A Novel Receiver Design for Training Sequence Inserted OFDM Transmission

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Abstract—**Orthogonal frequency division multiplexing (OFDM) has been attracting much attention due to the robustness against the frequency selective fading. Instead of well-known cyclic prefix (CP) insertion, a known training sequence (TS) insertion can also be used for OFDM block transmission. In this paper, we propose a new receiver design, which can obtain the frequency diversity gain through the use of frequency-domain equalization (FDE), for TS-OFDM. The conditional bit error rate (BER) analysis of the proposed FDE is presented. The average BER performance of the TS-OFDM signal transmission in a frequency-selective Rayleigh fading channel is evaluated by Monte-Carlo numerical computation method using the derived conditional BER and is confirmed by computer simulation. Numerical results show the proposed TS-OFDM receiver improves BER and throughput performances of TS-OFDM compared to the conventional TS-OFDM receiver. It was also shown that the TS-OFDM with the proposed receiver provides better BER and throughput performances than the CP-OFDM due to the frequency diversity gain.**

Keywords-component; OFDM, frequency-domain equalization, training sequence

I. INTRODUCTION

Since the mobile wireless channel is composed of many propagation paths with different time delays, the broadband channel becomes severely frequency-selective [1]. To avoid problems arising from the severe frequency-selective fading channel, orthogonal frequency division multiplexing (OFDM) [2, 3] has been attracting much attention. OFDM is a block transmission using a number of orthogonal subcarriers. Before the transmission, the cyclic prefix (CP) is often inserted into guard interval (GI) placed in front of each OFDM signal to make the received OFDM signal to be a circular convolution of the transmitted OFDM signal and the channel impulse response. Each data symbol in a block is transmitted in parallel using a different orthogonal subcarrier and hence, simple one-tap frequency-domain equalization (FDE) (i.e., zero-forcing (ZF)-FDE) can be used.

Instead of CP insertion, a known training sequence (TS) insertion [4, 5] can be used. Since CP is replaced by TS, the TS acts as GI. Furthermore, the TS can be utilized for channel estimation and therefore, no pilot block is needed or no pilot subcarrier is needed unlike CP inserted OFDM transmissions. In the conventional TS-OFDM, receiver performs overlap and add (OLA) processing [5, 6], in which the GI part of the next OFDM signal is added to the beginning of the present OFDM signal to make the received OFDM signal to be a circular convolution of the transmitted OFDM signal and the channel impulse response. After OLA processing, the same signal processing as the CP-OFDM (i.e, discrete Fourier transform (DFT) and simple one-tap ZF-FDE) can be applied. The conventional TS-OFDM with OLA processing cannot obtain the frequency diversity gain similar to the CP-OFDM.

In this paper, we propose a novel receiver design, which can obtain the frequency diversity gain through the use of FDE for TS-OFDM. The conditional bit error rate (BER) analysis is presented. The average BER performance in a frequency-selective Rayleigh fading channel is evaluated by Monte-Carlo numerical computation method using the derived conditional BER and is confirmed by computer simulation of the TS-OFDM signal transmission.

Packet access will be the core technology of the next generation mobile data communication systems. High-speed packet transmissions can be realized by the use of hybrid automatic repeat request (HARQ). HARQ using incremental redundancy (IR) strategy is known to achieve the high throughput performance [7]. In this paper, the throughput performance of TS-OFDM with HARQ is also presented. Since the TS-OFDM with the proposed receiver design can obtain the frequency diversity gain, the throughput of HARQ with IR strategy can be improved compared to the TS-OFDM with the OLA processing and CP-OFDM.

The rest of the paper is organized as follows. In Section II, transmission system model of the proposed TS-OFDM with FDE is presented. Section III derives the conditional BER analysis. In Section IV, the performance evaluation is shown, and finally Section V concludes the paper.

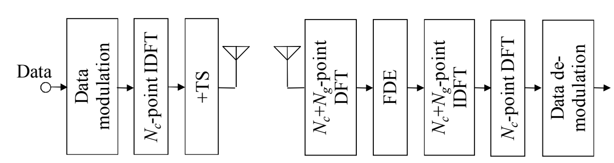

Figure 1. Transmission system model of TS-OFDM with FDE.

