Impact of Imperfect Channel Estimation on OFDM/TDM Performance

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***Summary*:** Orthogonal frequency division multiplexing (OFDM) combined with time division multiplexing (TDM), called OFDM/TDM, can bridge the conventional OFDM and single-carrier (SC) transmission by using the frequency-domain equalization (FDE). Accurate channel estimation is required for FDE. A well-known channel estimation scheme is pilot-based channel estimation that uses periodically transmitted pilot signals. If channel estimation error is present, the inter-symbol interference (ISI) is produced in OFDM/TDM. In this paper, we apply a Gaussian approximation to the channel estimation error and theoretically investigate the impact of the imperfect channel estimation on the average bit error rate (BER) performance of OFDM/TDM. It is shown by the numerical evaluation that the channel estimation error degrades more the BER performance when OFDM/TDM approaches SC.

1. Introduction

Mobile radio channel is composed of many paths with different time delays due to the reflection, refraction or diffraction of the transmitted signal by obstacles placed between a transmitter and a receiver. In such multipath channel, the received signal is a superposition of several delayed and scaled copies of the transmitted signal, giving rise to frequency-selective fading. Frequency-selective fading produces inter-symbol interference (ISI) and hence significantly degrades the transmission performance of the single carrier (SC) transmission systems [1]. Recently, orthogonal frequency division multiplexing (OFDM) has been attracting much attention because of its robustness against frequency-selective fading and simple one-tap frequency domain equalization (FDE) and has already been adopted in some wireless communications systems [2]-[4]. However, OFDM signal has high peak-to-average-power ratio (PAPR) [5]-[7] and the bit error rate (BER) performance severely degrades when multipaths with time delays longer than the guard interval (GI) are present [8].

Recently, we have shown [9], [10] that OFDM combined with TDM (OFDM/TDM) can overcome the PAPR problem or improve the delay robustness by changing the design parameters, without sacrificing the transmission data rate and the bandwidth at all. The same OFDM/TDM was presented in [11], but the objective of [11] is to increase the transmission data rate for the given bandwidth, but we keep the data rate the same as the conventional OFDM. OFDM/TDM can bridge the conventional OFDM and SC transmission with the FDE. It was shown by computer simulation [9], [10] that OFDM/TDM can improve the BER performance in comparison to the conventional OFDM by applying the FDE based on minimum mean square error (MMSE) criterion; however, the perfect channel estimation was considered. In this paper, we apply a Gaussian approximation to the channel estimation error and theoretically evaluate how the channel estimation error affects the BER performance of OFDM/TDM.

Remainder of this paper is organized as follows. In Sect. 2, the overall system model of OFDM/TDM is described. Sects. 3 and 4 present the channel estimation error model and the principle of the zero forcing (ZF)-FDE and MMSE-FDE, respectively. In Sect. 5, the instantaneous signal-to-interference plus noise power ratio (SINR) and the conditional BER are derived. Sect. 6 evaluates, by the Monte-Carlo numerical method, the BER performance of OFDM/TDM in a frequency-selective Rayleigh fading channel and then, compares it with those of the conventional OFDM and SC transmissions. Sect. 7 provides some conclusions and future work.

2. OFDM/TDM System Model

Figure 1 shows the transmitter/receiver structure of OFDM/TDM and Figure 2 illustrates the OFDM/TDM frame structure. In OFDM/TDM designed for PAPR reduction, a sequence of K OFDM signals with $N_m=N_c/K$ subcarriers is transmitted during the N_c-sample inverse fast Fourier transform (IFFT) block. K=1 and N_c give the conventional OFDM with N_c subcarriers and SC transmission, respectively. If the number of subcarriers, $N_m=N_c/K$, and the IFFT block length, N_c, are simply replaced by N_c and KN_c, repectively, then the OFDM/TDM signal designed for improving the delay robustness can be expressed. Therefore, for the same of brevity, we describe only the case for reducing the PAPR.

