A Comparative Study of Two Receiver Schemes for Interleaved OFDMA Uplink

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Abstract—**This paper presents a comparative study of two receiver schemes for interleaved OFDMA uplink. In the first scheme, called the post-DFT processing based interference cancellation receiver, the CFOs are compensated in frequency domain and an iterative interference-cancellation scheme is used to reduce the inter carrier interference (ICI) and/or multiuser interference (MUI). The other scheme, the signal structure-based MMSE receiver, exploits the signal inner structure on the interleaved OFDMA uplink and separates single user signal waveform by compensates the effective carrier frequency offsets. Simulation results show that the signal structure-based MMSE receiver is robust to large CFOs but its computation complexity increases considerably with the number of users. Furthermore, it assumes to know the signal power. When the CFOs are small, the post-DFT processing based interference cancellation receiver is preferable because its computation is easy with the number of users and it needs not know signal power.**

Keywords-orthogonal frequency division multiple access; uplink, carrier frequency offset compensation;

I. INTRODUCTION

Recently, much research has been devoted to orthogonal frequency division multiplexing (OFDM) in broadband wireless communications for its high spectral efficiency and ability in mitigating the effects of multipath propagation. To accommodate the variable data rates and Quality of Service (QoS) requirements of multimedia communication, OFDM, with clusters of subcarriers allocated to different users (normally referred to as OFDMA), has attracted much attention.

In OFDMA uplink, sub-band based subcarrier assignment scheme has been studied in [1-3], in which signals from different users occupy non-overlapping frequency bands in a similar fashion as traditional FDMA. Guard subcarriers are suggested in [2][3] to be put in the edge of each sub-band such that the multiple access interference can be minimized. Signals from different users can thus be separated by filter banks and existing synchronization algorithms for OFDM is applicable for the signal on each sub-band. Interleaved carrier assignment scheme is superior over the sub-band based scheme in that it provides maximum frequency diversity in frequency selective fading environment.

In OFDM, there are two synchronization methods for uplink receiver, namely compensation and feedback. The first one compensates the effect of carrier frequency offset on the received signal to recover the ideal waveform which will be received when the transmitter and the receiver are frequency synchronized. In the other method, the estimated carrier frequency offset is fed back to the transmitter side for adjustment. In OFDMA, however, the feedback method will introduce traffic overhead and is not applicable in an environment involving fast changing Dopper shift.

In the following, we will tackle compensation-based synchronization schemes for interleaved OFDMA uplink receiver. In [4], the CFOs are compensated at the BS for each user using the single-user detector, where the sampled sequence is compensated for in the time domain, and then discrete Fourier transform (DFT)-processed for each user. As a result, multiple DFT blocks, each for one user, are used. Performance degradation in single-user detector is caused by the fact that a frequency compensation for one user before the DFT can increase the frequency offsets in the data of other users within the DFT block. Reference [5] used a post-DFT processing technique by employing the fact that time domain multiplication is equivalent to frequency-domain circular convolution to reduce and was shown to outperform the single-user detector in that a frequency offset compensation for one user after the DFT does not affect the data the other users. To significantly reduce the ICI and/or MUI, an iterative interference-cancellation scheme was proposed to be used in an OFDMA system. In [6], the original signals are tentatively restored using the post-FFT processing method in [5] before the detection process. In [7], a signal structure-based minimum mean squared error (MMSE) receiver algorithm was proposed by exploiting the signal algebraic structure on the uplink of OFDMA system adopting interleaved carrier assignment scheme.

The rest of this paper is organized as follows. In section II, the system model for interleaved OFDMA uplink is presented. The single user detector, the post-FFT processing scheme and the post-FFT-processing-based interference cancellation scheme are presented in Section III. In section IV, the signal inner structure for interleaved OFDMA uplink and the signal structure-based MMSE receiver algorithm are presented. Further discussions are presented in section V. Simulation results and performance comparisons are reported in section VI. Conclusions are drawn in Section VII.

