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RESEARCH ARTICLE

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A method of ICI cancellation for OFDM with delay diversity

Abstract In orthogonal frequency division multiplexing (OFDM) system, the time variation of a wireless channel destroys orthogonality among the sub-carriers, and this induces inter-carriers interference (ICI) and degrades system performance severely in mobile environment. In this paper, a new method of ICI cancellation based on delay diversity (DD) was proposed, which provides a way to mitigate the negative effect from the time variation of the wireless paths, thus improve the system performance greatly. The new method was called time-domain self-interference cancellation (TDSIC) algorithm, which is different from other existing methods, such as frequency-domain method. In a cyclically extended OFDM system, the fading characteristics of extended OFDM symbols with different cyclic delay are different with each other, so in our TDSIC method, a new diversity collection scheme at the receiver end is proposed, which can be used to improve the system performance by suppressing ICI through selecting appropriate parameters. Moreover, the cyclically extended OFDM symbol at the transmitter side and diversity collection with different delay added OFDM symbols at the receiver side are used in the TDSIC method with the tradeoff of time-expense, so the well-known fixed delay for symbol at the transmitter side may be detected by the receiver side through estimating several parameters of wireless channels. In summary, the key of the TDSIC method is to improve the system performance with the cost of time. Based on performance analysis, simulation has proved that TDSIC may effectively improve the performance of the time-variant wireless channel.

Keywords inter-carriers interference (ICI), delay diversity (DD), cyclic extended, delay added

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1 Introduction

In orthogonal frequency division multiplexing (OFDM) system, the entire channel bandwidth is divided into many parallel independent sub-channels in which data are transmitted in parallel fashion. In terms of the good spectrum efficiency and being robust to the frequency-selective fading channel, OFDM is a promising application technology in wireless communication. However, because of using parallel transmission in time domain, OFDM causes a long duration in transmission, and this becomes a very sensitive factor to the time variation of the channel in high-speed mobile environment. It is difficult to preserve the orthogonality specially when operating in a time dispersion environment. Temporal variation of a channel destroys orthogonality among sub-carriers and thus degrades the system performance severely. Inter-carriers interference (ICI) cancellation or suppression is a key issue in OFDM system. ICI may be generated from many unideal parameters, for example, frequency offset between transmitters and receivers, timing offset between analog-to-digital conversion and digital-to-analog conversion (ADC & DAC), and Doppler spread from temporal variation of mobile environment. Although ICI caused by the mentioned parameters can be cancelled or suppressed through various methods [1], ICI that originated from channel time variation can only be mitigated because of the instantaneous variations of wireless channels.

Since the time variation of a wireless channel is related with the relative speed and the frequency offset at both ends, it becomes critical to reduce the negative effect with frequency-domain equalization (FDE) [2], time-domain windowing technology [3–5], and self-interference cancellation [6]. In Ref. [7], the minimum mean square error frequency-domain equalization (MMSE-FDE) achieves a lower bit error rate (BER), but the channel state information (CSI) must be provided manually by the designer. In reality, this method is not really effective as it requires a priori knowledge of CSI. In time-domain equalization [4], a window function is proposed in equivalent to the correlative polynomial used in the frequency domain and multiplied by the time-domain signal. In this method, time-domain windowing is still under further investigation into second-order polynomial window to achieve better performance. Time-domain windowing technology may have better performance than frequency correlation coding technology because it prevents error propagation. Single carrier frequency-domain equalizer (SC/FDE) today is recognized as an attractive alternative to combat with OFDM application, which has large channel dispersion in Long Term Evolution (LTE), but the disadvantage of coefficients is involved in this equalizer. All of the mentioned methods have a common point, i.e., too complex in real practice. Thus, first of all, a good method selected should be able to reduce the complex in field. It is important to find an appropriate method in time domain.