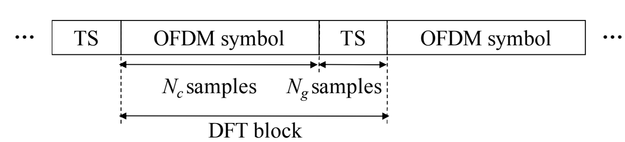

Figure 2. Block structure.

II. TS-OFDM TRASNMISSION WITH FDE

A. Transmit and received signal representations

Transmission system model of the proposed TS-OFDM with FDE is illustrated in Fig. 1. We consider an OFDM transmission with N_c subcarriers. At the transmitter, information bit sequence is transformed into a data-modulated symbol sequence. Then, the data-modulated symbol sequence is divided into a sequence of symbol blocks of N_c symbols each. The data symbol block is expressed using the vector form as $\mathbf{D}=[D(0),\ldots,D(i),\ldots,D(N_c-1)]^T$, where $(.)^T$ expresses the transposition. The data symbol block $\mathbf{D}$ is transformed into the time-domain OFDM symbol block $\mathbf{d}=[d(0),\ldots,d(t),\ldots,d(N_c-1)]^T$ by using an N_c-point inverse DFT (IDFT), which is given by

$$\mathbf{d}=\mathbf{F}_{N_c}^H\mathbf{D}\,, \tag{1}$$

where $(.)^H$ is the Hermitian transpose operation and $\mathbf{F}_K$ is the DFT matrix of size $K\times K$ given by

$$\mathbf{F}_K=\frac{1}{\sqrt{K}}\begin{bmatrix}1 & 1 & \cdots & 1\\ 1 & e^{-j2\pi\frac{1\times 1}{K}} & \cdots & e^{-j2\pi\frac{1\times(K-1)}{K}}\\ \vdots & \vdots & \ddots & \vdots\\ 1 & e^{-j2\pi\frac{(K-1)\times 1}{K}} & \cdots & e^{-j2\pi\frac{(K-1)\times(K-1)}{K}}\end{bmatrix}. \tag{2}$$

Before the transmission, the TS of length N_g ($\geq L$) samples is appended at the end of each OFDM symbol block, where L denotes the channel length. The block $\mathbf{s}=[s(0),\ldots,s(t),\ldots,s(N_c+N_g-1)]^T$ to be transmitted is expressed as

$$\mathbf{s}=\begin{bmatrix}\mathbf{d}\\ \mathbf{u}\end{bmatrix}=\begin{bmatrix}\mathbf{F}_{N_c}^H\mathbf{D}\\ \mathbf{u}\end{bmatrix}, \tag{3}$$

where $\mathbf{u}=[u(0),\ldots,u(t),\ldots,u(N_g-1)]^T$ denotes the TS vector which is identical for all blocks. The TS-OFDM block structure is illustrated in Fig. 2. The difference from CP-OFDM transmission is that CP is replaced by TS.

The OFDM symbol block is transmitted over a frequency-selective fading channel. The propagation channel is assumed to be a frequency-selective block fading channel composed of sample-spaced L distinct propagation paths. The channel impulse response $h(\tau)$ is given by

$$h(\tau)=\sum_{l=0}^{L-1}h_l\delta(\tau-\tau_l)\,, \tag{4}$$

where h_l and τ_l are respectively the complex-valued path gain with $E[\sum_{l=0}^{L-1}|h_l|^2]=1$ and the time delay of the l-th path. We assume the l-th path has a time delay of l samples, i.e. ($\tau_l=l$). The received signal block $\mathbf{y}=[y(0),\ldots,y(t),\ldots,y(N_c+N_g-1)]^T$ can be expressed as

$$\mathbf{y}=\sqrt{\frac{2E_s}{T_s}}\mathbf{hs}+\mathbf{n}\,, \tag{5}$$

where E_s and T_s are respectively the symbol energy and duration, $\mathbf{h}$ is the $(N_c+N_g)\times(N_c+N_g)$ channel impulse response matrix given as

