흐름 다중 심벌 검파를 사용한 트렐리스 부호화된 π/8 shift 8PSK-OFDM

안필승* · 김종일* · 김한종** · Zhengyuan Xu***

Trellis-coded π/8 shift 8PSK-OFDM with Sliding Multiple Symbol Detection

Phil-Seung Ahn*, Chong-Il Kim*, Han-Jong Kim**, Zhengyuan Xu***

요 약

본 논문에서는 π/4 shift QPSK를 트렐리스 부호화 변조에 적용시키기 위해 π/8 shift 8PSK와 BER 특성을 향상시키기 위한 트렐리스 부호화된 π/8 shift 8PSK-OFDM을 제안한다. 트렐리스 부호화 변조는 위상차에 의한 신호 집합 확장과 분할을 수행한다. 수신측에서 슬라이딩 방식의 다중 심벌 검파를 수행하기 위해서 연속 수신된 신호로부터 L개의 위상차를 추출하고 이를 이용한 비터비 디코더를 설계한다. 슬라이딩 방식의 다중 심벌 검파는 트렐리스 부호화된 π/8 shift 8PSK-OFDM에서 향상된 BER 성능을 보여준다. 본 논문에서 제안한 다중 심벌 검파를 이용한 π/8 shift 8PSK-OFDM은 대역폭과 전력의 효율성을 감소시키지 않고 같은 SNR에서 BER 성능을 향상시킬 수 있다는 것을 보여준다. 또한 제안된 디코더 방식과 알고리듬은 다중 반송파뿐만 아니라 전통적인 단일 반송파 변조에도 사용될 수 있다.

ABSTRACT

In this paper, we propose π/8 shift 8PSK and trellis-coded π/8 shift 8PSK-OFDM techniques by applying π/4 shift QPSK to trellis-coded modulation (TCM), and performing signal set expansion and set partition correspondingly based on phase difference. In our Viterbi decoding algorithm, up to L phase differences from successively received symbols are employed in the new branch metrics. Such sliding multiple symbol detection (SMSD) method provides improved bit-error-rate (BER) performance in the differential detection of the trellis-coded π/8 shift 8PSK-OFDM signals. The performance improvements are achieved for different communication channels without sacrificing bandwidth and power efficiency. It thus makes the proposed modulation and sliding detection scheme more attractive for power and band-limited systems.

키워드

TCM, 다중 심벌 검파, OFDM, 다중 반송파

I. Introduction

Recently, multicarrier modulation schemes, often called orthogonal frequency-division multiplexing (OFDM), have received a lot of attention in the radio communications and multimedia communications [1], [2], [3]. This is mainly because of the need to transmit high data rate in a mobile environment that makes a highly hostile radio channel [1]. The OFDM system has to cope with

* 관동대학교

** 한국기술교육대학교

*** University of California, Riverside(UCR)

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multipath propagation channels in a realistic radio environment. Intersymbol interference (ISI) can be avoided due to inserted guard interval. However frequency-selective fading may still cause the subcarriers strongly attenuated by the radio channel. Therefore it is difficult to achieve considerable data rate for a multimedia data transmission. To attack this problem, channel coding seems to be a solution. In order to gain high data rate, an efficient channel coding scheme has to be used. In 1982, the trellis-coded modulation (TCM) method was developed by Ungerboeck for power and bandwidth efficient digital communications [4] where the coding gain was improved by combining the channel coding and modulation.

In this paper, in order to apply the π/4 shift QPSK to TCM, we propose trellis-coded (TC) π/8 shift 8PSK and the π/8 shift 8PSK modulation techniques by performing signal set expansion and set partition by phase difference. Also, the Viterbi decoder with branch metrics of the squared Euclidean distance of the first up to the L-th phase differences is introduced in order to improve the BER performance in differential detection of TC π/8 shift 8PSK. Exploitation of branch metrics is thus well integrated in our detailed Viterbi algorithm.