Delay diversity was previously used to achieve better frequency diversity with a cost of inducing inter-symbol interference (ISI) and large delay diversity. To avoid these disadvantages, cyclic delay diversity (CDD) was introduced in Ref. [8], where the transmitted signal was not delayed but cyclically shifted instead. CDD could be used for both transmit and receive diversity to improve the performance in either time domain or frequency domain.

In this paper, a method of modified delay diversity (DD), called time-domain self-interference cancellation (TDSIC), is proposed to mitigate ICI in the time-variant channel. The main idea of TDSIC is given as follows: Since cyclically extended symbol is used at the transmitter in OFDM single input single output (SISO) system, these extended OFDM symbols with different cyclic delay added (CDA) can be reconstructed at the receiver side. In case any of these reconstructed OFDM symbols experience different time fading, diversity collection can be used to improve the system ICI performance.

The paper is organized as follows: Transceiver structure of TDSIC OFDM system is introduced first in Sect. 2. In Sect. 3, the algorithm of TDSIC is described and analyzed. Then, simulation and comparison with polynomial cancellation coding (PCC) are presented in Sect. 4. Finally, the conclusion and some suggestion for future research of TDSIC are discussed in Sect. 5.

2 TDSIC transceiver and signal analysis

Discrete-time baseband equivalent model of the TDSIC OFDM SISO transceiver system is shown in Fig. 1. Suppose a transmitted OFDM symbol $\{S(k)\}$ is processed by the IDFT block, then $x(n)$ may be obtained as follows:

$$x(n) = \sqrt{N}\mathrm{IDFT}\{S(k)\} = \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} S(k)\mathrm{e}^{\frac{\mathrm{j}2\pi kn}{N}},$$
$$n = 0,1,\ldots,N-1. \quad (1)$$

Data $x(n)$ is cyclically extended, and cyclic prefix (CP) is added, then the time-domain transmitted OFDM signal $\tilde{x}(n)$ could be obtained and just expressed as follows, assuming that the length of CP is T:

$$\tilde{x}(n) = \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} S(k)\mathrm{e}^{\frac{\mathrm{j}2\pi kn}{N}},$$
$$n = -T,\ldots,0,1,\ldots,N-1,N,\ldots,P-1. \quad (2)$$

The factor $R = P/N$ is defined as cyclic extended ratio (CER), which denotes the symbol on predefined repeated ratio in the OFDM symbol. The parameter P_d refers to the space between repeated symbols in time domain. Since CP does not affect the ICI performance, it is ignored in the latter discussion.

Suppose that a cyclically extended OFDM symbol with CER = 2 passes through digital up conversion (DUC), D/A, and up-modulation blocks; after that these symbols will be coupled into the time-variant channel. At the receiver end, the dataset symbols couples in and then is fed through down-conversion, A/D, and digital down conversion (DDC) blocks. After CP is removed (assuming a perfect time synchronization has already been achieved), the resulted baseband received signal $y(n)$ is shown as follows:

$$y(n) = \frac{1}{\sqrt{N}}\sum_{k=0}^{N-1} S(k)\sum_{i=1}^{M} h_i\mathrm{e}^{\frac{\mathrm{j}\left[2\pi\left(k+\mathrm{fD}_i\right)\left(n-T_i\right)+\mathrm{j}\varphi_i\right]}{N}}$$
$$+\mathrm{Noi}(n)$$

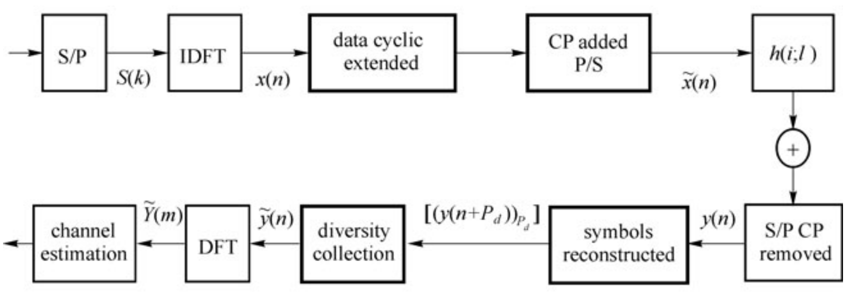

Fig. 1 Discrete-time baseband equivalent model of TDSIC

$$= \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S(k) \sum_{i=1}^{M} h_{k,i} e^{\frac{j2\pi(k+fD_i)n}{N}} + \mathrm{Noi}(n),$$

