Duobinary Partial Response Filtered SC-FDE

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Abstract— **Frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion significantly improves the bit error rate (BER) performance of single-carrier (SC) transmission by suppressing the channel frequency-selectivity. Partial response (PR) filtering, e.g., duobinary PR filtering, achieves higher spectral efficiency by permitting certain amount of inter-symbol interference (ISI). However, employing the MMSE-FDE solely is not sufficient to mitigate the effect of such amount of ISI produced by PR filtering, and hence BER performance degrades. To reduce the effect of ISI, we introduce frequency-domain iterative ISI cancellation and time-domain maximum likelihood sequence estimation (MLSE). Performance of duobinary PR filtered SC-FDE is evaluated in terms of BER by computer simulation.**

Keywords-component; Frequency-domain equalization, partial response filter, frequency-domain ISI cancellation, MLSE

I. INTRODUCTION

High-speed and high-quality data services are strongly demanded in the next-generation mobile communication systems. However, the broadband wireless channel is composed of many propagation paths having different time delays. This makes the channel become frequency-selective, in which inter-symbol interference (ISI) degrades the system performance of single-carrier (SC) transmission in terms of bit error rate (BER) [1]. To remedy this problem, frequency-domain equalization (FDE) based on minimum mean square error (MMSE) criterion technique is serviceable for suppressing the frequency selectivity [2, 3].

Square-root Nyquist filter is typically used as transmit/receive filters in the SC transmission system [4]. However, as the filter roll-off factor increases, the required transmission bandwidth increases. On the other hand, partial response (PR) filter achieves higher spectral efficiency because of narrower transmission bandwidth, whereas the BER performance becomes worse as a result from permitting certain amount of ISI produced by the filter itself [1, 5].

Time-domain approaches for mitigating the problem of ISI such as transmit precoding and maximum likelihood sequence estimation (MLSE) has been discussed in the literature [5, 6], but frequency-nonselective channel was assumed. We alternatively consider the SC-FDE using frequency-domain duobinary PR transmit filter. MMSE-FDE is employed at the receiver. However, employing only MMSE-FDE is not sufficient to mitigate the effect of ISI [7].

In this paper, we introduce two approaches for SC-FDE using duobinary PR transmit filter. One acquires the advantage of using frequency-domain processing by employing the frequency-domain iterative ISI cancellation [8] at the receiver. Another approach considers the MLSE [1], which is a time-domain signal detection approach.

The remainder of this paper is organized as follows. In Sect. II, duobinary PR filtered SC-FDE with frequency-domain iterative ISI cancellation is presented. In Sect. III, duobinary PR filtered SC-FDE with MLSE is presented. In Sect. IV, computer simulation results are shown in aspect of BER performance. Finally, Sect. V summarizes the paper.

II. SC-FDE WITH FREQUENCY-DOMAIN ITERATIVE ISI CANCELLATION

The SC-FDE system model using duobinary PR filter and frequency-domain iterative ISI cancellation are illustrated in Fig. 1. Throughout the paper, the T_s symbol-spaced discrete-time signal representation is considered.

At the transmitter, a block of data-modulated symbols $\{d(n);n=0\sim N_c-1\}$ is transformed into N_c frequency components $\{D(k);k=0\sim N_c-1\}$ by applying N_c-point fast Fourier transform (FFT). Then, the duobinary PR filter is applied by multiplying the frequency-domain signal by filter coefficients $\{W_T(k);k=0\sim N_c-1\}$. N_c-point inverse FFT (IFFT) is applied to the transmit filter output $\{S(k);k=0\sim N_c-1\}$ to obtain the time-domain transmit signal $\{s(n);n=0\sim N_c-1\}$. To avoid the presence of inter-block interference (IBI), the last N_g samples of each transmission block are copied as a cyclic prefix (CP) and inserted into the guard interval (GI) placed at the beginning of each bock.

At the receiver, after the removal of the GI, the received signal $\{r(n);n=0\sim N_c-1\}$ is transformed into N_c frequency components $\{R(k);k=0\sim N_c-1\}$ by N_c-point FFT. MMSE-FDE and ISI cancellation are carried out in the frequency-domain. Finally, IFFT is applied to obtain the soft decision symbol for data modulation.