II. Transmission System and Trellis-Coded π/8 shift 8PSK-OFDM

Fig. 1 shows the considered TC π/8 shift 8PSK-OFDM transmission system. The coded bit sequence is mapped into π/8 shift 8PSK modulation signals by the signal mapper for each subcarrier, i.e., for the subcarrier in the current OFDM symbol, the bits z_k^0, z_k^1, z_k^2 are mapped onto the phase difference symbol x_k.

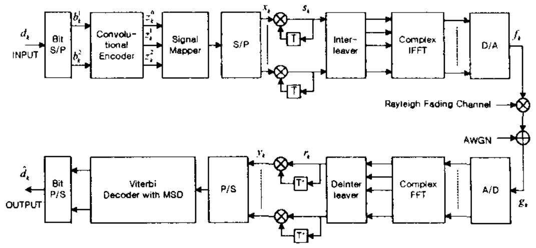

Fig. 1. TC π/8 shift 8PSK-OFDM Transmission System

By using signal set expansion and signal mapping by set partitions, significant coding gain can be achieved. To apply the ideas of π/4 shift QPSK to TCM, the $n/(n+1)$ convolutional encoder is used to expand the signal set from 2^n to $2^{n+1}(n=2)$, and to maximize the Euclidean distance between the signals. The signal sets of the π/4 shift QPSK are expanded from 4 phase difference states (±π/4 and ±3π/4) to 8 phase difference states (±π/8, ±3π/8, ±5π/8 and ±7π/8). For a brief explanation, the constellation diagrams of the incoherent π/8 shift 8PSK are shown in Fig. 2. The convolutional encoder and trellis diagrams are the same as those of [4], [5], [6].

A TCM is used as an inner code in order to correct the error events that consist of few bit errors. If a burst of symbol errors occurs at the channel output, the TCM can not be decoded correctly. Therefore, symbol interleaving is used in order to spread the bit errors over the entire OFDM symbol.

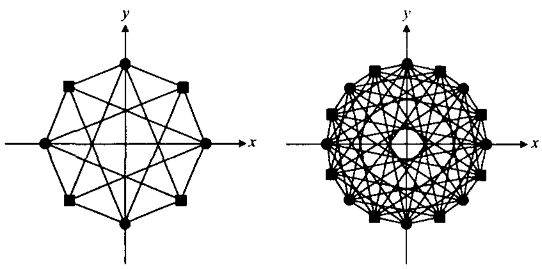

Fig. 2. The constellation of the π/4 shift QPSK and π/8 shift 8PSK

As the next step in the transmitter, an inverse fast Fourier transform (IFFT) is performed in order to calculate the discrete time OFDM signal. Furthermore, a guard interval is added to the signal as a periodic extension of the IFFT output sequence. The transmitted signal is distorted by the frequency-selective radio channel and, furthermore, AWGN is added.

III. Trellis-Coded π/8 shift 8PSK-OFDM with Sliding Multiple Symbol Detection

In the π/4 shift QPSK or the TC π/8 shift 8PSK, information is transmitted by the phase difference between the current signal and the previous signal. At the receiver, only the phase difference between the two sampling instants is needed for detecting information. But as shown in [7], the L-th order phase difference (phase difference between the current signal and the L-th previous signal) is capable of nonredundant error correction by operating as a syndrome, or as in [8], [9] one can perform multiple symbol detection to improve the performance of MDPSK. In this paper, we will show that the Viterbi decoder can improve the BER performance by using the branch metric with first and L-th order phase differences. The Viterbi decoder is equivalent to the sliding multiple symbol detection by means of using the branch metric with the first up to the L-th order phase differences. Also, the proposed algorithm can be used in the TC MDPSK as well as the TC π/8 shift 8PSK-OFDM.