$$n = 0,1,...,2N-1, \quad (3)$$

where i denotes path index; M is the number of distinguishable multi-paths; fD_i, h_i, T_i, and φ_i denote the normalized Doppler offset, path gain, delay time, and phase shift of the path i, respectively; $h_{k,i}$ denotes the channel impulse response of the sub-carrier k within path i; and $\mathrm{Noi}(n)$ denotes the Gaussian additive noise.

Since $y(n)$ is fed into OFDM symbol reconstruction block, these OFDM symbols with different CDA $[(y(n+P_d))_{P_d}]$ are attained. Data $(y(n+P_d))_{P_d}$, namely, a TDSIC symbol, is added to the DFT block, and then, $Y_d(m)$ is calculated as follows:

$$Y_d(m) = \frac{1}{\sqrt{N}} \mathrm{DFT}\left\{(y(n+P_d))_{P_d}\right\}$$

$$= \sum_{\substack{k=0 \\ k \neq m}}^{N-1} S(k) \sum_{i=1}^{M} H_{k,i}(k-m) F_{i,d}(k-m)$$

$$+ S(m) \sum_{i=1}^{M} H_{k,i}(0) F_{i,d}(0) + N_d(m),$$

$$m = 0,1,...,N-1; \; 0 \leqslant p_d \leqslant N, \quad (4)$$

where d means the dth TDSIC symbol, and P_d denotes the additive delay of the TDSIC symbol. $y(n+P_d)$ is the reconstructed OFDM symbol with CDA P_d, and function $(y(n+P_d))_{P_d}$ denotes a cyclic shift with the symbol $y(n+P_d)$ by P_d. $H_{k,i}(z)$ denotes the inter-carrier fading factor (ICFF) caused by path i, while $F_{i,d}(z)$, an additional fading factor added to the ICFF $H_{k,i}(z)$, denotes the fading characteristic raised by P_d. $H_{k,i}(z)$ and $F_{i,d}(z)$ can be expressed as follows:

$$H_{k,i}(z) = \frac{1}{N} h_{k,i} \sum_{n=0}^{N-1} e^{\frac{j2\pi(z+fD_i)n}{N}}$$

$$= \frac{1}{N} h_i e^{j\phi_k} \sum_{n=0}^{N-1} e^{\frac{j2\pi(z+fD_i)n}{N}}, \quad (5)$$

$$F_{i,d}(z) = e^{\frac{j2\pi(z+fD_i)P_d}{N}}. \quad (6)$$

The first term in Eq. (4) is the ICI due to loss of orthogonality among sub-carriers, the second term is the desired output with a random phase rotation for sub-carriers, and the third term is the Gaussian additive noise.

3 TDSIC algorithm

3.1 Theory of TDSIC

After data has been fed through the symbol reconstruction block, a number of TDSIC symbols could be gathered at the receiver. Supposing a diversity collection symbol $\tilde{y}(n)$ has been fed through the DFT block, we get $\tilde{Y}(m)$, which could be expressed as follows:

$$\tilde{y}(n) = \sum_{d=1}^{D} K_d (y(n+P_d))_{P_d}, \; 0 \leqslant n < N; \; 0 \leqslant P_d \leqslant N, \quad (7)$$

$$\tilde{Y}(m) = \frac{1}{\sqrt{N}} \mathrm{DFT}\left\{\tilde{y}(n)\right\}$$

$$= \sum_{\substack{k=0 \\ k \neq m}}^{N-1} S(k) \sum_{i=1}^{M} H_{k,i}(k-m) \sum_{d=1}^{D} K_d F_{i,d}(k-m)$$

$$+ S(m) \sum_{i=1}^{M} H_{k,i}(0) \sum_{d=1}^{D} K_d F_{i,d}(0) + \tilde{N}(m), \quad (8)$$

where K_d is a weighting factor of the dth TDSIC symbol $(y(n+P_d))_{P_d}$, and D denotes the number of TDSIC symbols for diversity collection.

