Iterative Padding Subtraction of the PN Sequence for the TDS-OFDM over Broadcast Channels

Jun Wang, Zhi-Xing Yang, Chang-Yong Pan, Jian Song and Lin Yang

Abstract —*PN-sequence padding (PNP) TDS-OFDM transmission scheme has recently been proposed as an appealing alternative to the traditional Cyclic Prefix (CP) OFDM technology as it can provide significant improvement in the spectrum efficiency. In this paper, an iterative method is introduced to subtract the PN sequence padding for the TDS-OFDM systems to mitigate the residual inter symbol interference (ISI) from the imperfect channel estimation, and the inherent channel estimation techniques are also performed to track the channel changes. Results show that this iterative algorithm can effectively remove the impact from the residual ISI on the slow-fading broadcast channels, even in the presence of very large channel delays in the single-frequency simulcast networks (SFN)*[1].

Index Terms —PN sequence padding, TDS-OFDM, inter symbol interference (ISI), Zero Padding (ZP).

I. INTRODUCTION

Digital television (TV) broadcasting over terrestrial VHF and UHF radio channels requests a universal transmission format for both fixed and mobile receivers under the multipath propagation environment. The Digital Video Broadcasting for Terrestrial Television (DVB-T)[1], uses the cyclic prefix (CP) guard interval for the OFDM (orthogonal frequency division multiplexing) signal to avoid the multipath impact from the adjacent symbols and sends a significant amount of training symbols (more than 10% of the data symbols) for the better system performance. This, however leads to the non-negligible reduction in the channel throughput. To maintain the system performance with the minimal capacity reduction, a new scheme competing for the Chinese Terrestrial TV standard, Terrestrial Digital Multimedia Broadcasting (DMB-T), was proposed by Tsinghua University using the key technology of TDS-OFDM [2]. Instead of using CP as that in DVB-T, TDS-OFDM uses the pseudo-noise (PN) sequence padding as the guard interval, which also serves the purpose of the training symbols. The design can greatly reduce the overall transmission overhead with the typical improvement in pectrum efficiency between 5% and 15%, and have been demonstrated that it can provide even better performance [3].

Ideally, after subtracting the PN sequences from the TDS-OFDM signal, the remaining signal can be taken as the Zero Padding (ZP) OFDM signal and all the techniques for the ZP-OFDM signal processing in [4][5] can be applied. In practice, however, ideal performance can be realized if and only if the receiver has the perfect channel information. Otherwise, the impact of the residual inter-symbol-interference (ISI) could be devastating even in the small amount. A new method to mitigate the residual ISI is introduced in this paper, which is the iterative version of the echo cancellation method in the CP-OFDM systems with insufficient length of CP [6][7]. The details on performing the channel estimation as well as tracking the fading channel frame by frame using the periodical PN sequence are also addressed.

The remaining paper is organized as follows. Section II briefly describes the PN-sequence padding TDS-OFDM modulator. The iterative subtraction algorithm is introduced in Section III and its performance is evaluated in Section IV by the computer simulation. We then conclude this paper in Section V.

II. SYSTEM MODEL FOR PNP TDS-OFDM

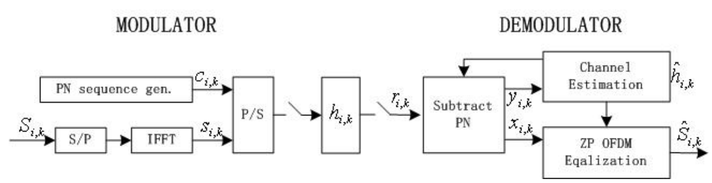

Fig. 1. Discrete model of TDS-OFDM

Figure 1 is the baseband equivalent TDS-OFDM system with the discrete-time blocks. The ith (i≥ 0) transmitted information sequence of $\{S_{i,k}\}_{k=0}^{N-1}$ is first modulated by the N-point IFFT:

$$s_{i,k} = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} S_{i,n} \exp\left\{ j \frac{2\pi nk}{N} \right\}, \quad 0 \le k < N \qquad (1)$$

The predefined PN sequence $\{c_{i,k}\}_{k=0}^{M-1}$ with length of M is then inserted before each IFFT output $\{s_{i,k}\}_{k=0}^{N-1}$. The corresponding transmitted signal frame vector is given in Figure 2.(a).