$$\mathbf{h}=\begin{bmatrix}h_0 & & & & h_{L-1} & \cdots & h_1\\ h_1 & h_0 & & \mathbf{0} & & \ddots & \vdots\\ \vdots & h_1 & \ddots & & & & h_{L-1}\\ h_{L-1} & \vdots & \ddots & \ddots & & & \\ & h_{L-1} & & \ddots & \ddots & & \\ & & \ddots & & \ddots & \ddots & \\ \mathbf{0} & & & h_{L-1} & \cdots & h_1 & h_0\end{bmatrix} \tag{6}$$

and $\mathbf{n}=[n(0),\ldots,n(t),\ldots,n(N_c+N_g-1)]^T$ is the noise vector. The t-th element, $n(t)$, of $\mathbf{n}$ is the zero-mean additive white Gaussian noise (AWGN) having the variance $2N_0/T_s$ with N_0 being the one-sided noise power spectrum density.

B. FDE

For TS-OFDM, the same N_g-sample TS is inserted to the end of each N_c-sample OFDM symbol block. Therefore, it can be viewed as a CP an N_c+N_g-sample block. The frequency-domain signal block obtained by (N_c+N_g)-point DFT $\mathbf{Y}=[Y(0),\ldots,Y(k),\ldots,Y(N_c+N_g-1)]^T$ is expressed as

$$\begin{aligned}\mathbf{Y}&=\mathbf{F}_{N_c+N_g}\mathbf{y}\\ &=\sqrt{\frac{2E_s}{T_s}}\mathbf{HS}+\mathbf{N}\end{aligned}\,, \tag{7}$$

where $\mathbf{S}=[S(0),\ldots,S(k),\ldots,S(N_c+N_g-1)]^T=\mathbf{F}_{N_c+N_g}\mathbf{s}$ represents the frequency-domain transmit block, $\mathbf{N}=[N(0),\ldots,N(k),\ldots,N(N_c+N_g-1)]^T=\mathbf{F}_{N_c+N_g}\mathbf{n}$ is the frequency-domain noise vector, and $\mathbf{H}=\mathbf{F}_{N_c+N_g}\mathbf{hF}_{N_c+N_g}^H$ is the channel gain matrix. Due to the circulant property of $\mathbf{h}$, the channel gain matrix $\mathbf{H}$ is diagonal. The k-th diagonal element of $\mathbf{H}$ is given by

$$H(k)=\sum_{l=0}^{L-1}h_l\exp\left(-j2\pi k\frac{\tau_l}{N_c+N_g}\right). \tag{8}$$

FDE is carried out to obtain

$$\hat{\mathbf{Y}}=\mathbf{WY}\,, \tag{9}$$

where $\mathbf{W}=diag[W(0),\ldots,W(k),\ldots,W(N_c+N_g-1)]$ is the FDE weight matrix. The k-th diagonal element of $\mathbf{W}$ is given by

$$W(k)=\frac{H^*(k)}{|H(k)|^2+(E_s/N_0)^{-1}}\,, \tag{10}$$

which is the minimum mean square error (MMSE) weight derived so as to the minimize the MSE between the transmit block $\mathbf{S}$ and output of the FDE $\hat{\mathbf{Y}}$.

$\hat{\mathbf{Y}}$ is transformed into a time-domain TS-OFDM signal block $\hat{\mathbf{s}}=[\hat{s}(0),\ldots,\hat{s}(t),\ldots,\hat{s}(N_c+N_g-1)]^T$ by (N_c+N_g)-point IDFT as

$$\hat{\mathbf{s}} = \mathbf{F}^{H}_{N_c+N_g}\hat{\mathbf{Y}}. \tag{11}$$

C. Data demodulation

The time-domain TS-OFDM signal block $\hat{\mathbf{s}}$ is divided into two parts; first N_c-sample signal block which corresponds to the OFDM data symbol block and last N_g-sample block which corresponds to the TS. Therefore, to obtain the decision variable vector $\hat{\mathbf{D}} = [\hat{D}(0),...,\hat{D}(i),...,\hat{D}(N_c-1)]^T$, N_c-point DFT is applied to transform N_c-sample signal block $\{\hat{s}(0),...,\hat{s}(t),...,\hat{s}(N_c-1)\}$ into N_c frequency components as

$$\hat{\mathbf{D}} = \mathbf{F}_{N_c}[\hat{s}(0),...,\hat{s}(t),...,\hat{s}(N_c-1)]^T. \tag{12}$$