Considering that $S(k)$ and the phase of $H_{k,i}(z)$ are independent, the signal energy $\tilde{P}_S(m)$ and interference energy ICI power $\tilde{P}_{\mathrm{ICI}}(m)$ in the sub-carrier m can be expressed as follows:

$$\tilde{P}_S(m) = \left|S^2(m)\right| \left| \sum_{i=1}^{M} H_{k,i}(0) \sum_{d=1}^{D} K_d F_{i,d}(0) \right|^2$$

$$\approx \left|S^2(m)\right| \sum_{i=1}^{M} \left|H_i^2(0)\right| \left| \sum_{d=1}^{D} K_d F_{i,d}(0) \right|^2, \quad (9)$$

$$\tilde{P}_{\mathrm{ICI}}(m) = \left| \sum_{\substack{k=0 \\ k \neq m}}^{N-1} S(k) \sum_{i=1}^{M} H_{k,i}(k-m) \sum_{d=1}^{D} K_d F_{i,d}(k-m) \right|^2$$

$$\approx \sum_{\substack{k=0 \\ k \neq m}}^{N-1} \left|S^2(k)\right| \sum_{i=1}^{M} \left|H_i^2(k-m)\right| \left| \sum_{d=1}^{D} K_d F_{i,d}(k-m) \right|^2. \quad (10)$$

Therefore, the cost function $p = \sum \tilde{P}_S(m) / \sum \tilde{P}_{\mathrm{ICI}}(m)$ is derived to show the relation P_d and K_d with p.

$$p = \mathrm{E}[\tilde{P}_S(m)]/\mathrm{E}[\tilde{P}_{\mathrm{ICI}}(m)]$$

$$\approx \frac{\sum_{i=1}^{M}\left|H_i^2(0)\right|\left|\sum_{d=1}^{D}K_dF_{i,d}(0)\right|^2}{\sum_{k-m=1}^{N-1}\sum_{i=1}^{M}\left|H_i^2(k-m)\right|\left|\sum_{d=1}^{D}K_dF_{i,d}(k-m)\right|}. \quad (11)$$

$H_i(z)$ can be expressed as follows:

$$\left|H_i^2(z)\right| = \frac{1}{N^2}h_i^2\left|\sum_{n=0}^{N-1}\mathrm{e}^{\frac{\mathrm{j}2\pi(z+\mathrm{fD}_i)n}{N}}\right|^2. \quad (12)$$

The p, which denotes system average signal-to-interference ratio (SIR), is mainly determined by time-variant channel information state, its probability density function (PDF), and the max Doppler spread. It could also be concluded from Eq. (11) that cost function is not much related with the transmitted data.

Supposing channel state is achieved, thus perfect system performance could be reached with diversity collection being applied with appropriate combinations of P_d and K_d.

3.2 Pursuing P_d and K_d

It is difficult to obtain an optimum combination of P_d and K_d just from Eq. (11) because parameters P_d and K_d are included in the same equation. However, from Eq. (5), there is an approximation of $H_{i,d}(0) \approx 1$ for all M paths when fD_i is small. The result shows that the signal energy is near constant although the interference energy P_{ICI} is variable. When the signal energy P_S is fixed, the minimum P_{ICI} will make the ratio p maximum.