One frame consists of K slots; in each slot, N_m data symbols are transmitted using N_m subcarriers. The transmitted data symbol sequence in the k-th slot is written in a vector form as

$$\mathbf{d}_k=[d_{k,0},d_{k,1},\cdots d_{k,N_m-1}] \quad \text{for } k=0\sim K\text{-}1, \ (1)$$

where k is the slot index. There are K OFDM signals in the transmission frame of OFDM/TDM. The transmitted data symbol sequence vector $(N_mK\times1)=(N_c\times1)$ in a frame is represented as

$$\mathbf{d}=[\mathbf{d}_0,\mathbf{d}_1,\cdots\mathbf{d}_{K-1}]^T, \ (2)$$

where T denotes the transpose. The transmitted OFDM/TDM signal vector $(N_c\times1)$ per transmission frame in the time-domain can be expressed as

$$\mathbf{s}=\sqrt{2E_s/T_c}\,\mathbf{F}^{-1}\mathbf{d} \quad (3)$$

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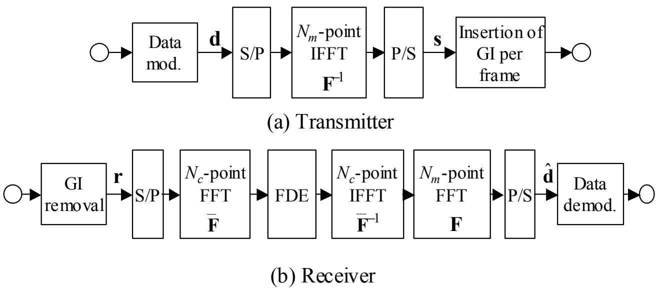


Fig. 1 Transmitter/receiver structures of OFDM/TDM.

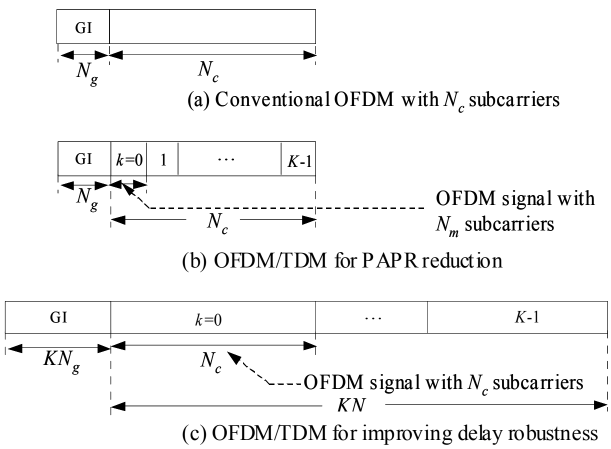


Fig. 2 OFDM/TDM frame structure.

with

$$\mathbf{F} = \mathbf{I}_{K\times K} \otimes \mathbf{f} = \begin{bmatrix} \mathbf{f} & & & \\ & \mathbf{f} & & \\ & & \ddots & \\ & & & \mathbf{f} \end{bmatrix}, \quad (4)$$

where E_s and T_c denote the data symbol energy and the fast Fourier transform (FFT) sampling interval, respectively. $\otimes$ is the Kronecker product. $\mathbf{F}$ and $\mathbf{I}$ are the combined block Fourier transform matrix ($N_mK\times N_mK=N_c\times N_c$) and the identity matrix ($K\times K$), respectively. $\mathbf{f}$ is the Fourier transform matrix ($N_m\times N_m$). Before transmission, the OFDM/TDM frame is cyclically extended, that is, the last N_g samples in the OFDM/TDM frame are inserted as the guard interval (GI) placed at the beginning of the frame.