In principle, the above described FDE process for TS-OFDM is similar to the FDE process for single-carrier (SC) block transmission [8, 9] except for the last DFT operation. Since the received TS-OFDM signal using N_c orthogonal subcarriers is decomposed by N_c+N_g-point DFT into more than N_c frequency components, the frequency diversity gain can be obtained unlike the CP-OFDM and the conventional TS-OFDM with OLA processing. It should be noted that the intersymbol interference (ISI) is also produced. It can also be noted that the computational complexity of the proposed FDE increases considerably compared to the conventional TS-OFDM with OLA processing because additional operations are IDFT and DFT only.

III. BER Analysis

A. Decision variable

From (7)-(12), the i-th subcarrier component $\hat{D}(i)$, i=0~N_c–1, can be written as

$$\begin{aligned}\hat{D}(i) &= \frac{1}{\sqrt{N_c}}\frac{1}{\sqrt{N_c+N_g}}\sqrt{\frac{2E_s}{T_s}}\sum_{k=0}^{N_c+N_g-1}\hat{H}(k)S(k)\\ &\quad\times\exp\left[j\pi(N_c-1)\left(\frac{k}{N_c+N_g}-\frac{i}{N_c}\right)\right]\Phi(k,i)\\ &\quad+\frac{1}{\sqrt{N_c}}\frac{1}{\sqrt{N_c+N_g}}\sum_{k=0}^{N_c+N_g-1}\hat{N}(k)\\ &\quad\times\exp\left[j\pi(N_c-1)\left(\frac{k}{N_c+N_g}-\frac{i}{N_c}\right)\right]\Phi(k,i)\end{aligned}, \tag{13}$$

where $\hat{H}(k) = W(k)H(k)$ and $\hat{N}(k) = W(k)N(k)$ and

$$\Phi(k,i) = \begin{cases} N_c & \text{if } k = (1+N_g/N_c)i \\ \dfrac{\sin\pi N_c\left(\dfrac{k}{N_c+N_g}-\dfrac{i}{N_c}\right)}{\sin\pi\left(\dfrac{k}{N_c+N_g}-\dfrac{i}{N_c}\right)} & \text{otherwise}\end{cases}. \tag{14}$$

As an example, $\Phi(k, i$=32) for N_c=64, N_g=16 is plotted in Fig. 3.

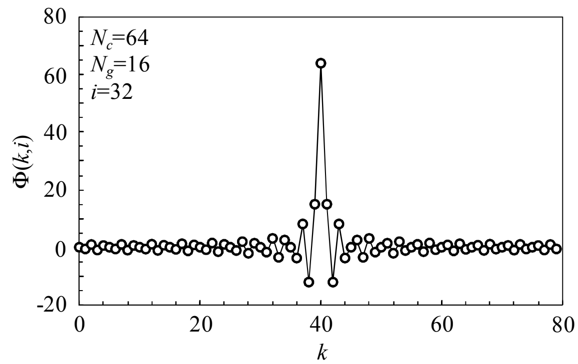

Figure 3. $\Phi(k, i$=32) for N_c=64, N_g=16.

Since the first term is decompose into the desired signal, residual ISI, and TS components, (13) can be rewritten as

$$\begin{aligned}\hat{D}(i) &= \sqrt{\frac{2E_s}{T_s}}\left(\frac{1}{N_c}\frac{1}{N_c+N_g}\sum_{k=0}^{N_c+N_g-1}\hat{H}(k)\Phi^2(k,i)\right)D(i)\\ &\quad+\mu_{TS}(i)+\mu_{ISI}(i)+\mu_{noise}(i)\end{aligned}, \tag{15}$$

where second, third, and fourth components are respectively denote the TS, residual ISI, and noise components. If the ideal knowledge of channel state information is assumed, TS component can be cancelled perfectly (i.e., $\mu_{TS}(i)$=0) and this can be performed simply in the frequency-domain as

$$\hat{\mathbf{Y}} = \mathbf{WY} - \sqrt{\frac{2E_s}{T_s}}\mathbf{WHF}_{N_c+N_g}\begin{bmatrix}\mathbf{0}\\ \mathbf{u}\end{bmatrix}. \tag{16}$$