Consider a transmitted signal sequence set of the TC π/8 shift 8PSK-OFDM with multicarrier number MC

$$S=(S^0, S^1\cdots, S^i, \cdots S^{MC-1}) \quad (1)$$

Each S_i is a transmitted signal sequence set of the ith subcarrier with length N and represented by

$$S^i=(S_0^i, S_1^i\cdots, S_k^i, \cdots S_{N-1}^i) \quad (2)$$

In each sequence S_k^i is a symbol of the TC π/8 shift 8PSK-OFDM to be transmitted in each subcarrier. It has the following complex form

$$S_k^i=\exp(j\theta_k^i) \quad (3)$$

where θ_k denotes the transmitted signal phase and takes one of the 16 values from the set $\{2\pi m/16+\pi/8, m=0,1,\cdots,15\}$. Then in the case of the TC π/8 shift 8PSK-OFDM

$$\begin{aligned} S_k^i &= \exp(j\theta_k^i) \\ &= \exp(j(\theta_{k-L}^i+\Delta\theta_{k-L+1}^i \\ &\qquad +\Delta\theta_{k-L+2}^i+\cdots+\Delta\theta_k^i)) \\ &\equiv S_{k-L}^i\cdot x_{k-L+1}^i\cdot x_{k-L+2}^i\cdots x_k^i \end{aligned} \quad (4)$$

The differential encoded TC π/8 shift 8PSK-OFDM signals are transmitted over a Rayleigh fading (frequency selective radio) channels with AWGN. The received TC π/8 shift 8PSK-OFDM symbol sequences for each subcarrier can be written in the form of $R^i=(R_0^i, R_1^i, \cdots, R_{N-1}^i)$. Here, R_k^i is the k-th received symbol in each subcarrier i,

$$R_k^i=\alpha_k^i\cdot S_k^i+N_k^i$$
$$k=0,\cdots N-1 \text{ and } i=0,\cdots MC-1 \quad (5)$$

where α_k^i and N_k^i are complex Gaussian random variables representing the channel fading and additive Gaussian noise of the considered subcarrier respectively. As the result of channel estimation, the receiver is assumed to know the channel transfer factor α_k^i exactly. The N_k^i's are a set of statistically independent and identically distributed variables with variance

$$\sigma^2=\frac{1}{2}E\left(N_k^i\cdot N_k^{i*}\right)=N_0 \quad (6)$$

N_0 is the one-sided power spectral density of the AWGN.

In order to improve the BER performance, we will apply the SMSD technique to trellis decoding by extracting total L phase differences

$y^i_{1,k}, \cdots, y^i_{L,k}$, where L has to be smaller than the variation of the channel transfer factor. The l-th order phase difference of the received signals, $y^{i,a}_{l,k}$ can be derived as

$$\begin{aligned} y^{i,a}_{l,k} &= R^i_k \cdot R^{i\,*}_{k-l} \\ &\approx |\alpha^i_k|^2 \cdot x^i_k \cdot x^i_{k-1} \cdots x^i_{k-l+1} \\ &\quad + \alpha^i_k \cdot S^i_k \cdot N^{i\,*}_{k-l} + \alpha^{i\,*}_k \cdot S^{i\,*}_{k-l} \cdot N^i_k \end{aligned} \tag{7}$$

Let $y^i_{l,k}$ be

$$\begin{aligned} y^i_{l,k} &\equiv \frac{y^{i,a}_{l,k}}{|\alpha^i_k|^2} = x^i_k \cdot x^i_{k-1} \cdots x^i_{k-l+1} \\ &\quad + \frac{1}{\alpha^{i\,*}_k \cdot S^{i\,*}_k} \cdot N^{i\,*}_{k-l} + \frac{1}{\alpha^i_k \cdot S^i_{k-l}} \cdot N^i_k \\ &= x^i_k \cdot x^i_{k-1} \cdots x^i_{k-l+1} + N^i_{l,k} \end{aligned} \tag{8}$$

and the pdf $y^i_{l,k}$ of can be expressed as

$$\begin{aligned} &P_Y(y^i_{l,k} \mid x^i_k, x^i_{k-1}, \cdots, x^i_{k-l+1}) \\ &\approx \frac{1}{\sigma_{l,k}\sqrt{2\pi}} \exp\left\{ \frac{-(y^i_{l,k} - x^i_k \cdot x^i_{k-1} \cdots x^i_{k-l+1})^2}{2\sigma^2_{l,k}} \right\} \end{aligned} \tag{9}$$