II. SYSTEM MODEL

Consider an interleaved OFDMA system with K users and N subcarriers, including all data bearing subcarriers, pilot

The research is mainly funded by Nature Science Fund of China (NSFC), Project NO.60302025 and partially by National 863 project: FuTURE (2003AA12331004).

subcarreirs and virtual subcarriers. All subcarriers are sequentially indexed with $\{0,1,\dots,N\text{-}1\}$. The N subcarreirs are divided into Q sub-channel in an interleaved fashion with $P=N/Q$ subcarriers in each sub-channel. Sub-channel $\{q\}$, $q=0,1,\dots Q\text{-}1$, is composed of subcarreirs with index set $\{q,Q+q,\dots,(P\text{-}1)Q+q\}$.

Without loss of generality, we assume that each user only occupies one sub-channel. Let user k occupies sub-channel $\{q^{(k)}\}$ and $\left[X_0^{(k)},X_1^{(k)},\dots,X_{P-1}^{(k)}\right]^T$ be the P modulation symbols it transmits on its P subcarriers in one OFDMA block. The subscript $[\bullet]^T$ denotes vector or matrix transpose and $(\bullet)^{(k)}$ represents the ownership of the kth user. Also let $\left[H_0^{(k)},H_1^{(k)},\dots,H_{P-1}^{(k)}\right]^T$ denote the channel frequency responses between the kth user and the base station on sub-channel $\{q^{(k)}\}$.

Assume that only the kth user sends signal on subchannel $\{q^{(k)}\}$ during one OFDMA block and $\xi^{(k)}$ is the normalized carrier frequency offset of the received signal from the kth user ($|\xi^{(k)}|<0.5$). $\xi^{(k)}$ is defined as $\xi^{(k)}=\Delta f^{(k)}/\Delta f$, where $\Delta f^{(k)}$ is the carrier frequency offset in Hz and Δf is subcarrier spacing. Hence, after the removal of cyclic prefix, the OFDMA signal samples received by the base station is given by

$$\Upsilon^{(k)}(n)=e^{j2\pi\varphi^{(k)}}e^{j2\pi\xi^{(k)}n/N}\cdot\sum_{p=0}^{P-1}X_p^{(k)}H_p^{(k)}e^{j2\pi\left(pQ+q^{(k)}\right)\left(n-n_d^{(k)}\right)/N}\qquad(1)$$

for $n=0,1,\dots,N\text{-}1$, where $n_d^{(k)}$ is the time offset between the base station and the kth user; $\varphi^{(k)}$ is a common phase shift due to previous OFDMA blocks. For simplification, in the following we assume $n_d^{(k)}=0$ and $\varphi^{(k)}=0$:

$$\begin{aligned}\Upsilon^{(k)}(n)&=e^{j2\pi\xi^{(k)}n/N}\cdot\sum_{p=0}^{P-1}X_p^{(k)}H_p^{(k)}e^{j2\pi\left(pQ+q^{(k)}\right)n/N}\\&=e^{j2\pi\xi^{(k)}n/N}\cdot y^{(k)}(n)\end{aligned}\qquad(2)$$

III. THE POST-DFT PROCESSING-BASED INTERFERENCE CANCELLATION RECEIVER

The received baseband signal observed in additive noise is given by

$$\begin{aligned}r(n)&=\sum_{k=1}^{K}\Upsilon^{(k)}(n)+z(n)\\&=\sum_{k=1}^{K}y^{(k)}(n)e^{j2\pi\xi^{(k)}n/N}+z(n)\end{aligned},$$

for $n=0,1,..,N\text{-}1$ (3)

In the single-user detector for the kth user, the received sequence $r(n)$ is multiplied by a time domain sequence $\exp(-j2\pi\xi^{(k)}n/N)$ before the DFT processing. After the multiplication, the signal at the kth branch is then given by

$$\begin{aligned}y_{\sin}^{(k)}(n)&=r(n)e^{-j2\pi\xi^{(k)}n/N}\\&=y^{(k)}(n)+\sum_{\substack{m=1\\m\neq k}}^{K}y^{(m)}(n)e^{j2\pi\left(\xi^{(m)}-\xi^{(k)}\right)n/N}\\&\quad+z(n)e^{-j2\pi\xi^{(k)}n/N}\end{aligned}\qquad(4)$$

The first term in (4) is the signal for the kth user, the second term is MUI, and the third term is the additive noise.

