \sum_{n=1}^N \sqrt{\frac{2E_b}{N}} d_1(2i+1) \sqrt{\alpha_{1,n}^2 + \alpha_{2,n}^2}$ is the desired output; $\alpha_{l,n} = |h_{l,n}|$ ($l = 1, 2; n = 1, 2, \dots, N$); $I = \sum_{k=2}^K \sum_{n=1}^N \sqrt{\frac{2E_b}{N}} d_k(2i+1) c_k(n) c_1(n) \sqrt{\alpha_{1,n}^2 + \alpha_{2,n}^2}$ is the MAI term from other users $k > 1$ with zero mean; $\eta = \sum_{n=1}^N [h_{1,n}^* w_n(2i+1) + h_{2,n} w_n^*(2i+2)] / \sqrt{\alpha_{1,n}^2 + \alpha_{2,n}^2}$ is the Gaussian noise term with zero mean and variance $\text{var}(\eta) = NN_0$.

According to the previous assumption that $\alpha_{l,n}$ is Rayleigh distributed with second-moment σ^2 , we can evaluate $E[\sqrt{\alpha_{1,n}^2 + \alpha_{2,n}^2}]$ as $E[V_n] = 3\sqrt{\pi}\sigma/4$, where $V_n = \sqrt{\alpha_{1,n}^2 + \alpha_{2,n}^2}$. So $E[V_n^2] = 2\sigma^2$.

When the number of subcarriers N is large, the law of large number (LLN) can be employed, thus we have

$$\frac{1}{N} \sum_{n=1}^N \sqrt{\alpha_{1,n}^2 + \alpha_{2,n}^2} \cong E[V_n] = 3\sqrt{\pi} \frac{\sigma}{4} \quad (9)$$

Hence

$$D \cong \frac{3}{2} \sqrt{\frac{\pi N E_b}{2}} d_1(2i+1) \sigma \quad (10)$$

Considering $V_n = \sqrt{\alpha_{1,n}^2 + \alpha_{2,n}^2}$, then I can be changed into

$$I = \sum_{k=2}^K \sum_{n=1}^N \sqrt{\frac{2E_b}{N}} d_k(2i+1) c_k(n) c_1(n) V_n \quad (11)$$

For a large number of users and subcarriers, the interference I can be approximated by a Gaussian random variable according to the central limit theorem (CLT). Thus the variance of I can be evaluated by

$$\begin{aligned} \text{var}(I) &= \frac{2E_b}{N} E \left[\sum_{k=2}^K d_k(2i+1) \sum_{n=1}^N c_k(n) c_1(n) V_n \sum_{s=2}^K d_s^*(2i+1) \sum_{m=1}^N c_s(m) c_1(m) V_m \right] = \\ &= \frac{2E_b}{N} \left\{ \sum_{k=2}^K \sum_{n=1}^N \sum_{m=1, m \neq n}^N c_k(n) c_1(n) c_k(m) c_1(m) E[V_n] E[V_m] + \sum_{k=2}^K \sum_{n=1}^N E[V_n^2] c_k^2(n) c_1^2(n) \right\} = \\ &= \frac{2E_b}{N} \left\{ \sum_{k=2}^K \sum_{n=1}^N \sum_{m=1}^N c_k(n) c_1(n) E[V_n] c_k(m) c_1(m) E[V_m] - \sum_{k=2}^K \sum_{n=1}^N E^2[V_n] c_k^2(n) c_1^2(n) + \sum_{k=2}^K \sum_{n=1}^N E[V_n^2] \right\} = \\ &= \frac{2E_b}{N} \left\{ \sum_{k=2}^K \left[\sum_{n=1}^N c_k(n) c_1(n) E[V_n] \right]^2 + \sum_{k=2}^K \sum_{n=1}^N [E[V_n^2] - (E[V_n])^2] \right\} = \end{aligned}$$

$$\frac{2E_b}{N} \left\{ \sum_{k=2}^K \frac{9}{16} \pi \sigma^2 \rho_{k,1}^2 + (K-1)N \left[2 - \frac{9\pi}{16} \right] \sigma^2 \right\} \quad (12)$$

where $\rho_{k,1} = \sum_{n=1}^N c_k(n) c_1(n)$ represents the correlation between user k and user 1's spreading code. Owing to super-orthogonality of the Walsh-Hadamard code applied, the first term of $\text{var}(I)$ equals zero. Hence,

$$\text{var}(I) = (K-1)E_b\sigma^2 \left(4 - \frac{9\pi}{8} \right) \quad (13)$$

Thus we can calculate the signal to interference ratio (SIR) approximately for the STBC-CWP-MC-CDMA system with 2Tx and 1Rx (receiver antenna) as follows:

$$R_{\text{SIR}} = \frac{|D|^2}{\text{var}(I)} \cong \frac{(9/8)\pi E_b N \sigma^2}{(K-1)E_b\sigma^2[4 - (9\pi/8)]} = \frac{7.589N}{K-1} \quad (14)$$

Similarly, according to Ref. [5], we can calculate the SIR for the 1Tx and the 1Rx-based CWP-MC-CDMA system with EGC as $R_{\text{SIR}} \cong 3.6598N/(K-1)$. Hence, the SIR for the 2Tx and the 1Rx-based STBC-CWP-MC-CDMA is double that of the single antenna CWP-MC-CDMA system. Thus, the 2Tx and the 1Rx-based STBC-CWP-MC-CDMA system achieves full space diversity (i. e. 2-diversity order) and full transmission rate (i. e. the code rate of STBC is 1).