A. Transmit and Received Signals

The time-domain transmit filter output $\{s(n);n=0\sim N_c-1\}$ can be expressed as

$$\begin{aligned} s(n) &= \frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1} S(k)\exp\left(j2\pi n\frac{k}{N_c}\right) \\ &= \frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1} D(k)W_T(k)\exp\left(j2\pi n\frac{k}{N_c}\right), \end{aligned} \tag{1}$$

where $W_T(k)$ can be expressed as

$$W_T(k)=\sqrt{2}\cos\left(\frac{k\pi}{N_c}\right)\exp\left(-j\frac{k\pi}{N_c}\right). \quad (2)$$

We assume a symbol-spaced frequency-selective block Rayleigh fading channel composed of L distinct propagation paths. The channel impulse response $h(\tau)$ is given by

$$h(\tau)=\sum_{l=0}^{L-1}h_l\cdot\delta(\tau-\tau_l), \quad (3)$$

where τ_l is the delay time and h_l is the complex-valued path gain with $E[\sum_{l=0}^{L-1}|h_l|^2]=1$.

The CP-removed received signal $r(n)$ can be expressed as

$$r(n)=\sqrt{\frac{2E_s}{T_s}}\sum_{l=0}^{L-1}h_l\cdot s(n-l)+\eta(n), \quad (4)$$

where E_s denotes the symbol energy. $\eta(n)$ is the zero-mean additive white Gaussian noise (AWGN) with variance $2N_0/T_s$ with N_0 being the one-sided power spectrum density of AWGN.

B. MMSE-FDE and ISI Cancellation

N_c-point FFT is applied to decompose $\{r(n)\}$ into frequency-domain received signal $R(k)$, which can be expressed as

$$R(k)=\sqrt{\frac{2E_s}{T_s}}H(k)S(k)+\Pi(k), \quad (5)$$

where

$$\begin{cases}H(k)=\sum_{l=0}^{L-1}h_l\exp\left(-j2\pi k\frac{l}{N_c}\right)\\ \Pi(k)=\frac{1}{\sqrt{N_c}}\sum_{n=0}^{N_c-1}\eta(n)\exp\left(-j2\pi k\frac{n}{N_c}\right)\end{cases}. \quad (6)$$

Here, $H(k)$ and $\Pi(k)$ are the channel gain and the noise component due to AWGN, respectively.

MMSE-FDE and frequency-domain ISI cancellation are done iteratively, where the ith $(i=0,1,\ldots,I)$ iteration stage of MMSE-FDE and ISI cancellation is given by

$$\begin{aligned}\hat{R}^{(i)}(k)&=R(k)W_R^{(i)}(k)\\&=\sqrt{\frac{2E_s}{T_s}}D(k)\hat{H}^{(i)}(k)+\hat{\Pi}^{(i)}(k)\end{aligned}, \quad (7)$$

where

$$\begin{cases}\hat{H}^{(i)}(k)=W_T(k)H(k)W_R^{(i)}(k)\\ \hat{\Pi}^{(i)}(k)=\Pi(k)W_R^{(i)}(k)\end{cases}. \quad (8)$$

$W_R^{(i)}(k)$ is the equalization weight at the ith iteration stage. $\hat{H}^{(i)}(k)$ and $\hat{\Pi}^{(i)}(k)$ are the equivalent channel gain and the noise component after performing MMSE-FDE at the ith iteration stage, respectively. The derivation of FDE weight for each iteration stage is later described in Sect. II-*D*.