In the single-user detector, one DFT block is needed for each user to detect the information symbols. To reduce the required number of DFT blocks, the CFOs can be compensated for in the frequency domain. In the scheme [5], after the DFT processing of $r(n)$ in (4) , the received signal is as follows:

$$\begin{aligned}R(m)&=\mathrm{DFT}\left(\sum_{k=1}^{K}y^{(k)}(n)e^{j2\pi n\xi^{(k)}/N}+z(n)\right)\\&=\sum_{k=1}^{K}Y^{(k)}(m)\otimes C^{(k)}(m)+Z(m)\end{aligned}\qquad(5)$$

where $\otimes$ denotes circular convolution, $Y^{(k)}(m)=\mathrm{DFT}_N(y^{(k)}(n))$, $C^{(k)}(m)=\mathrm{DFT}_N(\exp(j2\pi\xi^{(k)}n/N))$, $Z(m)=\mathrm{DFT}_N(z(n))$.

For convenience, we can rewrite (5) into a vector form as follows:

$$\begin{aligned}\mathbf{R}&=\sum_{k=1}^{K}\mathbf{Y}^{(k)}\otimes\mathbf{C}^{(k)}+\mathbf{Z}\\&=\mathbf{Y}^{(k)}\otimes\mathbf{C}^{(k)}+\sum_{\substack{m=1\\m\neq k}}^{K}\mathbf{Y}^{(m)}\otimes\mathbf{C}^{(m)}+Z\end{aligned}\qquad(6)$$

where $\mathbf{R}=[R(0),R(1),\dots,R(N\text{-}1)]^T$, $\mathbf{Y}^{(k)}=[Y^{(k)}(0), Y^{(k)}(1),\dots,Y^{(k)}(N\text{-}1)]^T$, $\mathbf{C}^{(k)}=[C^{(k)}(0),C^{(k)}(1),\dots,C^{(k)}(N\text{-}1)]^T$, $\mathbf{Z}=[Z(0), Z(1),\dots,Z(N\text{-}1)]^T$. The kth user's received signal can then be represented by an $N\times1$ vector $\mathbf{Y}'^{(k)}=[Y'^{(k)}(0), Y'^{(k)}(1),\dots,Y'^{(k)}(N\text{-}1)]^T$ as follows:

$$Y'^{(k)}=Y^{(k)}\otimes C'^{(k)}\qquad(7)$$

since $\mathbf{C}^{(k)}$ is the frequency-domain representation of $\exp(j2\pi\xi^{(k)}n/N)$, $n=0,1,\dots,N\text{-}1$, restoring $\mathbf{Y}^{(k)}$ from $\mathbf{Y}'^{(k)}$ can be achieved as follows:

$$Y^{(k)}=Y'^{(k)}\otimes C'^{(k)}\qquad(8)$$

where $\mathbf{C}'^{(k)}=[C'^{(k)}(0), C'^{(k)}(1),\dots,C'^{(k)}(N\text{-}1)]^T$ and $C'^{(k)}(m)=\mathrm{DFT}(\exp(-j2\pi\xi^{(k)}n/N))$.

When the CFO values are small compared with the subcarrier spacing, the received kth subscriber's power is mainly concentrated in the prescribed subcarrier positions. We can then use $\mathbf{A}^{(k)}\mathbf{R}$ to replace $\mathbf{Y}'^{(k)}$ in to obtain $\mathbf{Y}^{(k)}$, where $\mathbf{A}^{(k)}$ is a diagonal matrix, and

$$\mathbf{A}^{(k)}(i+1,i+1)=\begin{cases}1, & i\in\Gamma_k\\ 0, & i\notin\Gamma_k\end{cases} \quad (9)$$

where Γ_k is the set of subcarriers assigned to user k.