The cyclic-extended OFDM/TDM signal is transmitted over a frequency-selective fading channel. The fading channel is assumed to be composed of sample-spaced L discrete propagation paths and the maximum time delay is less than the GI length of N_g. We assume a block fading, where the paths gains remain constant during one OFDM/TDM frame. The channel impulse response vector $\mathbf{h}$ ($N_c\times 1$) is defined as

$$\mathbf{h} = [h_0, \cdots, h_{L-1}, h_L(=0), \cdots h_{N_c-1}(=0)]^T. \quad (5)$$

The received OFDM/TDM signal after removal of GI can be expressed using $N_c\times 1$ vector represented as

$$\begin{aligned}\mathbf{r} &= \mathbf{Hs} + \mathbf{n} \\ &= \sqrt{2E_s/T_c}\,\mathbf{HF}^{-1}\mathbf{d} + \mathbf{n}\end{aligned}, \quad (6)$$

where $\mathbf{n}$ is the $N_c\times 1$ noise vector due to the additive white Gaussian noise (AWGN). The $N_c\times N_c$ channel matrix $\mathbf{H}$ is given by

$$\mathbf{H} = [\mathbf{h}^0, \mathbf{h}^1, \cdots, \mathbf{h}^q, \cdots, \mathbf{h}^{N_c-1}], \quad (7)$$

where $\mathbf{h}^q$ is the q-cyclic time shifted channel impulse response vector ($N_c\times 1$):

$$\begin{cases}\mathbf{h}^0 = \mathbf{h} \\ \mathbf{h}^q = [h_{N_c-q}, \cdots, h_{N_c-1}, h_0, \cdots h_{N_c-q-1}]^T\end{cases}. \quad (8)$$

Consider the Fourier transform of $\mathbf{h}^q$:

$$\mathbf{D}^q = \overline{\mathbf{F}}\mathbf{h}^q, \quad (9)$$

where $\overline{\mathbf{F}}$ is the N_c-point discrete Fourier transform matrix ($N_c\times N_c$). A cyclic shift in the time domain results in a phase rotation in the frequency domain. Therefore, $\mathbf{D}^q$ is written as

$$\mathbf{D}^q = \mathbf{Z}\mathbf{f}^q \quad (10)$$

where

$$\begin{cases}\mathbf{Z} = diag\{\overline{\mathbf{F}}\mathbf{h}^0\} \\ \mathbf{f}^q = \dfrac{1}{N_c}[1, e^{-j2\pi q/N_c}, \cdots, e^{-j2\pi q(N_c-1)/N_c}]^T\end{cases} \quad (11)$$

and $diag\{a_0, a_1, \cdots, a_{Nc-1}\}$ is the diagonal matrix whose diagonal elements are $a_0, a_1, \cdots, a_{Nc-1}$. Therefore, the following equation is obtained from Eqs. (5), (9) and (10) as

$$\overline{\mathbf{F}}\mathbf{H} = \mathbf{Z}\overline{\mathbf{F}}. \quad (12)$$

Therefore, we have

$$\mathbf{H} = \overline{\mathbf{F}}^{-1}\mathbf{Z}\overline{\mathbf{F}}. \quad (13)$$

3. Channel Estimation Error Model

We adopt the following error model in channel estimation:

$$\hat{\mathbf{h}} = \mathbf{h} + \mathbf{e}, \quad (14)$$

where $\hat{\mathbf{h}} = [\hat{h}_0, \hat{h}_1, \cdots, \hat{h}_{L-1}, 0, \cdots, 0]^T$ is the $N_c\times 1$ vector representing the estimated channel impulse response and $\mathbf{e} = [e_0, e_1, \cdots, e_{L-1}, 0, \cdots, 0]^T$ represents the $N_c\times 1$ error vector. It is assumed that $\mathbf{e}$ is independent of $\mathbf{h}$ and that $\{e_l; l=0\sim L\text{-}1\}$ are modeled as independent zero-mean complex-valued Gaussian variables with the variance of $2\sigma_{error}^2$. The estimate of the channel frequency transfer function can be expressed as the following $N_c\times N_c$ matrix:

$$\hat{\mathbf{Z}} = diag\{\overline{\mathbf{F}}\hat{\mathbf{h}}\} = \mathbf{Z} + diag\{\overline{\mathbf{F}}\mathbf{e}\}. \quad (15)$$

4. FDE

The transmitted OFDM/TDM signal is distorted by the frequency-selective fading channel and thus, the equalizer is required to compensate for the frequency-selectivity. We consider ZF-FDE and MMSE-FDE.