ISI cancellation is performed on $\hat{R}^{(i)}(k)$ in the frequency-domain as

$$\tilde{R}^{(i)}(k)=\hat{R}^{(i)}(k)-\tilde{M}^{(i)}(k), \quad (9)$$

where $\tilde{M}^{(i)}(k)$ is the residual ISI replica which is given as

$$\tilde{M}^{(i)}(k)=\begin{cases}0 & \text{for } i=0\\ \sqrt{\frac{2E_s}{T_s}}\left\{\hat{H}^{(i)}(k)-A^{(i)}\right\}\tilde{D}^{(i-1)}(k) & \text{for } i>0\end{cases}, \quad (10)$$

with

$$A^{(i)}=\frac{1}{N_c}\sum_{k'=0}^{N_c-1}\hat{H}^{(i)}(k'). \quad (11)$$

$\tilde{D}^{(i)}(k)$ is the kth frequency component of the transmitted replica which will be derived in Sect. II-*C*. After the residual ISI cancellation, N_c-point IFFT is applied to decompose $\{\tilde{R}^{(i)}(k);k=0\sim N_c-1\}$ into time-domain received signal before data-demodulation $\{\tilde{r}^{(i)}(n);n=0\sim N_c-1\}$. $\tilde{r}^{(i)}(n)$ can be given as

$$\tilde{r}^{(i)}(n)=\frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}\tilde{R}^{(i)}(k)\exp\left(j2\pi n\frac{k}{N_c}\right). \quad (12)$$

C. ISI Replica Generation

In this section, we explain how to generate ISI replica $\tilde{M}^{(i)}(k)$. By using the soft decision variable $\tilde{r}^{(i-1)}(n)$ associated with $d(n)$, the log-likelihood ratio (LLR) for the xth (x=0~$\log_2M-1$, where M is the modulation level) bit in the nth symbol $d(n)$ computed as [8]

$$\begin{aligned}\Lambda_n^{(i-1)}&=\ln\left(\frac{p(b_{n,x}=1)}{p(b_{n,x}=0)}\right)\\&\approx\frac{\left|\tilde{r}^{(i-1)}(n)-\sqrt{2E_s/T_s}A^{(i-1)}d_{b_{n,x}=0}^{\min}\right|^2}{2(\hat{\sigma}^{(i-1)})^2}-\frac{\left|\tilde{r}^{(i-1)}(n)-\sqrt{2E_s/T_s}A^{(i-1)}d_{b_{n,x}=1}^{\min}\right|^2}{2(\hat{\sigma}^{(i-1)})^2}\end{aligned}, \quad (13)$$

where $p^{(i-1)}(b_{n,x}=1)$ and $p^{(i-1)}(b_{n,x}=0)$ are the probabilities of transmitted bit $b_{n,x}$ is $b_{n,x}=1$ and $b_{n,x}=0$ at the $(i-1)$th iteration stage, respectively. $d_{b_{n,x}=0}^{\min}$ (or $d_{b_{n,x}=1}^{\min}$) is the symbol which has the shortest Euclidean distance from $\tilde{r}^{(i-1)}(n)$ among all candidate symbols with $b_{n,x}$=0 (or 1). $2(\hat{\sigma}^{(i-1)})^2$ is the variance of the noise plus residual ISI and is given by [10]

$$\begin{aligned}2(\hat{\sigma}^{(i-1)})^2&=\frac{2E_s}{T_s}\rho^{(i-1)}\left[\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|\hat{H}^{(i-1)}(k)\right|^2-\left|A^{(i-1)}\right|^2\right]\\&\quad+\frac{2N_0}{T_s}\frac{1}{N_c}\sum_{k=0}^{N_c-1}\left|W_R^{(i-1)}(k)\right|^2\end{aligned}, \quad (14)$$

where $\rho^{(i-1)}$ will be derived in Sect. II-*D*.

For QPSK data modulation, the transmitted replica $\{\tilde{d}^{(i-1)}(n);n=0\sim N_c-1\}$ is given by [8]

$$\tilde{d}^{(i-1)}(n) = \frac{1}{\sqrt{2}}\tanh\left(\frac{\Lambda_0^{(i-1)}(n)}{2}\right) + j\frac{1}{\sqrt{2}}\tanh\left(\frac{\Lambda_1^{(i-1)}(n)}{2}\right) . \quad (15)$$

Then, N_c-point FFT is applied to transform the transmitted replica $\tilde{d}^{(i-1)}(n)$ into N_c frequency components $\{\tilde{D}^{(i)}(k); k = 0 \sim N_c - 1\}$ as

$$\tilde{D}^{(i-1)}(k) = \frac{1}{\sqrt{N_c}}\sum_{n=0}^{N_c-1}\tilde{d}^{(i-1)}(n)\exp\left(-j2\pi k\frac{n}{N_c}\right). \quad (16)$$

Substituting (16) into (10), we obtain the frequency-domain ISI replica $\tilde{M}^{(i)}(k)$.