Residual ISI and noise components are given by

$$\begin{cases}\mu_{ISI}(i) = \sqrt{\dfrac{2E_s}{T_s}}\sum\limits_{\substack{i'=0\\ \neq i}}^{N_c-1}d(i')\exp\left[j\pi(N_c-1)\dfrac{i-i'}{N_c}\right]\\ \quad\times\left(\dfrac{1}{N_c}\dfrac{1}{N_c+N_g}\sum\limits_{k=0}^{N_c+N_g-1}\hat{H}(k)\Phi(k,i)\Phi(k,i')\right)\\ \mu_{noise}(i) = \dfrac{1}{\sqrt{N_c}}\dfrac{1}{\sqrt{N_c+N_g}}\sum\limits_{k=0}^{N_c+N_g-1}\hat{N}(k)\\ \quad\times\exp\left[j\pi(N_c-1)\left(\dfrac{k}{N_c+N_g}-\dfrac{i}{N_c}\right)\right]\Phi(k,i)\end{cases}. \tag{17}$$

B. Conditional BER

We assume the Quadrature phase shift keying (QPSK) data modulation. It can be understood from (15) that $\hat{D}(i)$ is a complex-valued random variable with mean $\sqrt{2E_s/T_s}(1/N_c)(1/(N_c+N_g))\sum_{k=0}^{N_c+N_g-1}\hat{H}(k)\Phi^2(k,i)D(i)$. Approximating $\mu_{ISI}(i)$ and $\mu_{noise}(i)$ as a zero mean complex-valued

Gaussian variable, $\mu(i)=\mu_{ISI}(i)+\mu_{noise}(i)$ can be treated as a new zero mean complex-valued Gaussian variable. The variance $2\sigma_\mu^2(i)$ of $\mu(i)$ is given by $2\sigma^2_{\mu_{ISI}}(i)+2\sigma^2_{\mu_{noise}}(i)$. From (17), $\sigma^2_{\mu_{ISI}}(i)$ and $\sigma^2_{\mu_{noise}}(i)$ can be derived as (for the sake of brevity, the derivation is omitted)

$$\begin{cases} \sigma^2_{\mu_{ISI}}(i)=\dfrac{E_s}{T_s}\displaystyle\sum_{\substack{i'=0\\ \neq i}}^{N_c-1}\left|\frac{1}{N_c}\frac{1}{N_c+N_g}\sum_{k=0}^{N_c+N_g-1}\hat{H}(k)\Phi(k,i)\Phi(k,i')\right|^2 \\ \sigma^2_{\mu_{noise}}(i)=\dfrac{1}{N_c}\dfrac{1}{N_c+N_g}\dfrac{N_0}{T_s}\displaystyle\sum_{k=0}^{N_c+N_g-1}|W(k)\Phi(k,i)|^2 \end{cases}. \tag{18}$$

The conditional BER for the given E_s/N_0 and $\mathbf{H}$ can be given by [1]

$$p_b\left(\frac{E_s}{N_0},\mathbf{H}\right)=\frac{1}{2}erfc\left[\sqrt{\frac{1}{4}\gamma\left(i,\frac{E_s}{N_0},\mathbf{H}\right)}\right], \tag{19}$$

where $erfc(x)$ is the complementary error function and $\gamma(i, E_s/N_0, \mathbf{H})$ is the conditional signal-to-interference plus noise power ratio (SINR), which is given by

$$\gamma\left(i,\frac{E_s}{N_0},\mathbf{H}\right)=\frac{\dfrac{2E_s}{T_s}\left|\dfrac{1}{N_c}\dfrac{1}{N_c+N_g}\displaystyle\sum_{k=0}^{N_c+N_g-1}\hat{H}(k)\Phi^2(k,i)\right|^2}{\sigma^2_{\mu_{ISI}}(i)}$$
$$=\frac{\dfrac{2E_s}{N_0}\left|\dfrac{1}{N_c}\dfrac{1}{N_c+N_g}\displaystyle\sum_{k=0}^{N_c+N_g-1}\hat{H}(k)\Phi^2(k,i)\right|^2}{\left[\dfrac{E_s}{N_0}\displaystyle\sum_{\substack{i'=0\\ \neq i}}^{N_c-1}\left|\dfrac{1}{N_c}\dfrac{1}{N_c+N_g}\displaystyle\sum_{k=0}^{N_c+N_g-1}\hat{H}(k)\Phi(k,i)\Phi(k,i')\right|^2+\dfrac{1}{N_c}\dfrac{1}{N_c+N_g}\displaystyle\sum_{k=0}^{N_c+N_g-1}|W(k)|^2\,\Phi^2(k,i)\right]}. \tag{20}$$

The theoretical average BER is numerically computed by averaging (20) over all possible $\mathbf{H}$ and is confirmed by computer simulation in the next section.