4.1 ZF-FDE

The decision variable after ZF-FDE can be given by

$$\hat{\mathbf{d}}_{ZF} = \left[(\hat{\mathbf{H}}\mathbf{F}^{-1})^H(\hat{\mathbf{H}}\mathbf{F}^{-1})\right]^{-1}(\hat{\mathbf{H}}\mathbf{F}^{-1})^H\mathbf{r} \quad (16)$$

where

$$\hat{\mathbf{H}} = \overline{\mathbf{F}}^{-1}\hat{\mathbf{Z}}\overline{\mathbf{F}} \quad (17)$$

and $(.)^H$ is the conjugate transpose. Using Eqs. (15) and (17), Eq. (16) can be approximated as

$$\begin{aligned}\hat{\mathbf{d}}_{ZF} &\approx \sqrt{2E_s/T_c}\,\mathbf{d} + \mathbf{\Psi}\mathbf{Z}^{-1}\overline{\mathbf{F}}\mathbf{n} \\ &\quad - \mathbf{\Psi}\mathbf{Z}^{-1}diag\{\overline{\mathbf{F}}\mathbf{e}\}(\sqrt{2E_s/T_c}\,\mathbf{\Psi}^{-1}\mathbf{d} + \mathbf{Z}^{-1}\overline{\mathbf{F}}\mathbf{n})\end{aligned}, \quad (18)$$

where $\mathbf{\Psi} = \mathbf{F}\overline{\mathbf{F}}^{-1}$. In Eq. (18), the first, second and third terms denote the desired signal, noise and interference due to channel estimation error, respectively.

4.2 MMSE-FDE

The decision variable after MMSE-FDE can be given by

$$\hat{\mathbf{d}}_{MMSE} = \left[(\hat{\mathbf{H}}\mathbf{F}^{-1})^H(\hat{\mathbf{H}}\mathbf{F}^{-1}) + SNR^{-1}\mathbf{I}_{N_c\times N_c}\right]^{-1}(\hat{\mathbf{H}}\mathbf{F}^{-1})^H\mathbf{r}, \quad (19)$$

where SNR represents the signal-to-noise power ratio. Using Eqs. (15) and (17), Eq. (19) can be approximated as

$$\begin{aligned}\hat{\mathbf{d}}_{MMSE} &\approx \sqrt{2E_s/T_c}\,diag\{\mathbf{B}_{MMSE}\}\mathbf{d} \\ &\quad + \sqrt{2E_s/T_c}\left(\mathbf{B}_{MMSE} - diag\{\mathbf{B}_{MMSE}\}\right)\mathbf{d} \\ &\quad + \hat{\mathbf{B}}_{MMSE}\mathbf{\Psi}\mathbf{Z}^{-1}\overline{\mathbf{F}}\mathbf{n} \\ &\quad + SNR^{-1}\mathbf{B}_{MMSE}\mathbf{\Psi}(\mathbf{Z}^H\mathbf{Z})^{-1}\mathbf{C}(\mathbf{Z}^H\mathbf{Z})^{-1}\mathbf{\Psi}^{-1}\mathbf{B}_{MMSE} \\ &\qquad \times\sqrt{2E_s/T_c}\,\mathbf{d} \\ &\quad - \hat{\mathbf{B}}_{MMSE}\mathbf{\Psi}\mathbf{Z}^{-1}diag\{\overline{\mathbf{F}}\mathbf{e}\}(\sqrt{2E_s/T_c}\,\mathbf{\Psi}^{-1}\mathbf{d} + \mathbf{Z}^{-1}\overline{\mathbf{F}}\mathbf{n})\end{aligned}, \quad (20)$$