The PN sequences used in TDS-OFDM system are defined as a set of shifted m-sequence. They satisfy the following orthogonality condition to provide the unique frame address [1]:

$$c_{i,k} * c_{j,k} = \delta(i, j) \qquad (2)$$

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Here, * is the convolution operator and $\delta(i,j)$ is a modified delta function. With the above description, each OFDM frame can be decomposed into two non-overlapping parts in time domain: PN sequence $\{c_{i,k}\}_{k=0}^{M-1}$ and $\{s_{i,k}\}_{k=0}^{N-1}$, as illustrated in Figures 2.(b).

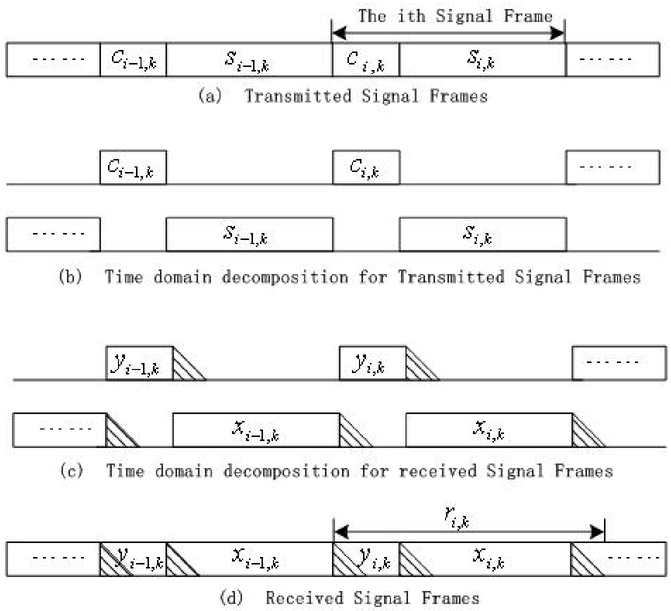

Fig. 2. Decomposition for transmitted and received signal frames

This OFDM signal then passes through a multipath fading channel. For the explanation purpose, the channel during the transmission of ith OFDM block can be modeled by a quasi static Lth-order FIR filter with the impulse response (CIR) $\{h_{i,k}\}_{k=0}^{L-1}$, we assume the normalized Doppler frequency $f_D N T_s$ is very small, and the inter-channel-interference (ICI) caused by the channel time variations over each OFDM frame can therefore be ignored, where f_D is the maximum Doppler frequency, and T_s is the baseband symbol rate.

In practice, the TDS-OFDM system is designed such that the duration of the PN sequence exceeds the channel memory, i.e. $M \ge L$. The ith received OFDM frame $\{r_{i,k}\}_{k=0}^{M+N+L-2}$ excluding noise now consists of two overlapping parts: $\{y_{i,k}\}_{k=0}^{M+L-2}$ representing the linear convolution output between the PN sequence $\{c_{i,k}\}_{k=0}^{M-1}$ and the channel CIR, and $\{x_{i,k}\}_{k=0}^{N+L-2}$ denoting the convolution output between the information sequence $\{s_{i,k}\}_{k=0}^{N-1}$ and the channel CIR. The expressions of these two parts are given by:

$$x_{i,k} = s_{i,k} * h_{i,k} = \sum_{l=0}^{L-1} s_{i,k-l} \cdot h_{i,l} \quad 0 \le k < N+L-1 \tag{3}$$

$$y_{i,k} = c_{i,k} * h_{i,k} = \sum_{l=0}^{L-1} c_{i,k-l} \cdot h_{i,l} \quad 0 \le k < M+L-1 \tag{4}$$

And the received ith signal frame $\{r_{i,k}\}_{k=0}^{M+N+L-2}$ with AWGN of $n_{i,k}$ can be expressed as:

$$r_{i,k} = u_{i,k} + n_{i,k} \quad 0 \le k < M+N+L-1 \tag{5}$$

Where,

$$u_{i,k} = \begin{cases} x_{i-1,k+N} + y_{i,k} & 0 \le k < L-1 \\ y_{i,k} & L-1 \le k < M \\ x_{i,k-M} + y_{i,k} & M \le k < M+L-1 \\ x_{i,k-M} & M+L-1 \le k < N+M \\ x_{i,k-M} + y_{i+1,k-N-M} & N+M \le k < N+M+L-1 \end{cases} \tag{6}$$

If $\{y_{i,k}\}_{k=0}^{M+L-2}$ can be subtracted from the received signal $\{r_{i,k}\}_{k=0}^{M+N+L-2}$, the remaining signal $\{x_{i,k}\}_{k=0}^{N+L-2}$ will be equivalent to the ZP-OFDM signal, and all the well-established methods related to the ZP-OFDM can be applied, for example, the corresponding ZF and MMSE equalizers provided in [4]. In the following section, the proposed iterative method for the PN-sequence subtraction and channel estimation are discussed in details.