TABLE I. NUMERICAL AND SIMULATION CONDITION

Transmitter	Data modulation	QPSK
	Number of subcarriers	N_c=64
	TS lengths	N_g=16
	TS	Chu sequence [10]
Channel	Fading type	Frequency-selective block Rayleigh
	Power delay profile	L=16 path uniform power delay profile
Receiver	Channel estimation	Ideal

IV. PERFORMANCE EVALUATION

The numerical and simulation conditions are summarized in Table I. QPSK data modulation, N_c=64, N_g=16, and 16-path frequency-selective block Rayleigh fading channel having uniform power delay profile are assumed. Ideal channel estimation and perfect DFT block synchronization at the receiver are assumed for the performance evaluation.

A. Average BER performance

Figure 4 plots the theoretical and simulated average BER performances of TS-OFDM with the proposed receiver as a function of the average received $E_b/N_0(=(E_s/N_0)(1+N_g/N_c)/2)$. For comparison, the BER performances of the TS-OFDM with OLA processing and the CP-OFDM are also plotted. It can be seen from Fig. 4 that a fairly good agreement between the theoretical and simulated results is seen. The proposed receiver can improve the BER performance of TS-OFDM than the conventional OLA processing. Also, the BER performance of TS-OFDM with the proposed receiver is better than that of the CP-OFDM. This is because the frequency diversity gain can be obtained.

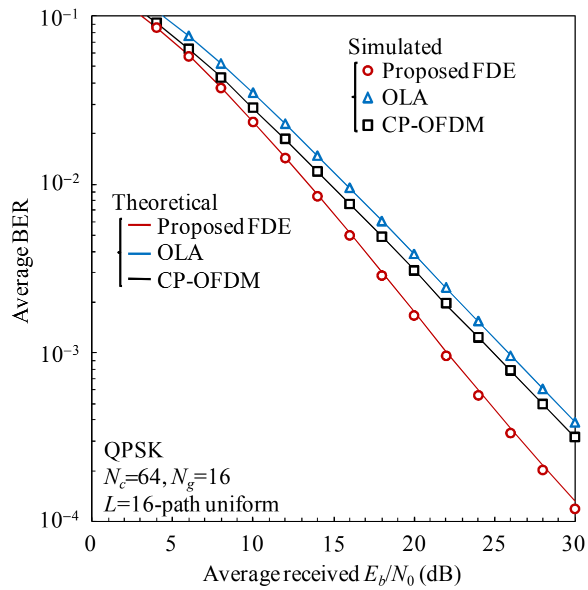

Figure 4. Average BER performacne of TS-OFDM with FDE.

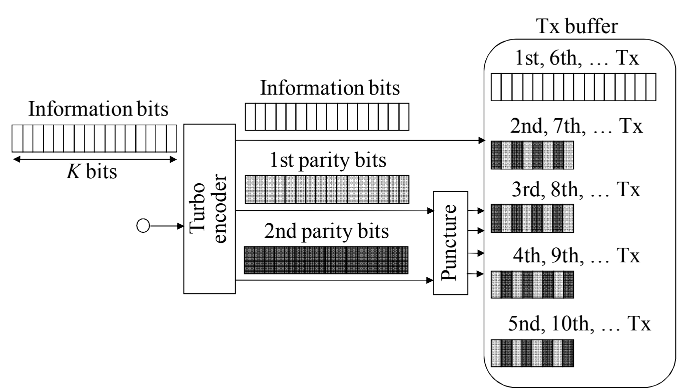

Figure 5. HARQ type II S-P4.