In summary, the suboptimum may be provided to obtain P_d, assuming that all K_d are equal; therefore, an optimum gathering of P_d can be obtained to make p the maximum. Then, in order to obtain K_d, another two cost functions are defined as follows:

$$P_S = \frac{\sum_{m=0}^{N-1}\tilde{P}_S(m)}{\sum_{m=0}^{N-1}\left|S^2(m)\right|} \approx \sum_{i=1}^{M}\left|H_i^2(0)\right|\left|\sum_{d=1}^{D}K_dF_{i,d}(0)\right|^2, \quad (13)$$

$$P_{\mathrm{ICI}} = \frac{\sum_{m=0}^{N-1}\tilde{P}_{\mathrm{ICI}}(m)}{\sum_{m=0}^{N-1}\left|S^2(m)\right|}$$

$$\approx \sum_{k-m=1}^{N-1}\sum_{i=1}^{M}\left|H_i^2(k-m)\right|\left|\sum_{d=1}^{D}K_dF_{i,d}(k-m)\right|^2. \quad (14)$$

To get the maximum p, we need to make P_{ICI} the minimum, while P_S remains constant. Thus, P_d and K_d may be derived from the equations as follows:

$$P_d:\ p = \max(p)|_{K_d=1}, \quad (15)$$

$$\begin{cases}\sum_{i=1}^{M}|h_i|\frac{\sin(\pi\mathrm{fD}_i)}{N\sin(\pi\mathrm{fD}_i/N)}\sum_{d=1}^{D}K_d\mathrm{e}^{\mathrm{j}2\pi\mathrm{fD}_iP_d/N} = 1,\\ \frac{\partial P_{\mathrm{ICI}}}{\partial K_d} = 0.\end{cases} \quad (16)$$

The suboptimum of P_d may be calculated from the equation above when $K_d = 1$, and the vector P_d is determined by Eq. (15). Since fD_i is small, P_S is almost identical with fD_i. P_d in P_{ICI} will provide suboptimum solution for p. In fact, the value of P_d is supplied in space of P_d with iterative method.

To obtain K_d, assuming $\mathrm{E}[S(k)] = 0$, $\mathrm{E}[S^2(k)] = P_S$, $S(k)$ and h_i are both independent factors here, $|h_i|^2 = |h|^2$. From Eqs. (11), (13), and (14), the cost function is actually equivalent as follows:

$$p(K_1,K_2,...,K_D) \approx \frac{\left|\sum_{d=1}^{D}K_d\mathrm{E}\left[\sum_{i=1}^{M}\mathrm{e}^{\frac{2\pi\mathrm{fD}_iP_d}{N}}\left|\frac{\sin(\mathrm{fD}_i\pi)}{N\sin(\mathrm{fD}_i\pi/N)}\right|\right]\right|^2}{\sum_{z=1}^{N-1}\left|\sum_{d=1}^{D}K_d\mathrm{E}\left[\sum_{i=1}^{M}\mathrm{e}^{\frac{2\pi(z+\mathrm{fD}_i)P_d}{N}}\left|\frac{\sin\left((z+\mathrm{fD}_i)\pi\right)}{N\sin\left((z+\mathrm{fD}_i)\pi/N\right)}\right|\right]\right|^2}$$

$$= \frac{\left|\sum_{d=1}^{D}K_dG_{P_d}(0)\right|^2}{\sum_{z=1}^{N-1}\left|\sum_{d=1}^{D}K_dG_{P_d}(z)\right|^2}. \quad (17)$$

The channel is a classic Clark model and is described by Jakes, assuming that some parameters for the Doppler spectrum shape and fD_{max} are known, and p is a function of the weight vector $\{K_d\}$. $G_{P_d}(z)$ may be determined after some computation as follows:

$$G_{P_d}(z) = \mathrm{E}\left[\sum_{i=1}^{M} e^{\frac{2\pi(z+fD_i)P_d}{N}}\left|\frac{\sin\left((z+fD_i)\pi\right)}{N\sin\left((z+fD_i)\pi/N\right)}\right|\right]$$