with expected value

$$E[y^i_{l,k}] = x^i_k \cdot x^i_{k-1} \cdots x^i_{k-l+1} \tag{10}$$

and variance

$$\begin{aligned} \sigma^2_{l,k} &= \left(\frac{1}{|\alpha^i_k \cdot S^i_k|^2} + \frac{1}{|\alpha^i_k \cdot S^i_{k-l}|^2} \right) \sigma^2 \\ &= \frac{2\sigma^2}{|\alpha^i_k|^2} \end{aligned} \tag{11}$$

The task of the demodulator is to process the extracted phase difference signal sequence $\boldsymbol{Y}$ set in order to produce an estimated value of $\hat{\boldsymbol{X}} = (\widehat{x^i_1}, \widehat{x^i_2}, \cdots, \widehat{x^i_{N-1}})$. Assume that all the information symbols are equally probable. In order to minimize the probability of symbol error, we process $\boldsymbol{Y}$ according to the maximum-likelihood decision. This procedure compares the conditional probabilities of the received signal given each possible transmitted sequence and chooses the transmitted sequence corresponding to the largest probability among them. That is, the decoder chooses $\hat{\boldsymbol{X}}$ if

$$P_Y(\boldsymbol{Y} \mid \hat{\boldsymbol{X}}) = \max_{all\ \boldsymbol{X}} P_Y(\boldsymbol{Y} \mid \boldsymbol{X}) \tag{12}$$

If we use L continuous phase differences, then

$$P_Y(Y|\hat{X}) = \max_{all\ X} \prod_{k=L}^{N-1} P(y^i_{1,k}, \cdots y^i_{L,k} \mid x^i_k \cdots x_{k-L+1}) \tag{13}$$

Generally, it is computationally easier to use the logarithm of the likelihood-function since this permits the summation, rather than the multiplication of terms. Then the log likelihood function [11] is expressed as :

$$\ln P_Y(Y|\hat{X}) = \max_{all\ X} \ln \prod_{k=L}^{N-1} f(y^i_{1,k}, \cdots y^i_{L,k} \mid x^i_k \cdots x_{k-L+1}) \tag{14}$$

We assume that the noise random variables are jointly Gaussian distributed and the channel is memoryless. They are also mutually independent. Therefore the conditional pdf of the channel differential signal becomes

$$\begin{aligned} \ln P_Y(Y|\hat{X}) &= \max_{all\ X} \ln \prod_{k=L}^{N-1} \frac{\left| [C_{Y^i_k Y^i_k}]^{-1} \right|^{1/2}}{(2\pi)^{L/2}} \\ &\times \exp\left\{ -\frac{[\boldsymbol{Y}^i_k - \overline{\boldsymbol{Y}^i_k}]^H [C_{Y^i_k Y^i_k}]^{-1} [\boldsymbol{Y}^i_k - \overline{\boldsymbol{Y}^i_k}]}{2} \right\} \end{aligned} \tag{15}$$

where $[C_{Y^i_k Y^i_k}]$ is the covariance matrix of a block of extracted phase difference :

$$\begin{aligned} [\boldsymbol{Y}^i_k - \overline{\boldsymbol{Y}^i_k}] &= \begin{bmatrix} y^i_{1,k} - \overline{y^i_{1,k}} \\ \vdots \\ y^i_{L,k} - \overline{y^i_{L,k}} \end{bmatrix} \\ &= [\boldsymbol{Y}^i_k - \boldsymbol{X}^i_k] = \begin{bmatrix} y^i_{1,k} - x^i_k \\ \vdots \\ y^i_{L,k} - x^i_k \cdots x^i_{k-L+1} \end{bmatrix} \end{aligned} \tag{16}$$