To further reduce the MUI, the interference cancellation is performed in an iterative fashion:

Initiation: set j=0 and $\hat{\mathbf{Y}}^{(k),j}=\mathbf{A}^{(k)}\left(\left(\mathbf{A}^{(k)}\mathbf{R}\right)\otimes\mathbf{C}'^{(k)}\right)$ for k=1, 2,…,K

Loop: j=j+1

$$\text{Set }\mathbf{Y}^{(k),j}=\mathbf{R}-\sum_{\substack{m=1\\ m\neq k}}^{K}\hat{\mathbf{Y}}^{(m),j-1}\otimes\mathbf{C}^{(m)} \quad (10)$$

for k=1,2,…K

$$\hat{\mathbf{Y}}=\mathbf{A}^{(k)}\left(\left(\mathbf{A}^{(k)}\mathbf{Y}^{(k),j}\right)\otimes\mathbf{C}'^{(k)}\right) \quad (11)$$

for k=1,2,…,K

Go back to *Loop*

IV. Signal structure-based user signal separation

By introducing the effective CFO of the kth user, $\theta^{(k)}$, which is given by

$$\theta^{(k)}=\frac{\xi^{(k)}+q}{Q} \quad (12)$$

Equation (2) can be simplified to

$$\Upsilon^{(k)}(n)=e^{j2\pi\theta^{(k)}n/P}\sum_{p=0}^{P-1}X_p^{(k)}H_p^{(k)}e^{j2\pi p/P} \quad (13)$$

From (12) and (13), we have

$$\Upsilon^{(k)}(n+\mu P)=e^{j2\pi\mu\theta^{(k)}}\Upsilon^{(k)}(n) \quad (14)$$

Equation (14) shows that the signal samples of $\{\gamma^{(k)}(n)\}$, for n=0,1,…,N-1 has a special periodic structure with every P sample. There fore, all these N samples can be arranged into a $Q\times P$ matrix $\mathbf{X}^{(k)}$ as

$$\mathbf{X}^{(k)}=\begin{bmatrix}\Upsilon^{(k)}(0) & \ldots\ \Upsilon^{(k)}(P-1)\\ \Upsilon^{(k)}(P) & \ldots\ \Upsilon^{(k)}(2P-1)\\ \vdots & \ldots\ \vdots\\ \Upsilon^{(k)}(N-P) & \ldots\Upsilon^{(k)}(N-1)\end{bmatrix}_{Q\times P} \quad (15)$$

From (14) and (15), $\mathbf{X}^{(k)}$ can be expressed as

$$\mathbf{X}^{(k)}=\mathbf{v}^{(k)}\mathbf{s}^{(k)} \quad (16)$$

where $\mathbf{v}^{(k)}=\left[1,e^{j2\pi\theta^{(k)}},\ldots,e^{j2\pi(Q-1)\theta^{(k)}}\right]^T$ and $\mathbf{s}^{(k)}$ is the first row of $\mathbf{X}^{(k)}$. From (15), $\mathbf{s}^{(k)}$ is given by

$$\mathbf{s}^{(k)}=\mathbf{u}^{(k)}\odot\left(\Lambda^{(k)}\mathbf{W}\right) \quad (17)$$

where $\odot$ represents Schur product or element by element product. $\mathbf{W}$ is a $P\times P$ IFFT matrix. Also,

$$\Lambda^{(k)}=\left[H_1^{(k)}X_1^{(k)}\ H_2^{(k)}X_2^{(k)}\ \cdots\ H_P^{(k)}X_P^{(k)}\right], \text{ and}$$

$$\mathbf{u}^{(k)}=\left[1\ e^{j2\pi\theta^{(k)}/P}\ \ldots\ e^{j2\pi\theta^{(k)}(P-1)/P}\right].$$