From (10) and (13) as well as $\text{var}(\eta) = NN_0$, the BER expression of channel-uncoded STBC-CWP-MC-CDMA system for large N can be approximated by

$$P_e = \frac{1}{2} \text{erfc} \left\{ \sqrt{\frac{|D|^2}{2[\text{var}(I) + \text{var}(\eta)]}} \right\} \cong \frac{1}{2} \text{erfc} \left\{ \sqrt{\frac{(9/8)\pi N \sigma^2 E_b}{(K-1)(8 - 9\pi/4)\sigma^2 E_b + 2NN_0}} \right\} \quad (15)$$

When using turbo codes as channel coding, the above system BER expression necessitates making corresponding modification. Unfortunately, the BER expression is difficult to achieve due to high calculation complexity. In addition, the data decision will be performed after turbo decoding, not after the channel equalization and combining mentioned above. Refer to Fig. 2 and section 1 for specific principle descriptions, where the turbo encoder and decoder structure are shown in Ref. [9]. In this paper, the turbo encoder employs two 8-state and 1/2-rate identical recursive systematical convolutional (RSC) encoders $(1, 1 + D + D^3/1 + D^2 + D^3)$ connected in parallel with a pseudorandom interleaver preceding the second RSC encoder. Both RSC encoders encode the information bits. The first encoder operates on the input bits in their original order, whereas the second encoder operates on the input bits as permuted by the turbo interleaver. Consider that the coded bits block of each encoder consists of systematic bit part and parity bit part, and the systematic bits are the same in both coded bit streams; hence, the systematic part is transmitted only once. The coded bits are not punctured. We choose the systematic bit from encoder 1 as the transmitted information bit, and add two parity bits, so the total code rate R_c of the turbo encoder equals 1/3, namely, one bit input, then three bits output. The turbo decoding is performed iteratively; the basic principle of iterative decoding is to feed forward/backward the decoder output (hard decisions and reliability information) to improve the next decoding. By multiple iterative runs, the performance can be significantly improved. Finally, the decision variable \hat{b}_k for transmitted bit b_k is obtained. Considering the tradeoff between different decoding algorithm performance and realization complexity, the Log-MAP algorithm is adopted. The number of iterations is four.

3 Simulation Results

In this section, we simulate the performance of the STBC-CWP-MC-CDMA system as well as the STBC-CWP-MC-CDMA system with turbo coding over channel model A defined by ITU-R M. 1225^[10]. The processing gain of spreading code (Walsh-Hadamard code) is equal to the number of sub-carriers N , and EGC is employed for channel equalization. The related simulation parameters are listed as follows: the code bit modulation scheme is QPSK; the carrier frequency $f_c = 2$ GHz; the mobile velocity $v = 50$ km/h; sampling frequency $f_s = 3.84$ MHz; the number of active users is 7 or 13; the bit rate $R_b = 384$ kbit/s. The 32-band complex wavelet packet (corresponding to the five-level wavelet packet) and the 16-band complex wavelet packet^[4] are used for simulation; the turbo parameters are set as in section 2. In simulation, we assume that channel estimation and system synchronization are perfect. The simulation results are shown in Fig. 4.

Fig. 4(a) gives the average bit error probability as a function of E_b/N_0 under the conditions that $K=7$ and $N=$

16. “DFT1T1R” and “DFT2T1R” represent the conventional DFT based MC-CDMA (DFT-MC-CDMA) system with 1Tx/1Rx and 2Tx/1Rx, respectively. “DFT1T1RCP x ” denotes the 1Tx/1Rx-based DFT-MC-CDMA system with CP; “DFT2T1RCP x TC” denotes the 2Tx/1Rx-based DFT-MC-CDMA system with turbo coding and CP; where “ x ” denotes that x cyclic prefix symbols are inserted. “CWP1T1R” and “CWP2T1R” represent the STBC-CWP-MC-CDMA system with 1Tx/1Rx and 2Tx/1Rx, respectively; “CWP2T1RTC” denotes the turbo coded STBC-CWP-MC-CDMA system with 2Tx/1Rx. From Fig. 4(a), we can see that the performance of CWP-MC-CDMA outperforms that of DFT based MC-CDMA system, and is slightly superior to that of the conventional MC-CDMA system with CP. Moreover, the application of STBC can effectively improve the performance of the CWP-MC-CDMA system, for four channel-uncoded systems. STBC-CWP-MC-CDMA with 2Tx and 1Rx has better performance than the other three comparative systems. When turbo code is used, the performance of the STBC-CWP-MC-CDMA system is increased significantly; it can make the above uncoded STBC-CWP-MC-CDMA system with 2Tx/1Rx achieve approximately 3 dB gains at the BER of 10^{-4} .