D. MMSE Weight Derivation

MMSE-FDE weight $\{W_R^{(i)}(k); k = 0 \sim N_c - 1\}$ which minimizes the mean square error (MSE) between the frequency-domain signal after ISI cancellation $\tilde{R}^{(i)}(k)$ and the transmitted symbol $D(k)$ can be derived as [8]

$$W_R^{(i)}(k) = \frac{W_T^*(k)H^*(k)}{\rho^{(i-1)}\left|W_T(k)\right|^2\left|H(k)\right|^2 + \left(E_s/N_0\right)^{-1}}, \quad (17)$$

with

$$\rho^{(i-1)} = \frac{1}{N_c}\sum_{n=0}^{N_c-1}\left\{E[|\,d(n)\,|^2] - |\,\tilde{d}^{(i-1)}(n)\,|^2\right\}. \quad (18)$$

where $E[|\,d(n)\,|^2] = 1$ in case of QPSK data modulation.

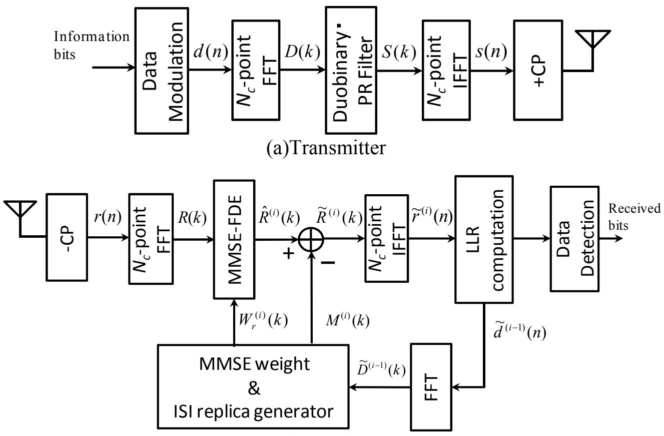

Fig. 1 System model of SC-FDE using duobinary PR filter with frequency-domain iterative ISI cancellation.

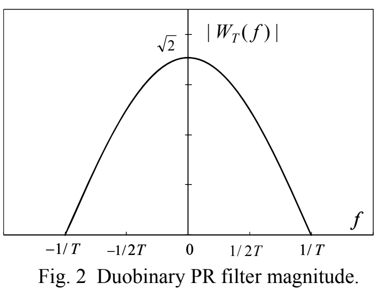

Fig. 2 Duobinary PR filter magnitude.

III. SC-FDE WITH MLSE

The transmission system model for SC-FDE using duobinary PR filter with MLSE is illustrated in Fig. 3. The transmitter is exactly the same as Sect. II. At the receiver, MMSE-FDE with the FDE weight $\{W_R'(k); k = 0 \sim N_c - 1\}$ is applied, which the weight is calculated so as to minimizes the MSE between the frequency-domain signal after MMSE-FDE $\hat{R}(k)$ and the transmitted symbol $S(k)$, can be derived as [8]

$$W_R'(k) = \frac{H^*(k)}{\left|H(k)\right|^2 + \left(E_s/N_0\right)^{-1}}. \quad (19)$$