with

$$\begin{cases}\mathbf{B}_{MMSE} = [\mathbf{I}_{N_c\times N_c} + SNR^{-1}\mathbf{\Psi}(\mathbf{Z}^H\mathbf{Z})^{-1}\mathbf{\Psi}^{-1}]^{-1} \\ \hat{\mathbf{B}}_{MMSE} = [\mathbf{I}_{N_c\times N_c} \\ \qquad + SNR^{-1}\mathbf{B}_{MMSE}\mathbf{\Psi}(\mathbf{Z}^H\mathbf{Z})^{-1}\mathbf{C}(\mathbf{Z}^H\mathbf{Z})^{-1}\mathbf{\Psi}^{-1}]\mathbf{B}_{MMSE} \\ \mathbf{C} = \mathbf{Z}^H diag\{\overline{\mathbf{F}}\mathbf{e}\} + (diag\{\overline{\mathbf{F}}\mathbf{e}\})^H(\mathbf{Z} + diag\{\overline{\mathbf{F}}\mathbf{e}\})\end{cases} \quad (21)$$

In Eq. (20), the first, second and third terms represent the desired signal, the interpath interference (IPI) and the noise components, respectively. The forth and fifth terms represent the interference components due to channel estimation error.

5. BER Analysis

The interference due to channel estimation error can be approximated as a zero-mean complex-valued Gaussian process. First, we obtain the SINR of decision variable for the given **Z** and then, the conditional BER for the given **Z**. The data symbols are assumed to be uncorrelated and have a variance of unity ($|d|^2=1$). In this section, we use the following notation.

- $[\mathbf{x}]_i$ denotes the i-th element in the column vector **x**.
- $[\mathbf{A}]_{ij}$ denotes the element in the i-th row and the j-th column of the matrix **A**.

5.1 Instantaneous SINR

Using $E[e_l^* e_{l'}] = 0\ (2\sigma_{error}^2)\ for\ l \neq l'\ (l = l')$ and $E[\mathbf{e}^H\mathbf{h}] = \mathbf{0}$, the instantaneous SINR γ_i of the i-th data symbol for ZF-FDE can be expressed as

$$\begin{aligned}\gamma_i &= \frac{E[\left|[\sqrt{2E_s/T_c}\,\mathbf{d}]_i\right|^2]}{\begin{array}{l}E[\left|[\mathbf{Z}^{-1}\overline{\mathbf{F}}\mathbf{n}]_i\right|^2] \\ \quad + E[\left|[\mathbf{\Psi}\mathbf{Z}^{-1}diag\{\overline{\mathbf{F}}\mathbf{e}\}(\sqrt{2E_s/T_c}\,\mathbf{\Psi}^{-1}\mathbf{d} + \mathbf{Z}^{-1}\overline{\mathbf{F}}\mathbf{n})]_i\right|^2]\end{array}} \\ &\approx (2E_s/N_0)\big/[\mathbf{\Psi}\mathbf{\Lambda}\mathbf{\Psi}^{-1}]_{ii} \quad (22)\end{aligned}$$

with

$$\begin{aligned}\mathbf{\Lambda} = diag\Bigg\{&\frac{1}{N_c|H_0|^2} + \frac{2L\sigma_{error}^2}{N_c|H_0|^2}\left(\frac{E_s}{N_0} + \frac{1}{N_c|H_0|^2}\right), \cdots, \\ &\frac{1}{N_c|H_{N_c-1}|^2} + \frac{2L\sigma_{error}^2}{N_c|H_{N_c-1}|^2}\left(\frac{E_s}{N_0} + \frac{1}{N_c|H_{N_c-1}|^2}\right)\Bigg\}\end{aligned}$$

$$, \text{for ZF-FDE} \quad (23)$$

where N_0 is the AWGN power spectrum density. The instantaneous SINR γ_i of the i-th data symbol for MMSE-FDE can be expressed as