In reality, the signal of one OFDMA block received by the base station is the summation of all users' signals. Let $\{\gamma(n)\}$, for n=0,1,…,N-1 denote the received signals of one OFDMA block at the base station. We have

$$\Upsilon(n)=\sum_{k=1}^{K-1}\Upsilon^{(k)}(n) \quad (18)$$

Arranging $\{\gamma(n)\}$, for n=0,1,…,N-1, into a $Q\times P$ matrix $\mathbf{X}$ in the same fashion as $\mathbf{X}^{(k)}$ and combining (16), (17) and (18), we have a useful structure based matrix representation of one interleaved OFDMA uplink block as

$$\mathbf{X}=\sum_{k=1}^{K}\mathbf{X}^{(k)}=\mathbf{VS}=\mathbf{V}\{\mathbf{U}\odot(\mathbf{\Lambda W})\} \quad (19)$$

where $\mathbf{V}$=[$\mathbf{v}^{(1)}$, $\mathbf{v}^{(2)}$,…, $\mathbf{v}^{(K)}$] and

$$\mathbf{U}=\begin{bmatrix}\mathbf{u}^{(1)}\\ \mathbf{u}^{(2)}\\ \vdots\\ \mathbf{u}^{(K)}\end{bmatrix}\quad \mathbf{\Lambda}=\begin{bmatrix}\Lambda^{(1)}\\ \Lambda^{(2)}\\ \vdots\\ \Lambda^{(K)}\end{bmatrix}\quad \mathbf{S}=\begin{bmatrix}\mathbf{s}^{(1)}\\ \mathbf{s}^{(2)}\\ \vdots\\ \mathbf{s}^{(K)}\end{bmatrix}$$

The matrix form representation of an OFDMA block observed in noise is given by

$$\mathbf{Y}=\mathbf{X}+\mathbf{Z}=\mathbf{V}\{\mathbf{U}\odot(\mathbf{\Lambda W})\}+\mathbf{Z} \quad (20)$$

where $\mathbf{Z}$ is a $Q\times P$ AWGN noise matrix with zero mean and variance σ^2.

Let A=$\mathbf{\Lambda W}$ and $\boldsymbol{y}_l$, $\boldsymbol{u}_l$, $\boldsymbol{a}_l$ and $\boldsymbol{z}_l$ denote the lth column of $\mathbf{Y}$, $\mathbf{U}$, $\mathbf{A}$ and $\mathbf{Z}$, l=1,2,…,P. We have $y_l=\mathbf{VD}_l a_l+z_l$. Where $\mathbf{D}_l$ =Diag[$\boldsymbol{u}_l$]. $\boldsymbol{a}_l$ is a $K\times 1$ vector and its kth element is given by

$$\sum_{p=0}^{P-1}X_p^{(k)}H_p^{(k)}e^{j2\pi(l-1)p/P} \quad (21)$$

The MMSE filter applied to the received signal vector $\boldsymbol{y}_l$ is the linear $K\times Q$ matrix transformation **:**

$$\overline{\mathbf{C}_l} = \left[\mathbf{V}\mathbf{V}^{\mathrm{H}} + (\sigma/\delta)^2 \mathbf{I}\right]^{-1} \mathbf{V}\mathbf{D}_l \quad (22)$$

Hence the recovered lth column of $\mathbf{A}$ is

$$\hat{a}_l = \mathbf{D}_l^{\mathrm{H}} \mathbf{V}^{\mathrm{H}} \left[\mathbf{V}\mathbf{V}^{\mathrm{H}} + \sigma^2 \mathbf{I}\right]^{-1} \left(\mathbf{V}\mathbf{D}_l a_l + z_l\right) \quad (23)$$

V. DISCUSSIONS

A. CFOs Estimation

All the schemes require the knowledge of the CFO values, which can be obtained using training signals. Because the received signal power for a specific user is mainly concentrated in the prescribed subcarrier positions. By exploiting this fact, the training symbols for different users can be transmitted simultaneously. For example, we can use two OFDM symbols to estimate the CFO values [8]. In this case, we use $\{Y_1(m)|k\in\Gamma_k\}$ and $\{Y_2(m)|k\in\Gamma_k\}$to denote the received signals at the kth user subcarrier positions of the first and the second received training OFDM symbols respectively. The CFO values for the kth user can be obtained as follows:

$$\hat{\xi}^{(k)} = \frac{1}{2\pi} \tan^{-1} \frac{\sum_{m\in\Gamma_k} \mathrm{Im}\left(Y_2(m) Y_1^*(m)\right)}{\sum_{m\in\Gamma_k} \mathrm{Re}\left(Y_2(m) Y_1^*(m)\right)} \quad (24)$$