III. ITERATIVE SUBTRACTION METHOD

The channel estimation in the real systems is generally imperfect, especially for the time-varying channels with long echoes. The proposed iterative algorithm subtracts the PN padding and performs channel estimation in an iterative manner. After several iterations, the channel estimation becomes more accuracy and the PN padding can be subtracted almost completely, and this can be summarized as follows:

(1) The subtraction is performed frame by frame, when we do with the ith signal frame, the estimations of the channel impulse response of $\{\hat{h}_{i-2,l}\}_{l=0}^{L-1}$, $\{\hat{h}_{i-1,l}\}_{l=0}^{L-1}$ and $\{\hat{x}_{i-1,k}\}_{k=0}^{N+L-2}$ for the previous two signal frames have already known. Assuming the channel echo delay is quite stable, i.e., unchanged during each OFDM symbol. Based on $\{\hat{h}_{i-2,l}\}_{l=0}^{L-1}$ and $\{\hat{h}_{i-1,l}\}_{l=0}^{L-1}$, via a linear interpolation for both echoes' amplitudes and phases, $\{\hat{h}_{i,l}^{iter=0}\}_{l=0}^{L-1}$ can be estimated [8]. Set the iteration index J to zero and go to step (2).

(2) $\{\hat{h}_{i+1,l}^{iter=J}\}_{l=0}^{L-1}$ is estimated via linear interpolation from $\{\hat{h}_{i-1,l}\}_{l=0}^{L-1}$ and $\{\hat{h}_{i,l}^{iter=J}\}_{l=0}^{L-1}$ in the same way as that in step (1).

(3) Since $\{c_{i,k}\}_{k=0}^{M-1}$ and $\{c_{i+1,k}\}_{k=0}^{M-1}$ associated with the signal frames are known after the synchronization, the linear convolution outputs $\{\hat{y}_{i,k}^{iter=J}\}_{k=0}^{M+L-2}$ and $\{\hat{y}_{i+1,k}^{iter=J}\}_{k=0}^{M+L-2}$ can be calculated with a N_1 points FFT respectively, where $N_1 \ge M+L-1$.

(4) Subtract $\left\{\hat{y}_{i,k}\right\}_{k=0}^{M+L-2}$ and $\left\{\hat{y}_{i+1,k}\right\}_{k=0}^{M+L-2}$ from the ith received signal frame $\left\{r_{i,k}\right\}_{k=0}^{M+N+L-2}$, and obtain an estimate of $\left\{\hat{x}_{i,k}^{iter=J}\right\}_{k=0}^{N+L-2}$, i.e.,

$$\hat{x}_{i,k}^{iter=J}=\begin{cases} r_{i,k+M}-\hat{y}_{i,k+M}^{iter=J} & 0\le k<L-1 \\ r_{i,k+M} & L-1\le k<N \\ r_{i,k+M}-\hat{y}_{i+1,k-N}^{iter=J} & N\le k<N+L-1 \end{cases} \tag{7}$$

(5) If iter J is equal to the predefined number of iterations J_0, stop the iteration, and $\left\{\hat{x}_{i,k}^{iter=J_0}\right\}_{k=0}^{N+L-2}$ becomes the final estimation of $\left\{x_{i,k}\right\}_{k=0}^{N+L-2}$ (i.e. $\left\{\hat{x}_{i,k}\right\}_{k=0}^{N+L-2}$) while $\left\{\hat{h}_{i,l}^{iter=J}\right\}_{l=0}^{L-1}$ is the final estimation of $\left\{h_{i,l}\right\}_{l=0}^{L-1}$ (i.e. $\left\{\hat{h}_{i,l}\right\}_{l=0}^{L-1}$). $\left\{\hat{x}_{i,k}\right\}_{k=0}^{N+L-2}$ is then equivalent to ZP-OFDM signal. From calculation complexity point of view, the ZP-OFDM-OLA equalization method [4] is used here. After decisions on $\left\{\hat{x}_{i,k}\right\}_{k=0}^{N+L-2}$ are made, move on to the (i+1)th signal frame and start again from 1). Else if iter J is less then J_0, go to step (6).