A. MLSE

The time-domain signal $\{\hat{r}(n); n = 0 \sim N_c - 1\}$ can be expressed as

$$\begin{aligned}\hat{r}(n) &= \frac{1}{\sqrt{N_c}}\sum_{k=0}^{N_c-1}\hat{R}(k)\exp\left(j2\pi n\frac{k}{N_c}\right) \\ &= \sqrt{\frac{2E_s}{T_s}}As(n) + \mu'(n) + \hat{\eta}'(n)\end{aligned} \quad (20)$$

with

$$\begin{cases} A = \dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}H(k)W_R'(k) \\ \mu'(n) = \sqrt{\dfrac{2E_s}{T_s}}\left\{\dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}\hat{H}(k)\right\}\displaystyle\sum_{\substack{n'=0 \\ n\neq n'}}^{N_c-1}d(n')\exp\left(-j2\pi k\frac{(n'-n)}{N_c}\right), \\ \hat{\eta}'(n) = \dfrac{1}{N_c}\displaystyle\sum_{k=0}^{N_c-1}\hat{\Pi}(k)\exp\left(j2\pi n\frac{k}{N_c}\right) \end{cases}$$

(21)

where $\mu'(n)$ and $\hat{\eta}'(n)$ are the residual ISI and the noise after performing MMSE-FDE, respectively. In this paper, $\mu'(n)$ and $\hat{\eta}'(n)$ are modeled as independent zero-mean complex-valued Gaussian variables. The residual ISI plus noise component $\tilde{\eta}'(n) = \mu'(n) + \hat{\eta}'(n)$ can be treated as a new zero-mean complex-valued Gaussian variable [10]. The variance of $\tilde{\eta}'(n)$ is given by

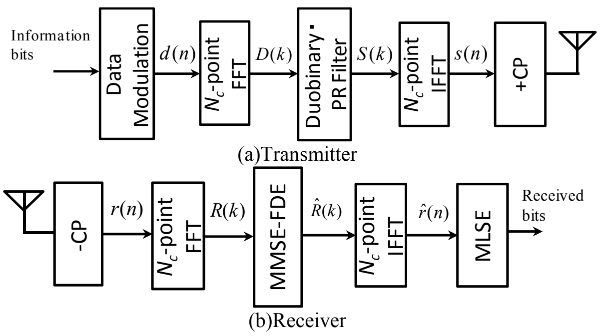

Fig. 3 SC-FDE using duobinary PR filter with MLSE.

$$2\hat{\sigma}'^2 = E[|\tilde{\eta}'(n)|^2] = 2\sigma'^2_{ISI} + 2\sigma'^2_{noise} \quad (22)$$

with

$$\begin{cases} 2\sigma'^2_{ISI} = \dfrac{2E_s}{T_s}\left[\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}\left|H(k)W'_R(k)\right|^2 - |A|^2\right] \\ 2\sigma'^2_{noise} = \dfrac{2N_0}{T_s}\dfrac{1}{N_c}\sum_{k=0}^{N_c-1}\left|W'_R(k)\right|^2 \end{cases}. \quad (23)$$

From (22), (20) can be rewritten as

$$\hat{r}(n) = \sqrt{\frac{2E_s}{T_s}}\frac{A}{\sqrt{2}}\left(d(n) + d(n-1 \mod . N_c)\right) + \tilde{\eta}(n). \quad (24)$$

It is observed that $s(n) = (1/\sqrt{2})\left(d(n) + d(n-1 \text{ mod}.N_c)\right)$ is in (24) as the characteristic of time-domain transmit symbol after applying duobinary PR signal [1].

After obtaining the output after applying MMSE-FDE, the posterior probability of the symbol candidate $d_i(n)$ (i=0~$\log_2 M$–1, where M is the modulation level) can be calculated by referencing to the Beyes' theorem as

$$P(d_i(n)\,|\,\hat{r}(n)) = \frac{P(d_i(n))p(\hat{r}(n)\,|\,d_i(n))}{p(\hat{r}(n))}, \quad (25)$$

where $P(d_i(n))$ and $p(\hat{r}(n))$ are the probability of the symbol candidate $d_i(n)$ and the probability density function (pdf) of the time-domain signal after MMSE-FDE, respectively, $p(\hat{r}(n)\,|\,d_i(n))$ is the conditional pdf given by

$$p(\hat{r}(n)\,|\,d_i(n)) = \frac{1}{2\pi\hat{\sigma}'^2}\exp\left(-\frac{\left|\hat{r}(n) - \sqrt{\frac{E_s}{T_s}}A(d_i(n) + d_i(n-1 \mod . N_c))\right|^2}{2\hat{\sigma}'^2}\right). \quad (26)$$