B. Computation Complexity

In the post-DFT based interference cancellation scheme, its complexity can be reduced, since most of the elements to be convolved are zero. In (10) and (11), both $\hat{\mathbf{Y}}^{(k),j}$ and $A^{(k)}Y^{(k),j}$ only have N/Q nonzero elements. Furthermore, most elements in $C^{(k)}$ and $C'^{(k)}$ are quite small and can be taker as zeros. For example, the $(m+1)$th element of $C'^{(k)}$ is given by

$$C_m'^{(k)} = \frac{\sin\left\{\pi\left(m+\xi^{(k)}\right)\right\}}{\sin\left\{\frac{\pi\left(m+\xi^{(k)}\right)}{N}\right\}} e^{-\frac{j\pi(N-1)\left(m+\xi^{(k)}\right)}{N}} \quad (25)$$

As a result, as long as CFOs are small, the elements in the middle of $C'^{(k)}$ are close to zero. We can replace $C'^{(k)}$ by using the following vector:

$$\left[C_0'^{(k)}, C_1'^{(k)}, \ldots, C_{(M-1)/2}'^{(k)}, 0, \ldots, 0, C_{N-(M-1)/2}'^{(k)}, \ldots, C_{N-1}'^{(k)}\right]^T$$

where M is the number of nonzero elements in $C'^{(k)}$. It is assumed that M is an odd number. Similarly, we can use

$$\left[C_0^{(k)}, C_1^{(k)}, \ldots, C_{(M-1)/2}^{(k)}, 0, \ldots, 0, C_{N-(M-1)/2}^{(k)}, \ldots, C_{N-1}^{(k)}\right]^T$$

to replace $C^{(k)}$. However, when the CFOs are large, a lot of the elements in $C^{(k)}$ and $C'^{(k)}$ cannot be neglected, and the complexity of the post-DFT processing based interference cancellation scheme cannot be significantly reduced. Table 1 compares the complexity of two receiver schemes for certain values of N, P and K.

TABLE I. NUMBER OF COMPLEX MULTIPLICATIONS REQUIRED FOR RECEIVER SCHEMES

	Type	*2 users*	*4 users*	*8 users*
N=256	Single user detector	2560	5120	10240
	Post-DFT processing (M=N)	68608	70656	74752
	Interference cancellation (1 iteration and M=N)	166912	152576	148480
	Signal structure-based	1942	3224	2223424
	Type	***4 users***	***8 users***	***16 users***
N=1024	Single user detector	24576	49152	98304
	Post-DFT processing (M=N)	1074176	1094656	1135616
	Interference cancellation (1 iteration and M=N)	2384896	2274304	2249728
	Signal structure-based	12696	2238656	1e+05

VI. SIMULATION RESULTS AND PERFORMANCE COMPARISONS

The transmission parameters of the OFDMA system in our simulation are as follows. The uplink bandwidth (BW) is 16MHz and the FFT size is N=1024. The sampling frequency Fs is 16MHz, hence the subcarrier spacing, Δf, is 15.625 kHz and sample interval Ts is 0.0625us. The CP is composed of 256 samples. In our simulations, the 1024 subcarriers are divided into Q=16 subchannels and the number of active users is 100% system capacity, i.e., K=16.

All user data symbols were independent QPSK symbols. The SNR of the kth user is defined as

$$\mathrm{SNR}^{(k)} = \frac{\mathrm{E}\left[\left|\Upsilon^{(k)}(n)\right|^2\right]}{\sigma^2} \quad (26)$$

All $\mathrm{SNR}^{(k)}$, k=1,2,…,K-1, are assumed to be same at the uplink receiver and each user only occupies one sub-channel in one OFDMA block. The channel model adopted is ITU vehicular channel A which is composed of six paths with path delays of 0, 4, 11, 17, 27, 40 samples, and all users' channel are assumed to be statistically independent and perfectly known at the BS.