(6) A time domain filter and decision feedback method is used to remove the residual ISI and the noise components from $\left\{\hat{x}_{i,k}^{iter=J}\right\}_{k=0}^{N+L-2}$ with given $\left\{z_{i,k}^{iter=J}\right\}_{k=0}^{N+L-2}$. The corresponding filter method is described in Section III-B.

(7) Reconstruct the $\left\{\hat{y}_{i,k}^{iter=J+1}\right\}_{k=0}^{M+L-2}$ as

$$\hat{y}_{i,k}^{iter=J+1}=\begin{cases} r_{i,k}-\hat{x}_{i-1,k+N} & 0\le k<L-1 \\ r_{i,k} & L-1\le k<M \\ r_{i,k}-z_{i,k-M}^{iter=J} & M\le k<M+L-1 \end{cases} \tag{8}$$

Where $\left\{\hat{x}_{i-1,k}\right\}_{k=0}^{N+L-2}$ is the final estimation of $\left\{x_{i-1,k}\right\}_{k=0}^{N+L-2}$ for the last (i-1)th signal frame.

(8) From $\left\{\hat{y}_{i,k}^{iter=J+1}\right\}_{k=0}^{M+L-2}$, a more accurate channel response, $\left\{\hat{h}_{i,l}^{iter=J+1}\right\}_{l=0}^{L-1}$, is obtained. The corresponding channel estimation method is described in Section III-A. Set J <=J+1 and jump to step (2) for iteration.

A. Channel estimation methods

Channel characteristics can be estimated either frame by frame or over a group of frames, using the periodically transmitted PN sequence associated with the signal frame. The estimation can be performed either in time domain or frequency domain.

The time domain method uses the cross correlation operation. At the receive end, the received PN sequence $r_{i,k}$ can be approximately expressed as:

$$r_{i,k}=c_{i,k}*h_{i,k}+mp_{i,k},\ 0\le k<M+L-1 \tag{9}$$

Where $mp_{i,k}$ denotes the ISI from the previous frame body of the data signal. Because of the orthogonal construction of each PN sequence in Eq.(2), the cross correlation of the received PN sequence and the locally stored PN sequence can be expressed as:

$$\begin{aligned} c_{i,k}*r_{i,k}&=c_{i,k}*(c_{i,k}*h_{i,k})+c_{i,k}*mp_{i,k} \\ &\approx h_{i,k} \end{aligned} \tag{10}$$

As the length of the FIR filter characterizing the channel may cause the implementation difficult, the cross-correlation method is not suitable for the iteration operation, so this time domain approach is only used for the initial channel estimation. During the iteration process, channel estimation is implemented in the frequency domain with FFT approach.

In step (8), $\left\{\hat{y}_{i,k}^{iter=J+1}\right\}_{k=0}^{M+L-2}$ is the estimation of the linear convolution result of $\left\{c_{i,k}\right\}_{k=0}^{M-1}$ and $\left\{h_{i,l}\right\}_{l=0}^{L-1}$, we conduct $\left\{Y_{i,k}^{iter=J+1}\right\}_{k=0}^{N_1-1}$ and $\left\{C_{i,k}\right\}_{k=0}^{N_1-1}$ are respectively the N_1-point FFT of $\left\{\hat{y}_{i,k}^{iter=J+1}\right\}_{k=0}^{M+L-2}$ and $\left\{c_{i,k}\right\}_{k=0}^{M-1}$ being zero-padded to the length of N_1. The channel estimation on $\left\{\hat{h}_{i,l}^{iter=J+1}\right\}_{l=0}^{N_1-1}$ is given by:

$$\hat{h}_{i,k}^{iter=J+1}=IFFT\left\{\frac{Y_{i,k}^{iter=J+1}}{C_{i,k}}\right\},\quad 0\le k<N_1-1 \tag{11}$$

Then $\left\{\hat{h}_{i,l}^{iter=J+1}\right\}_{l=0}^{N_1-1}$ is passed through a rectangular window to set $\hat{h}_{i,l}^{iter=J+1}=0$ for $l\ge L$ and the result $\left\{\hat{h}_{i,l}^{iter=J+1}\right\}_{l=0}^{L-1}$ will be used for the next iteration.