$$\gamma_i \approx \frac{2|[\mathbf{B}_{MMSE}]_{ii}|^2}{\begin{array}{l}\sum_{j=0,j\neq i}^{N_c-1}|[\mathbf{B}_{MMSE}]_{ij}|^2 \\ +\left(\frac{E_s}{N_0}\right)^{-2}\sum_{i=0}^{N_c-1}|[\mathbf{B}_{MMSE}]_{ii}|^2\frac{4L\sigma_{error}^2}{N_c} \\ \times[\mathbf{\Omega}\{(\mathbf{Z}^H\mathbf{Z})^{-3}+\frac{L\sigma_{error}^2}{N_c}(\mathbf{Z}^H\mathbf{Z})^{-4}\}\mathbf{\Omega}^H]_{ii} \\ +\left\{1+\frac{2L\sigma_{error}^2}{N_c}+\left(\frac{2L\sigma_{error}^2}{N_c}\right)^2\right\} \\ \times\left[\begin{array}{l}\left\{\left(\frac{E_s}{N_0}\right)^{-1}+\frac{2L\sigma_{error}^2}{N_c}\right\}[\mathbf{\Omega}(\mathbf{Z}^H\mathbf{Z})^{-1}\mathbf{\Omega}^H]_{ii} \\ +\left(\frac{E_s}{N_0}\right)^{-1}\frac{2L\sigma_{error}^2}{N_c}[\mathbf{\Omega}(\mathbf{Z}^H\mathbf{Z})^{-2}\mathbf{\Omega}^H]_{ii}\end{array}\right]\end{array}} \quad (24)$$

for MMSE-FDE

where $\mathbf{\Omega} = \mathbf{B}_{MMSE}\mathbf{\Psi}$.

5.2 Conditional BER

The conditional BER of i-th data symbol for the given $\mathbf{Z}$ can be given by

$$P_b(E_s/N_0, \mathbf{Z}) = 1/2erfc(\sqrt{\gamma_i/4}) \quad (25)$$

for QPSK, where $erfc(x) = (2/\sqrt{\pi})\int_x^\infty \exp(-t^2)dt$ is the complementary error function. The theoretical average BER of the i-th data symbol can be numerically evaluated by averaging Eq. (25) over $\mathbf{Z}$:

$$P_b^{(i)}(E_s/N_0) = \int\cdots\int \frac{1}{2}erfc(\sqrt{\gamma_i/4})p(\mathbf{Z})d\mathbf{Z}, \quad (26)$$

where $p(\mathbf{Z})$ is the joint probability density function (pdf) of $\mathbf{Z}$. Then, the average BER over one frame is obtained by

$$P_b(E_s/N_0) = \frac{1}{N_c}\sum_{i=0}^{N_c-1} P_b^{(i)}(E_s/N_0). \quad (27)$$

6. Numerical Results

The numerical evaluation of the theoretical average BER is done by Monte-Carlo numerical computation method as follows. The set of $\mathbf{Z}$ is generated and then the conditional BER of the i-th data symbol for the given symbol energy E_s/N_0 and $\mathbf{Z}$ is computed using Eq. (25). This is repeated sufficient number of times to obtain the theoretical average BER of Eq. (26). The numerical conditions are summarized in Table 1. We assume $N_m=N_c/K$ subcarriers (K=1~256), frame length of N_c=256 samples. QPSK data-modulation is assumed. The propagation channel is an L=16-path frequency-selective block Rayleigh fading channel having uniform power delay profile.

Table 1 Numerical conditions

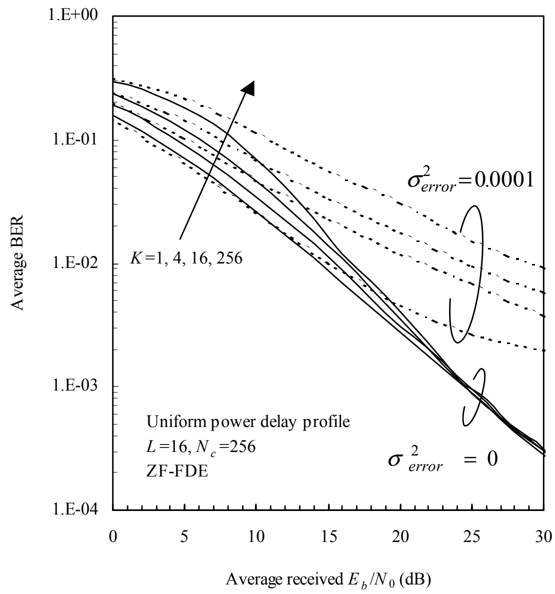
Transmitter	Data modulation	QPSK
	No. of FFT points	N_m=256/K
	Number of slots per frame	K=1~256
	Frame length	N_c=256
	Guard interval (GI)	N_g=32
Channel	Fading	Frequency-selective Rayleigh faidng
	Number of paths	L=16
Receiver	Number of FFT points	N_c=256 N_m=256/K
	Frequency-domain equalization	ZF, MMSE