$$= \lim_{\triangle\theta\to 0}\frac{1}{2\pi}\sum_{n=0}^{2\pi/\Delta\theta} e^{\frac{2\pi(z+fD_{max}\cos\theta)P_d}{N}}\left|\frac{\sin\left((z+fD_{max}\cos\theta)\pi\right)}{N\sin\left((z+fD_{max}\cos\theta)\pi/N\right)}\right|\Delta\theta$$

$$= \frac{1}{2\pi}\int_0^{2\pi} e^{\frac{2\pi(z+fD_{max}\cos\theta)P_d}{N}}\left|\frac{\sin\left((z+fD_{max}\cos\theta)\pi\right)}{N\sin\left((z+fD_{max}\cos\theta)\pi/N\right)}\right| d\theta,$$

$$z = 0,1,\ldots,N-1. \tag{18}$$

If $\{K_d\}$ is determined appropriately, without lost of generality, $P_S(K_1,K_2,\ldots,K_D) = 1$, and the P_{ICI} can be revised as follows:

$$P_{ICI}(K_1,K_2,\ldots,K_D) = \sum_{z=1}^{N-1}\left|\sum_{d=1}^{D-1} K_d G_{P_d}(z) + G_{P_D}(z)\left(\frac{1}{G_{P_D}(0)} - \sum_{d=1}^{D-1} K_d\frac{G_{P_d}(0)}{G_{P_D}(0)}\right)\right|^2$$

$$= \sum_{z=1}^{N-1}\left|\frac{G_{P_D}(z)}{G_{P_D}(0)} + \sum_{d=1}^{D-1} K_d\left(G_{P_d}(z) - G_{P_D}(z)\frac{G_{P_d}(0)}{G_{P_D}(0)}\right)\right|^2. \tag{19}$$

Now, the observed function may be set up as follows:

$$\boldsymbol{AK} = \boldsymbol{G}, \tag{20}$$

$$\boldsymbol{G} = \left[\frac{G_{P_D}(1)}{G_{P_D}(0)},\frac{G_{P_D}(2)}{G_{P_D}(0)},\ldots,\frac{G_{P_D}(N-1)}{G_{P_D}(0)}\right]^{\mathrm{T}},$$

$$\boldsymbol{K} = [K_1,K_2,\ldots,K_{D-1}]^{\mathrm{T}}, \tag{21}$$

$$[\boldsymbol{A}]_{z,d} = -G_{P_d}(z) + G_{P_D}(z)\frac{G_{P_d}(0)}{G_{P_D}(0)},$$

$$z = 1,2,\ldots,N-1;\ d = 1,2,\ldots,D-1.$$

Therefore, the $\boldsymbol{K}$ based on least square (LS) estimation is obtained as follows:

$$\boldsymbol{K}_{LS} = (\boldsymbol{A}^{\mathrm{H}}\boldsymbol{A})^{-1}\boldsymbol{A}^{\mathrm{H}}\boldsymbol{G}. \tag{22}$$

Now, $\boldsymbol{K} = [K_1,K_2,\ldots,K_{D-1}] = \boldsymbol{K}_{LS}$ is the solution of P_{ICI} function. The estimated value $\boldsymbol{K}'$ can be described by

$$\boldsymbol{K}' = \begin{bmatrix} \boldsymbol{K}_{LS} \\ \dfrac{1-\boldsymbol{G}_P\boldsymbol{K}_{LS}}{G_{P_D}(0)} \end{bmatrix}. \tag{23}$$

$\boldsymbol{G}_P$ is described as follows:

$$\boldsymbol{G}_P = [G_{P_1}(0),G_{P_2}(0),\ldots,G_{P_{D-1}}(0)]. \tag{24}$$

From above equations and processes, P_d and K_d are determined. With P_d and K_d, we can use Eqs. (7) and (8) to achieve diversity collection.

4 Simulation and analysis

In the TDSIC system, OFDM symbols are repeated by predefined delay value D. The offset frequency can be estimated with a simple method [9]. However, since residual frequency offset may exist and be negligible, simulation with system performance in frequency-offset channel is not the key point. All simulations are made in Doppler-spread channel to clarify the time variation of wireless channel.