B. Filter for $\left\{\hat{x}_{i,k}^{iter=J}\right\}_{k=0}^{N+L-2}$

In step (6), after remove the impact of the header (i.e., the PN sequence), $\left\{\hat{x}_{i,k}^{iter=J}\right\}_{k=0}^{N+L-2}$ is still contaminated by the noise as well as the residual ISI, and further filtering is needed to reduce the noise components as follows:

(1) Calculate the channel estimation $\left\{\tilde{h}_{i,l}^{iter=J}\right\}_{l=0}^{L-1}$ for $\left\{\hat{x}_{i,k}^{iter=J}\right\}_{k=0}^{N+L-2}$. For the simplicity, the average of $\left\{\hat{h}_{i,l}^{iter=J}\right\}_{l=0}^{L-1}$ and $\left\{\hat{h}_{i+1,l}^{iter=J}\right\}_{l=0}^{L-1}$ is used here, i.e.

$$\tilde{h}_{i,l}^{iter=J}=(\hat{h}_{i,l}^{iter=J}+\hat{h}_{i+1,l}^{iter=J})/2 \tag{12}$$

(2) Equalize $\left\{\hat{x}_{i,k}^{iter=J}\right\}_{k=0}^{N+L-2}$ in frequency domain through N_2-points FFT and IFFT operations, with $N_2\ge N+L-1$ and the result $\left\{\hat{s}_{i,k}^{iter=J}\right\}_{k=0}^{N_2-1}$ is:

$$\hat{s}_{i,k}^{iter=J} = IFFT\left\{\frac{FFT\left(\hat{x}_{i,k}^{iter=J}\right)}{FFT\left(\tilde{h}_{i,k}^{iter=J}\right)}\right\}, \quad 0 \le k < N_2 \tag{13}$$

(4) Then $\left\{\hat{s}_{i,k}^{iter=J}\right\}_{k=0}^{N_2-1}$ is passed through a time domain rectangular window to make $\hat{s}_{i,k}^{iter=J} = 0$ for k ≥ N to obtain $\left\{\hat{s}_{i,k}^{iter=J}\right\}_{k=0}^{N-1}$.

To further remove the residual noise, $\left\{\hat{s}_{i,k}^{iter=J}\right\}_{k=0}^{N-1}$ is converted to frequency domain with N-point FFT to get $\left\{\hat{S}_{i,k}^{iter=J}\right\}_{k=0}^{N-1}$. After making decisions on $\left\{\hat{S}_{i,k}^{iter=J}\right\}_{k=0}^{N-1}$, the decisions are then changed back to time domain with the result of $\left\{\tilde{s}_{i,k}^{iter=J}\right\}_{k=0}^{N-1}$.

(5) The final filtered output $\left\{z_{i,k}^{iter=J}\right\}_{k=0}^{N+L-2}$ is the linear convolution between $\left\{\tilde{s}_{i,k}^{iter=J}\right\}_{k=0}^{N-1}$ and $\left\{\tilde{h}_{i,l}^{iter=J}\right\}_{l=0}^{L-1}$, which can also be calculated by multiplication operation after both being converted to frequency domain with N_2 -point FFT.

IV. Simulation

With all the analysis and discussions provided above, we now evaluate the performance of the proposed algorithm over two typical broadcast channels for the TDS-OFDM system, using major simulation parameters in Table I. The 1st channel is the Fixed Reception F1 version of DVB-T[1] channel model with the static impulse response in Table II. The 2nd multipath channel consists of a 0dB-echo with very long delay of 30us, it characterizes the SFN environment for the State Administration of Radio Film and Television (SARFT) in the Chinese DTV test report with the impulse response in Table III. We use the maximal Doppler frequencies f_D =10Hz (the normalized Doppler frequency $f_D NT_s = 0.005$), corresponding to the receiver velocity ranging from 13 to 23km/h in the TV UHF band (@470~862MHz).