Fig. 3 plots the average BER performance of OFDM/TDM using ZF- and MMSE-FDE as a function of the average E_b/N_0, with K as a parameter, when $\sigma_{error}^2 = 0.0001$. For comparison, the BER performance of $\sigma_{error}^2 = 0$, which corresponds to the ideal channel estimation, is also plotted. OFDM/TDM represents OFDM and SC system when K=1 and K=256, respectively. It is found from Fig. 3 that with increase in K (as OFDM/TDM approaches the SC transmission), the impact of channel estimation error becomes severer. A possible reason for this is discussed below. When K=256 (SC), the N_c-point IFFT followed by N_m(=1)-point FFT is carried out after FDE and thus this is equivalent to N_c-point IFFT. On the other hand, when K=1 (OFDM), the above operation is equivalent to no IFFT/FFT processing due to $N_m=N_c$. Therefore, in the case of SC transmission, the channel estimation error at each frequency component $diag\{\overline{\mathbf{F}}\mathbf{e}\}$ is accumulated (or averaged) by N_c-point IFFT. This suggests that the SC transmission with FDE is very sensitive to the channel estimation error. On the other hand, the conventional OFDM dose not require the IFFT and FFT after FDE and thus, the BER degradation due to the channel estimation error is minimal.

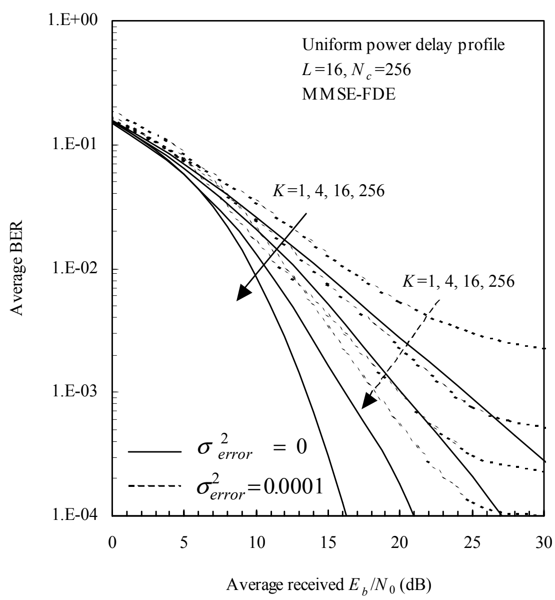
Fig. 4 plots the average BER performance as a function of the channel estimation error σ_{error}^2. It is confirmed from Fig. 4 that the BER degradation from ideal channel estimation ($\sigma_{error}^2 = 0$) becomes larger as K increases, even if the channel estimation error is the same.

7. Conclusion

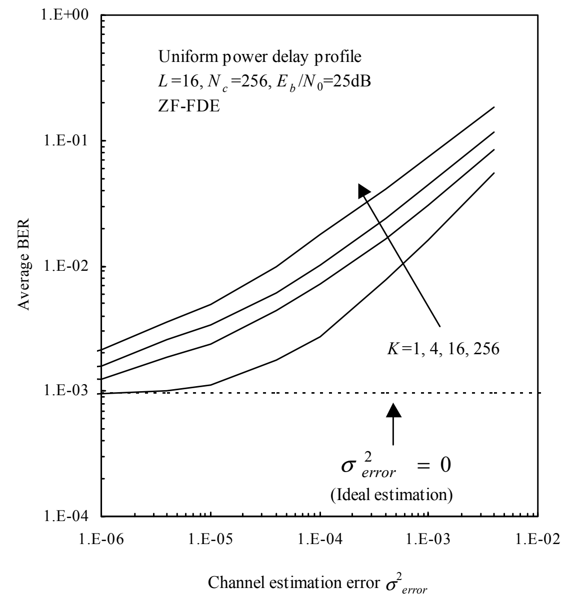
This paper investigated the impact of channel estimation error on the OFDM/TDM performance. The channel estimation error was modeled as the Gaussian approximation. It was found that the channel estimation error significantly affects the BER performance as OFDM/TDM approaches the SC transmission. In this paper, only the numerical results are shown and thus, the comparison to the simulation results is left for the future work.