Parameters in our simulations are described in detail as follows. OFDM system uses 128 sub-carriers, where the random input data are modulated with 16QAM symbols. Channel is Jakes' Doppler-spread channel with 12 multi-paths, and the available path parameter by other methods. As the impact of ICI from temporal variation is considered, noise is assumed to be negligible, and the ICI energy may be the main interference energy worth a consideration. In addition to all of these, assuming perfect time synchronization is achieved as well. For the purpose of comparing and verifying the algorithm with existing ICI cancellation algorithms and the proposed suboptimum method, 1/2 rate PCC-OFDM is also simulated and compared, in which just half of the data is transmitted as compared with in a normal OFDM system. In PCC-OFDM, the transmitted data abides by the operation rules and maps the data fed into a group of carriers with weight factor coming from polynomial function named coding cancellation. In the simulation, the weight factor is $\{-1,1\}$, so the spectrum efficiency only is one half of that in the normal OFDM system. Considering the fact that n normal OFDM, one sub-carrier is only loaded with one symbol, it is obvious that the spectrum efficiency of normal OFDM is better than PCC-OFDM.

The SIR and bit error ratio (BER) performance are presented in Figs. 2 and 3, respectively. In comparison to normal OFDM system, from Fig. 2, it can be seen that much better ICI performance gain is achieved with an approximate loss of frequency efficiency by 50%. With different delay D and identical channel conditions, at the lower fD_{max}, the lower D value has lower SIR comparison to the 1/2 PCC-OFDM, such as $D = 9$. However, at the fD_{max} beyond 0.25, the TDSIC system has better performance than 1/2 rate PCC-OFDM. Considering that system frequency efficiency will have a loss, while CP is added to combat ISI due to channel's delay spread, and the TDSIC system provides a little better frequency efficiency than PCC-OFDM when there are same number of sub-carriers. Comparing with the frequency-domain equalizer linear equalizer (FDE-LE) in time variation fading channel, E_b/N_0 is over 13 dB when BER is 10^{-2}, and the TDSIC also has a better superiority at the same mobile speed. When mobile speed is high, TDSIC system outperforms FDE-LE system.

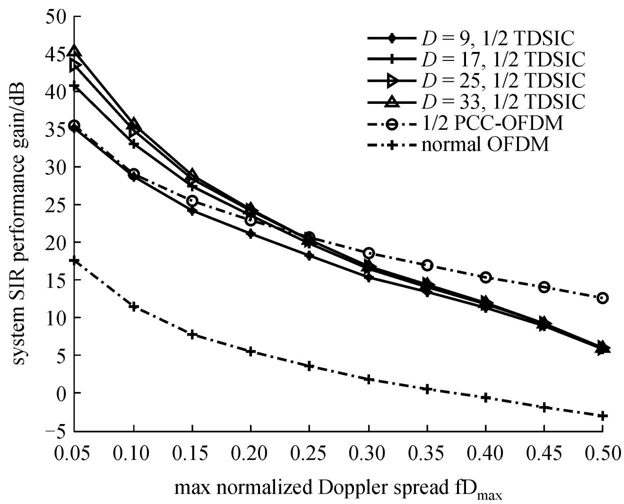

Fig. 2 SIR performances of TDSIC with different D as compared with PCC-OFDM in Jakes' Doppler-spread channel

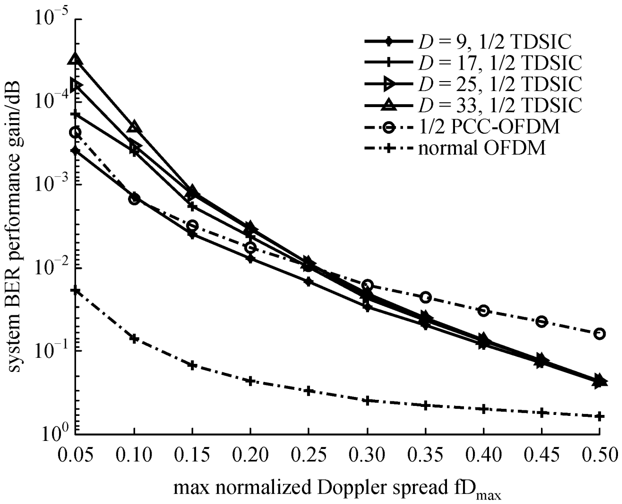

Fig. 3 BER performances of TDSIC with different D as compared with PCC-OFDM in Jakes' Doppler-spread channel