Then

$$\ln P_Y(Y|\hat{X}) = \max_{all\ X} \left[K - \sum_{k=L}^{N-1} A [\boldsymbol{Y}^i_k - \overline{\boldsymbol{Y}^i_k}]^H \cdot [C_{Y^i_k Y^i_k}]^{-1} [\boldsymbol{Y}^i_k - \overline{\boldsymbol{Y}^i_k}] \right] \tag{17}$$

where A and K are constants that can be disregarded in the maximization. In conclusion, the maximum-likelihood detector based on a sequence of signal $\boldsymbol{Y}^i_k$ seeks $\boldsymbol{X}$ which minimizes the following path metric

$$\lambda_p = \min_{all\ X} \left[\sum_{k=L}^{N-1} [\boldsymbol{Y}^i_k - \overline{\boldsymbol{Y}^i_k}]^H \cdot [C_{Y^i_k Y^i_k}]^{-1} [\boldsymbol{Y}^i_k - \overline{\boldsymbol{Y}^i_k}] \right] \tag{18}$$

We define the branch metric of the Viterbi decoder of the TC π/8 shift 8PSK-OFDM demodulation as the following:

$$\lambda_b = \left[Y_k^i - \overline{Y_k^i} \right]^H \left[C_{Y_k^i Y_k^i} \right]^{-1} \left[Y_k^i - \overline{Y_k^i} \right] \tag{19}$$

If we apply only the first phase difference i.e., $L=1$, the log likelihood function for the suboptimal MLSE can be approximated by a Gaussian distribution

$$\begin{aligned} P_Y\ (Y|\hat{X}) &= \max_{all\ X} \ln \prod_{k=1}^{N-1} P(y_{1,k}^i | x_k^i) \\ &= \max_{all\ X} \sum_{k=1}^{N-1} \ln \frac{1}{\sigma_{1,k}\sqrt{2\pi}} \cdot \exp\left\{ \frac{-(y_{1,k}^i - x_k^i)^2}{2\sigma_{1,k}^2} \right\} \\ &\approx \min_{all\ X} \sum_{k=1}^{N-1} (y_{1,k}^i - x_k^i)^2 \end{aligned} \tag{20}$$

Therefore the branch metric λ_b of the Viterbi decoder becomes

$$\lambda_b = | y_{1,k}^i - x_k^i |^2 \tag{21}$$

This is same as the conventional demodulator of the TCM. Also, if we apply two continuous phase differences i.e., $L=2$, the log likelihood function is expressed as

$$\begin{aligned} P_Y(Y|\hat{X}) &= \max_{all\ X} \ln \prod_{k=2}^{N-1} P(y_{1,k}^i, y_{2,k}^i | x_k^i, x_{k-1}^i) \\ &= \max_{all\ X} \sum_{k=2}^{N-1} \ln \frac{1}{2\pi\sigma_k^2\sqrt{1-\rho^2}} \cdot \\ &\quad \exp\left[\frac{-1}{2(1-\rho^2)\sigma_k^2} \left\{ \left| y_{1,k}^i - \overline{y_{1,k}^i} \right|^2 \right.\right. \\ &\quad - 2\rho \left| y_{1,k}^i - \overline{y_{1,k}^i} \right| \cdot \left| y_{2,k}^i - \overline{y_{2,k}^i} \right| \\ &\quad \left.\left. + \left| y_{2,k}^i - \overline{y_{2,k}^i} \right|^2 \right\} \right] \end{aligned} \tag{22}$$

where

$$\overline{y_{1,k}^i} = x_k^i \quad \overline{y_{2,k}^i} = x_k^i \cdot x_{k-1}^i \tag{23}$$

$$\begin{aligned} \sigma_k^2 &= E\left[\left| y_{1,k}^i - \overline{y_{1,k}^i} \right|^2 \right] = E\left[N_{1,k}^{i\ 2} \right] \\ &= E\left[\left| y_{2,k}^i - \overline{y_{2,k}^i} \right|^2 \right] = E\left[N_{2,k}^{i\ 2} \right] \end{aligned} \tag{24}$$

$$\rho = \frac{E\left[\left| y_{1,k}^i - \overline{y_{1,k}^i} \right| \cdot \left| y_{2,k}^i - \overline{y_{2,k}^i} \right| \right]}{\sigma_k^2} \tag{25}$$