The average symbol error rate obtained for the post-DFT processing based interference cancellation algorithm and the signal structure based MMSE receiver when the CFOs are in [-0.1 0.1], [-0.2 0.2] and [-0.3 0.3] are shown in Fig. 1, Fig. 2 and Fig. 3 respectively. It can be seen from Fig. 1 that when the CFOs is small, both the interference cancellation receiver and the signal structure based MMSE receiver are desirable. Fig. 2 and Fig. 3 show that when the CFOs is large, the signal structure based MMSE receiver is much better than the post-DFT processing based interference cancellation receiver, especially for high SNR values. For the post-DFT processing interference cancellation receiver, an error floor appears for large CFOs.

VII. Conclusions

In this paper, two different receiver schemes for the interleaved OFDMA uplink are compared. The first one, the post-DFT processing based interference cancellation receiver, generated the interference after the discrete Fourier transform processing, and removed them from the original received signal. Its computation is low when the number of users is large. The signal structure based receiver is robust to larger carrier frequency offsets, but its computation increases considerably with the increase of the number of users. Both algorithms show great potential for OFDMA systems using the interleaved subcarrier assignment in the uplink to reap frequency diversity gains in frequency selective fading channels.

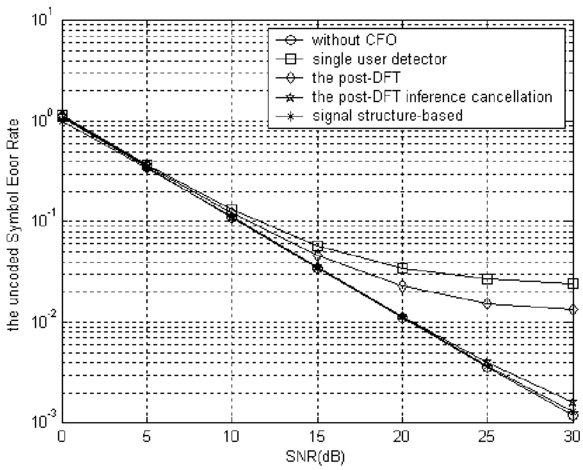

Figure 1. SER performance of receivers with the CFOs are in [-0.1 0.1]

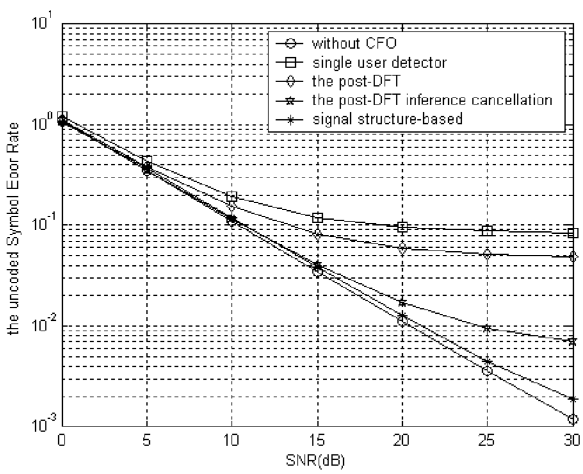

Figure 2. SER performance of receivers with the CFOs are in [-0.2 0.2]

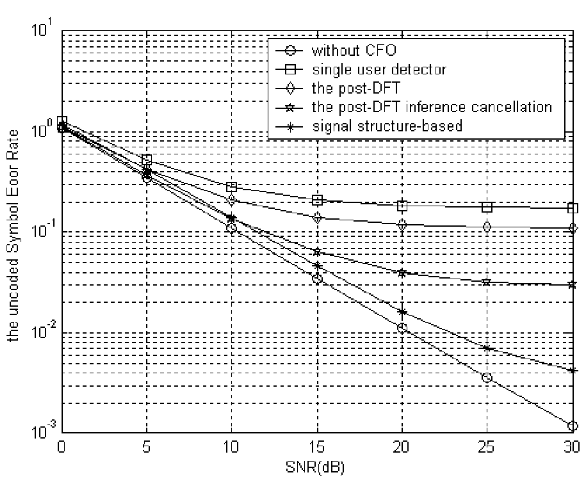

Figure 3. SER performance of receivers with the CFOs are in [-0.3 0.3]

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