We assume that the M signals are equally probable *a priori*, i.e., $P(d_i(n)) = 1/M$ for all M. Therefore, when $p(\hat{r}(n)\,|\,d_i(n))$ becomes maximum, $P(d_i(n)\,|\,\hat{r}(n))$ also becomes maximum. The nth symbol $d_i(n)$ being transmitted are statistically independent Gaussian variables. The conditional pdf of the random variables $\hat{\mathbf{r}} = [\hat{r}(0), \hat{r}(1), \ldots, \hat{r}(N_c-1)]$ are

$$p(\hat{\mathbf{r}}\,|\,\mathbf{d}_i) = \prod_{n=0}^{N_c-1} p(\hat{r}(n)\,|\,d_i(n)). \quad (27)$$

The natural logarithm of (27) is given by

$$\ln p(\hat{\mathbf{r}}\,|\,\mathbf{d}_i) = 2N_c \ln\left(\frac{1}{\sqrt{2\pi}\hat{\sigma}'}\right) - \frac{f(\mathbf{d}_i)}{2\hat{\sigma}'^2} \quad (28)$$

with

$$f(\mathbf{d}_i) = \sum_{n=0}^{N_c-1}\left|\hat{r}(n) - \sqrt{\frac{E_s}{T_s}}A\left(d_i(n) + d_i(n-1 \mod . N_c)\right)\right|^2, \quad (29)$$

where $f(\mathbf{d}_i)$ is the distance metrics. To maximize $p(\hat{r}(n)\,|\,d_i(n))$, $\mathbf{d}_i$ is selected so as to minimize $f(\mathbf{d}_i)$. In this paper, $d(N_c-1)$ is initially assumed to be known.

IV. Computer Simulation

The simulation parameters are summarized in Table I. We assume QPSK data modulation, FFT block size N_c=256, GI size N_g=16, and block Rayleigh fading channel having symbol-spaced 16-path uniform power delay profile. Ideal channel estimation is also assumed.

The average BER performance of SC-FDE with frequency-domain ISI cancellation is plotted in Fig. 4 as a function of the average received bit energy-to-AWGN noise power spectrum density ratio E_b/N_0, defined as E_b/N_0=((1/2)(E_s/N_0)(1+N_g/N_c)). For comparison, the BER performance of SC-FDE using ideal brick-wall filter is also plotted. It can be seen from Fig. 4 that the transmission using ISI cancellation can significantly improve the BER performance. For example, the required E_b/N_0 for achieving BER=10^{-4} when I=1 reduces from the transmission without ISI cancellation (I=0) about 3dB. However, when I=2, there is no improvement from I=1 due to the effect of the wrong symbol replica generation.

The average BER performance of SC-FDE with MLSE is plotted in Fig. 5 as a function of E_b/N_0, defined as E_b/N_0=((1/2)(E_s/N_0)(N_c+N_g)/(N_c–1)) while considering that $d(N_c-1)$ is initially assumed to be known. For comparison, the BER performance of SC-FDE using duobinary PR filter with precoding [1] and that of SC-FDE using ideal brick-wall filter are also plotted. It can be seen from Fig. 5 that the MLSE can reduce the required E_b/N_0 for BER=10^{-4} about 2dB compared with the transmission using precoding.

Fig. 6 illustrates the performance tradeoff between 3dB-bandwidth versus required E_b/N_0 for achieving BER=10^{-4}. SC-FDE using duobinary PR filter with MLSE provides better trade-off since it can narrow the 3dB-bandwidth to be a half of one for the conventional SC-FDE using ideal brick-wall filter, while the required E_b/N_0 for BER=10^{-4} is only 3dB higher.