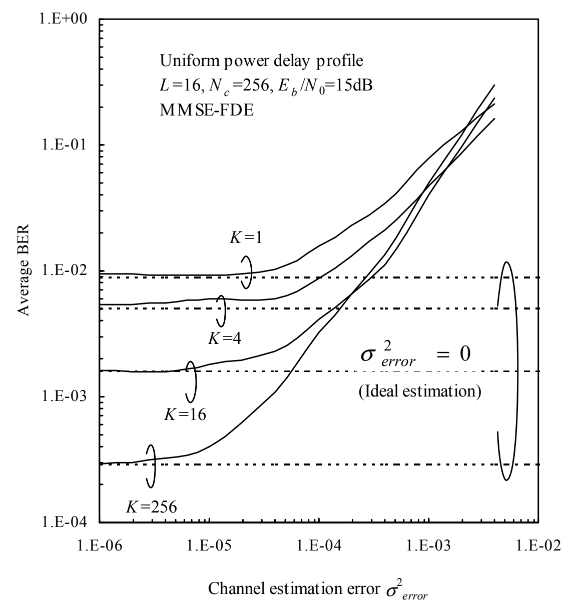
(a) ZF-FDE



(b) MMSE-FDE
Fig.3 BER performance.



(a) ZF-FDE



(b) MMSE-FDE
Fig.4 Impact of channel estimation error.

References

[1] J. G. Proakis, *Digital communications*, McGraw-Hill, 3rd Ed., McGraw Hill.

[2] R. D.J. van Nee, R. Prasad and R. van Nee, *OFDM for wireless multimedia communications,* Artech House, Jan. 2000.

[3] S. Hara and R. Prasad, *Multicarrier Techniques for 4G Mobile Communications*, Artech House, June 2003.

[4] L.Hanzo, W. Webb and T. Keller, *Single-and Multi-carrier Quadrature Amplitude Modulation*, John Willey & Sons, 2000.

[5] V. Tarokh and H. Jafarkhani, "On the computation and reduction of the peak-to-average power ration in multicarrier communications," IEEE Trans. Commun., Vol.48, No.1, pp.37-44, Jan. 2000.

[6] D. Wulich and L. Goldfield, "Reduction of peak factor in orthogonal multicarrier modulation by amplitude limiting and coding," IEEE Trans. Commun., Vol.47, No.1, pp. 18-21, Jan. 1999.

[7] X. Li and L. J. Cimini, "Effects of clipping and filtering on the performance of OFDM," IEEE Commun. Letters, Vol.2, No.5, pp. 150-159, May 1998.

[8] S. Suyama, M. Ito, H. Suzuki and K. Fukawa, "A scattered pilot OFDM receiver with equalization for multipath environments with delay difference greater than guard interval," IEICE Trans. Commun. Vol.E86-B, No.1, pp.283-290, Jan. 2003.

[9] H. Gacanin, S. Takaoka and F. Adachi., "Generalized OFDM for improving the robstness against the frequency-selective fading channel," Technicial Report of IEICE, RCS2004-63, Vol.104, No.63, pp.43-48, May 2004.

[10] H. Gacanin, S. Takaoka and F. Adachi, "Generalized OFDM for bridging between OFDM and single-carrier transmission," ICCS 2004, Singapore, pp. 145-149, Set. 2004.

[11] C. V. Sinn, J.Götze and M. Haardt, "Common architectures for TD-CDMA and OFDM based mobile radio systems without the necessity of a cyclic prefix," MS-SS Workshop, DLR, Oberpfaffenhofen, Germany, Sept. 2001.