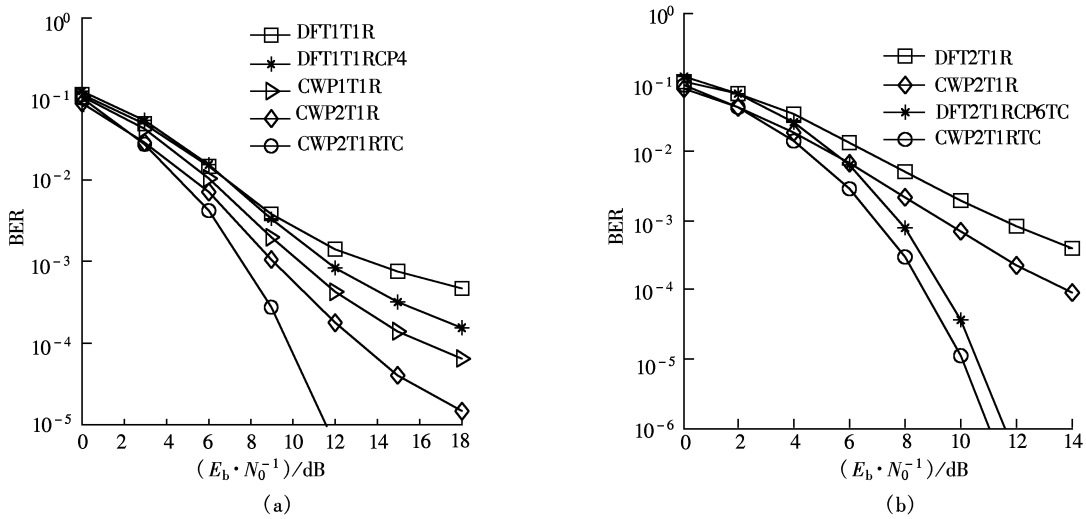


Fig. 4 BER against SNR for different MC systems. (a) $K=7$, $N=16$; (b) $K=13$, $N=32$

In further comparison, we also give the average bit error probability as a function of E_b/N_0 under $K=13$ and $N=32$. It is shown in Fig. 4(b) that the space-time coded CWP-MC-CDMA system performs better than the space-time coded MC-CDMA system, and the STBC-CWP-MC-CDMA system with turbo coding slightly outperforms the space-time block coded DFT-MC-CDMA system with turbo coding and CP. Especially, the 2Tx/1Rx-based STBC-CWP-MC-CDMA system with turbo coding has the best performance, it can get lower BER than the other three MC systems at the same SNR. All these results show that the STBC-CWP-MC-CDMA system with turbo coding has the strongest robustness against different interferences. In addition, we also notice that the application of the complex wavelet packet may increase the realization complexity of the system, but the optimized complex wavelet packet is produced in terms of the conventional real wavelet packet. Thus the corresponding ICWPT/CWPT can take a fast tree structured form^[11,4]. Hence, we apply a fast ICWPT/CWPT algorithm to the proposed system. It decreases the computational complexity greatly. The run time in simulation is almost the same as with corresponding conventional MC systems. Thus the computation efficiency is obviously improved.

4 Conclusion

On the basis of analyzing the principle of the space-time coding technique and the multi-carrier CDMA technique, we combine the two techniques effectively, and develop a space-time block coded MC-CDMA scheme based on the complex wavelet packet and turbo codes. The system can avoid the loss of spectrum efficiency due to inserting CP in the conventional MC-CDMA system, and its performance is close to or superior to that of the conventional MC-CDMA with CP, thus the spectrum efficiency and system performance are obviously increased. Moreover, the CWP-MC-CDMA system with 2Tx and 1Rx can implement full diversity and full transmission rate, and effectively improve system capacity and the ability to avoid various interferences in fading channel. Especially, by utilizing the turbo codes' superior ability to combat burst error from fading channel, and soft-input/soft-output infor-

mation provided by iterative decoding; the STBC-CWP-MC-CDMA system with turbo coding has much lower BER; thus corresponding performance is improved significantly.

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基于复小波包和 turbo 码的 STBC-MC-CDMA 系统及其性能

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摘要:在分析空时编码技术和多载波 CDMA 技术原理的基础上,采用 turbo 编码作为信道编码以及优化的复小波包作为多载波调制,提出一种基于复小波包和 turbo 码的空时分组编码的 MC-CDMA 系统,研究了其在瑞利衰落信道下的误码率性能. 该系统能充分利用空时分组编码的发送分集和 turbo 码的良好抗信道衰落能力显著提高系统性能,而且还能利用优化复小波包的优良特性避免通常 MC-CDMA 系统由于插入循环前缀(CP)所带来频谱效率的降低. 仿真结果表明,基于复小波包的空时分组编码的 MC-CDMA 系统要好于空时分组编码的通常 MC-CDMA (STBC-MC-CDMA) 系统,略好于采用 CP 的 STBC-MC-CDMA 系统;级联 turbo 码的空时编码技术的应用进一步增强了系统抗衰落信道下的各种干扰能力.

关键词:多载波技术;复小波包;空时分组码;码分多址;turbo 编码

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