In Fig. 3, the conclusion may be obtained for cases when $D = 9$ is better than 1/2 PCC-OFDM, and the fD_{max} has limit 0.25. When fD_{max} is less than 0.25, PCC-OFDM has better BER than other delay TDSIC systems. An analogous conclusion is that, when fD_{max} is beyond 0.25, PCC-OFDM is worse than other TDSIC systems.

If the two systems are to be applied in the same frequency band with the same frequency efficiency, PCC-OFDM should have much more sub-carriers than that of TDSIC. In other words, in the TDSIC system, the sub-channel span is just a half of that in the PCC. When they are transmitted on the same time-variant channel, the normalized Doppler spread in TDSIC system is only a half of that in the PCC-OFDM. In conclusion, it can be seen in Figs. 2 and 3 that TDSIC performs much better than PCC-OFDM.

Although the performance of BER with time-domain windowing technology outperforms the normal OFDM and PCC-OFDM when with same E_b/N_0, it has a tradeoff of increasing the normalized frequency offset to more than 0.1. With this face in mind, we get our next conclusion from that in Fig. 4, i.e., the optimum performance can be obtained with time windowing at the same SNR.

5 Conclusion

In this paper, based on delay diversity technology, we proposed a TDSIC system that can be used to compensate ICI caused by the time-variation wireless channel or in high-speed mobile environment. Because all symbols are repeatedly used, the TDSIC system is insensitive to frequency offset channel. In the algorithm of TDSIC, P_d and K_d are not at all related with the transmitted data

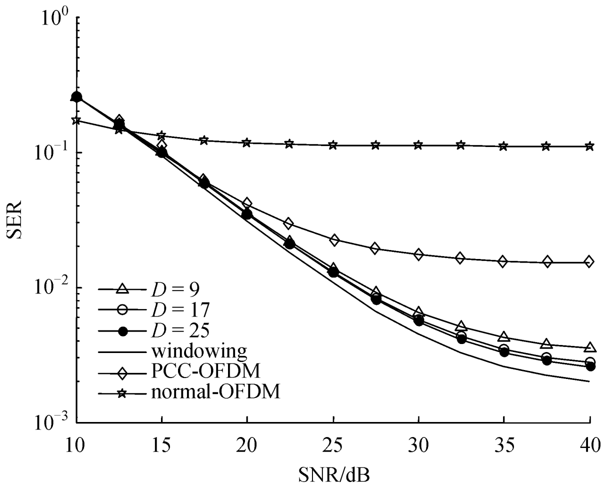

Fig. 4 SER performances of TDSIC with different D as compared with windowing and PCC-OFDM in Jakes' channel

according to presented equations; instead, they are mainly determined by the parameter of the time-variant channel. There are different performances for different fading channels. All mentioned parameters do not need to be pursued on sites and can be achieved offline; this way they become handy when needed, with the only exception of the max Doppler spread of the channel, which requires an estimation that can be obtained, as proposed in Ref. [10]. All of these not only make it become very effective in real practice but also reduce the system complexity greatly.

There is still room for improvement in the future, such as finding a method to avoid the high PAPR similar SC-FDE; another concern is with the existence of K_d, system complexity could increase. Finally, in order to estimate P_d and K_d offline and that of the real channel, not only the deviation between the estimated Doppler spread and the real one but also the difference that existed in the PDF between that of the channel need to be included in the consideration, and this could lead to a loss in system performance.

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