And then, the path metric and the branch metric of the Viterbi decoder can be approximated by

$$\begin{aligned} \lambda_p \approx \min_{all\ X} \sum_{k=2}^{N-1} & \left| y_{1,k}^i - x_k^i \right|^2 \\ & - 2\rho \left| y_{1,k}^i - x_k^i \right| \cdot \left| y_{2,k}^i - x_k^i \cdot x_{k-1}^i \right| \\ & + \left| y_{2,k}^i - x_k^i \cdot x_{k-1}^i \right|^2 \end{aligned} \tag{26}$$

$$\begin{aligned} \lambda_b &= \left| y_{1,k}^i - x_k^i \right|^2 \\ &\quad - 2\rho \left| y_{1,k}^i - x_k^i \right| \cdot \left| y_{2,k}^i - x_k^i \cdot x_{k-1}^i \right| \\ &\quad + \left| y_{2,k}^i - x_k^i \cdot x_{k-1}^i \right|^2 \end{aligned} \tag{27}$$

In eq. (27) , the first term is the Euclidean distance between the phase difference of the received sequence signal and the candidate signal x_k^i. The third term is the Euclidean distance between the second order phase difference of the received sequence signal and the candidate signal $x_k^i \cdot x_{k-1}^i$. The second term is the multiplication of the square root of the first and third term, which can be ignored under high SNR assumption. Thus, the suboptimal branch metric becomes

$$\begin{aligned} \lambda_b &= \left| y_{1,k}^i - x_k^i \right|^2 + \left| y_{2,k}^i - x_k^i \cdot x_{k-1}^i \right| \\ &= \left| R_k^i \cdot R_{k-1}^{i\ *} - x_k^i \right|^2 \\ &\quad + \left| R_k^i \cdot R_{k-2}^{i\ *} - x_k^i \cdot x_{k-1}^i \right|^2 \end{aligned} \tag{28}$$

If we use L continuous phase differences, the suboptimal branch metric finally becomes

$$\begin{aligned} \lambda_b &= \left| y_{1,k}^i - x_k^i \right|^2 + \left| y_{2,k}^i - x_k^i \cdot x_{k-1}^i \right|^2 + \cdots \\ &\quad + \left| y_{L,k}^i - x_k^i \cdot x_{k-1}^i \cdots x_{k-L+1}^i \right|^2 \\ &= \left| R_k^i \cdot R_{k-1}^{i\ *} - x_k^i \right|^2 \\ &\quad + \left| R_k^i \cdot R_{k-2}^{i\ *} - x_k^i \cdot x_{k-1}^i \right|^2 + \cdots \\ &\quad + \left| R_k^i \cdot R_{k-L}^{i\ *} - x_k^i \cdot x_{k-1}^i \cdots x_{k-L+1}^i \right|^2 \end{aligned} \tag{29}$$

A demodulation block diagram of the TC π/8 shift 8PSK-OFDM with SMSD is shown in Fig. 3.

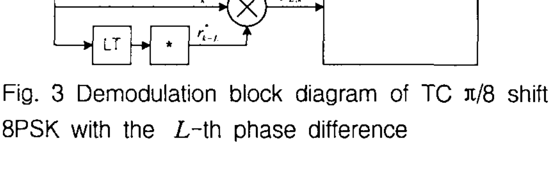

Fig. 3 Demodulation block diagram of TC π/8 shift 8PSK with the L-th phase difference