B. HARQ throughput performance

The throughput performance of TS-OFDM with HARQ using the proposed receiver is evaluated by computer simulation.

We employ a rate 1/3 turbo encoder using two (13, 15) recursive systematic convolutional (RSC) component encoders HARQ type II S-P4 [11] as illustrated in Fig. 5. Log-MAP decoding with 6 iterations is assumed. The packet size is set to K=512. The turbo encoder outputs the systematic bit sequence and two parity bit sequences. These sequences are punctured into five sequences (including systematic bit sequence) by the puncturing matrices given by

$$\begin{bmatrix}1&1&1&1\\0&0&0&0\\0&0&0&0\end{bmatrix}\begin{bmatrix}0&0&0&0\\1&0&0&0\\0&0&1&0\end{bmatrix}\begin{bmatrix}0&0&0&0\\0&1&0&0\\0&0&0&1\end{bmatrix}\begin{bmatrix}0&0&0&0\\0&0&1&0\\1&0&0&0\end{bmatrix}\begin{bmatrix}0&0&0&0\\0&0&0&1\\0&1&0&0\end{bmatrix}, \tag{21}$$

where the 1st, 2nd, and 3rd rows denote the puncturing pattern for the systematic bit sequence, 1st parity bit sequence, and 2nd parity bit sequence, respectively. For the first transmission, only the systematic bit sequence is transmitted. At the receiver, data decision and error detection are performed. If any error is detected in the received packet, second transmission is requested from the receiver by sending an NACK signal. When the NACK signal is received at the transmitter, the second packet (consisting of the punctured parity bit sequence) is transmitted. At the receiver, turbo decoding is carried out by using the first and second received packets. If any error is detected after turbo decoding, the NACK signal is transmitted again. One of the punctured parity bit sequences is transmitted each time the NACK signal is received at the transmitter until the 5th packet transmission. After the 5th packet transmission, the same packet is retransmitted.

The throughput performance of TS-OFDM with HARQ using the proposed receiver design is plotted in Fig. 6 as a function of average received symbol energy-to-noise power spectrum density ratio E_s/N_0. For comparison, the throughput performances of TS-OFDM using the OLA processing and CP-OFDM are also plotted. It can be seen form Fig. 6 that the proposed receiver design can achieve better throughput performance than the conventional OLA processing and improve the throughput compared to the CP-OFDM in high E_s/N_0 region due to the frequency diversity gain. TS-OFDM with the proposed receiver design provides at most the 15% improvement in throughput compared to TS-OFDM with OLA processing and the 6% improvement in throughput compared to CP-OFDM, respectively.

V. Conclusion

In this paper, we proposed a new receiver design for TS-OFDM with FDE which can obtain the frequency diversity gain. We presented the theoretical BER analysis and confirmed it by computer simulation. It was shown that the proposed receiver design improves BER and throughput performances of TS-OFDM compared to the conventional OLA processing. It was also shown that the TS-OFDM with the proposed receiver design provides better BER and throughput performances than the CP-OFDM due to the frequency diversity gain.

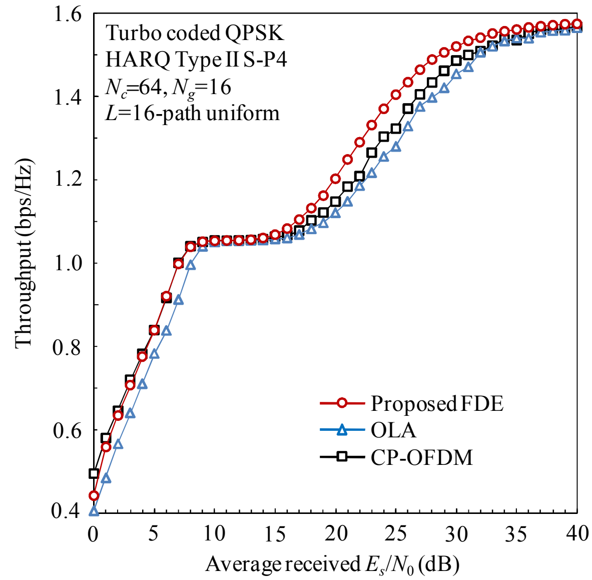

Figure 6. HARQ throughput performance of TS-OFDM with FDE.

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