TABLE I

MAJOR SIMULATION PARAMETERS

Symbol Rate $1/T_s$	7.56MSPS
Signal Constellation	QPSK, 16QAM, 64QAM
FFT Size N	3780
Sub carrier space	2KHz
PN Sequence Length M	420

TABLE II

THE 1ST CHANNEL MODEL

Multi-path propagation	Delay Time (T_s)	Relative Amplitude
1	0	1
2	2	0.225894
3	4	0.15034
4	5	0.051534
5	6	0.149723
6	7	0.170996
7	13	0.295723
8	16	0.407163
9	18	0.258782
10	19	0.221155
11	26	0.262909
12	28	0.24014
13	30	0.057662
14	31	0.061831
15	41	0.25973
16	42	0.116587
17	59	0.400967
18	83	0.303585
19	98	0.350825
20	101	0.185074
21	165	0.176809

TABLE III

THE 2ND CHANNEL MODEL

Multi-path propagation	Delay Time (T_s)	Relative Amplitude
1	14	1
2	0	0.126
3	15	0.1
4	27	0.1
5	57	0.316
6	241	1

In the simulations, we choose $N_1 = 2048$, $N_2 = 8096$, with 10,000 signal frames simulated. Figs. 3 and 4 illustrate the performance of the symbol-error-rate (SER) versus the average signal to noise ratio (SNR) in dB under the proposed algorithm for 1st and 2nd Channels, respectively, with no iteration (J_0=0), intermediate and final results of the three iterations (J_0=1, J_0=2 and J_0=3). It is quite clear that the proposed scheme offers huge improvement in the SER performance, especially after the first iteration. Several orders of magnitude improvement in the SER can be achieved for both channels after two or three iterations. For the 1st Channel without the 0dB-echoe, one iteration has already brought much better performance, average of 2 dB SNR improvement at SER=10^{-1} for QPSK, 16QAM and 64QAM Constellation, respectively, and more improvement can be achieved at high SNR. For the 2nd Channel with the 0dB-echo, one iteration is indispensable for the QPSK and 16QAM modulations while two iterations are required for 64QAM. More than 5 dB SNR improvement can be obtained for QPSK, and more than 3 dB for 16QAM and 64QAM, respectively. At high SNR the SER curves become flat because of the deep frequency null caused by the 0dB-echo.

V. CONCLUSIONS

This paper presented a method that can effectively remove the residual ISI for both static and slowly fading channels when CP guard interval is replaced by PN sequence padding. Simulation results show that the proposed technique can be applied to the TDS-OFDM terrestrial broadcasting systems with significant improvement in the spectrum efficiency and almost no SNR loss. The computational effort for the iteration process is quite limited and complexity is therefore very low.

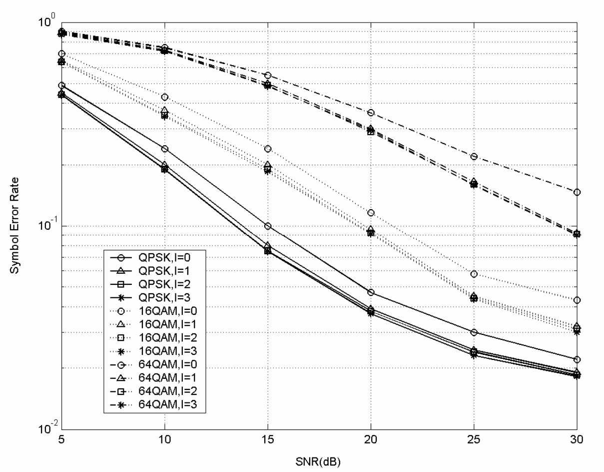

Fig. 3. SER performance of the iterative algorithm on the 1st channel.

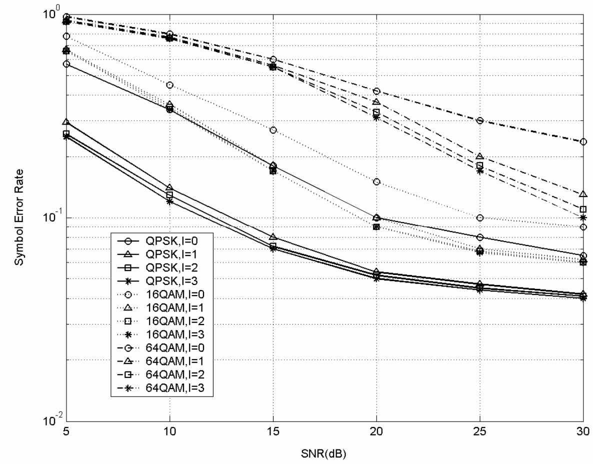

Fig. 4. SER performance of the iterative algorithm on the 2nd channel.

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