V. Conclusion

In this paper, we evaluated the performance of duobainary PR filtered SC-FDE which achieves higher spectral efficiency. By employing either the frequency-domain ISI cancellation or the MLSE, the BER performance improves compared with the transmission employing only the MMSE-FDE and gets closer to the conventional SC-FDE using ideal brick-wall filter. For system realization, the effect of channel estimation error will be considered as our future works.

TABLE I. COMPUTER SIMULATION CONDITION

Transmitter	Modulation	QPSK
	FFT block size	N_c=256
	GI size	N_g=16
	Filter	Duobinary PR
Channel	Fading type	16-path block Rayleigh
	Power delay profile	Uniform
	Time delay	$\tau_l=l$ (l=0~L-1)
Receiver	Channel estimation	Ideal
	Equalization	MMSE-FDE and ISI cancellation, MMSE-FDE and MLSE

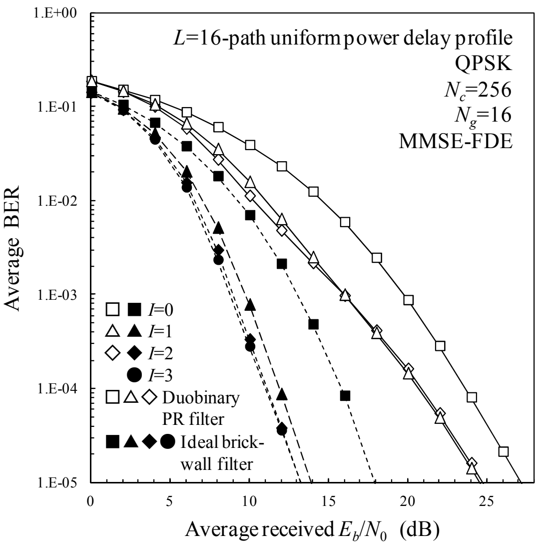

Fig. 4 BER performance with ISI cancellation.

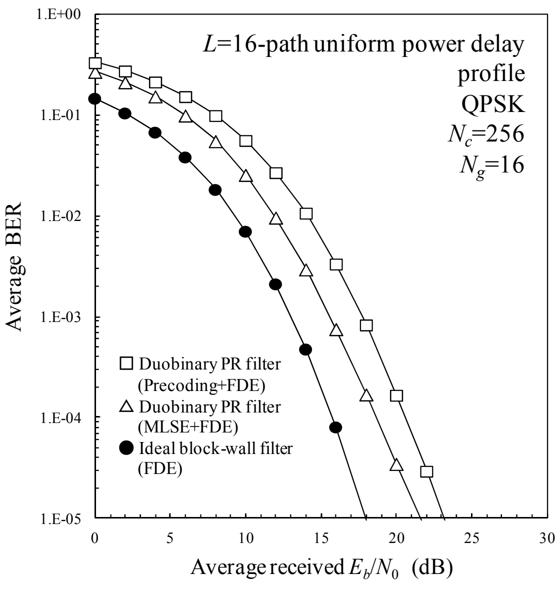

Fig. 5 BER performance with MLSE.

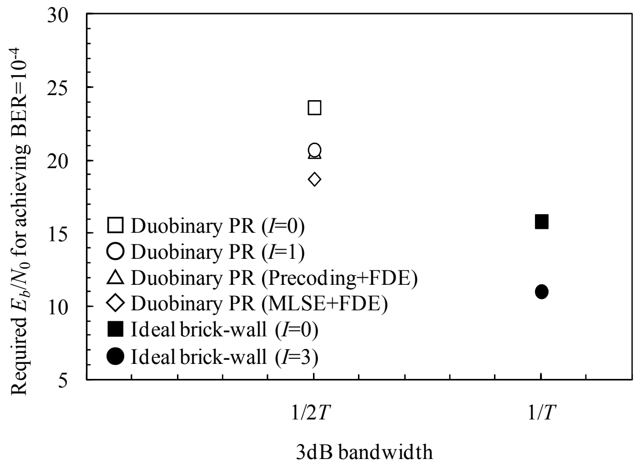

Fig. 6 the 3dB bandwidth- required E_b/N_0 for achieving BER=10^{-4}.

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