IV. Simulation Results

In this section, we present some computer simulation results for the uncoded π/4 shift QPSK OFDM and the TC π/8 shift 8PSK OFDM. The performance improvement of the TC π/8 shift 8PSK-OFDM is investigated based on the Monte Carlo method. OFDM systems with 1024 subcarriers are simulated respectively. Interleaving is performed by a block interleaver with 32×32 for each subcarrier number. A guard interval ($=MC/4$) is added to the discrete time signal as a periodic extension of the IFFT output sequence. It is assumed that the transmitted signal is corrupted by the AWGN, Rician or Rayleigh channel. In the simulations, the channel is frequency-selective but it is not time-varying, i.e., the Doppler frequency is $f_D=0$ [12]. At the receiver, the decoding is performed based on Viterbi algorithm with or without the second phase difference metric. The simulation calculates the Bit Error Rate (BER) of the uncoded π/4 shift QPSK-OFDM and the TC π/8 shift 8PSK-OFDM with or without the second order phase difference metric of 4, 8 and 16-state for different SNRs. Figs. 4-6 show the bit error probabilities of the TC π/8 shift 8PSK-OFDM system with the 1024 subcarriers number and 4, 8, 16-state of the trellis. At the 10^{-5} BER, numerical results corresponding to 16-state are summarized in Table I.

Table 1. SNR of the 16 State TC π/8 shift 8PSK-OFDM with or without SMSD and OFDM for 10^{-5} BER

	TC OFDM with SMSD	TC OFDM	OFDM
AWGN	7.7 dB	9.0 dB	12.0 dB
Rician	8.6 dB	10.2 dB	14.2 dB
Rayleigh	19.4 dB	21.0 dB	Error floor

Fig. 4 show the bit error probabilities of the TC π/8 shift 8PSK-OFDM system with $MC=1024$ subcarriers, respectively versus the E_b/N_0 for the AWGN channel. When $MC=1024$ system, the TC π/8 shift 8PSK-OFDM demonstrates an improvement of 2.5-3.0 dB over the uncoded π/4 shift QPSK-OFDM. The TC π/8 shift 8PSK-OFDM with the second order phase difference metric shows performance improvement up to 3.5-4.3 dB over the uncoded π/4 shift QPSK-OFDM.

Fig. 5 show the BER for the Rician channel. For $MC=1024$, the TC π/8 shift 8PSK-OFDM demonstrates an improvement of 3.8-4.0 dB over the uncoded π/4 shift QPSK-OFDM. The TC π/8 shift 8PSK-OFDM with the second order phase difference metric improves the performance up to 4.4-5.6 dB over the uncoded π/4 shift QPSK-OFDM.

The uncoded OFDM and 4 or 8-state TC π/8 shift 8PSK-OFDM with or without SMSD yield the error floor for the Rayleigh channel. The simulation results in Figs. 6 indicate that guard intervals dont eliminate the ISI completely for the 4-state and 8-state TC π/8 shift 8PSK-OFDM.

But for the 16-state TC π/8 shift 8PSK-OFDM, the error floor doesnt occur and the performance increases to 1.6 dB with the SMSD. So the multiple symbol detection can correct the error of the error floor for the Rayleigh channel.

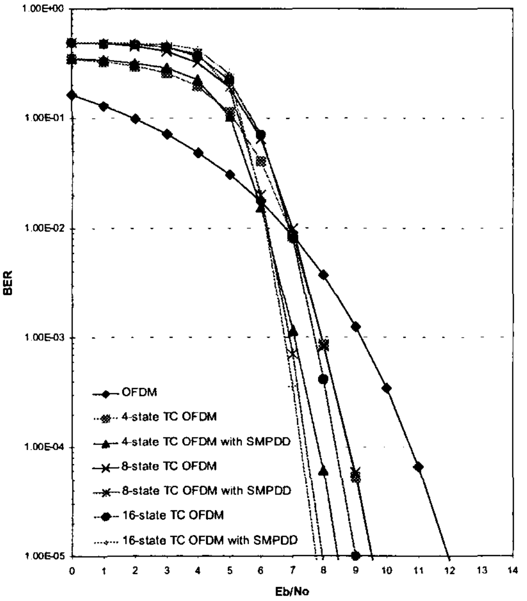

Fig. 4 Performance of the OFDM and the TC π/8 shift 8PSK-OFDM with SMSD for AWGN channel(subcarrier number=1024).

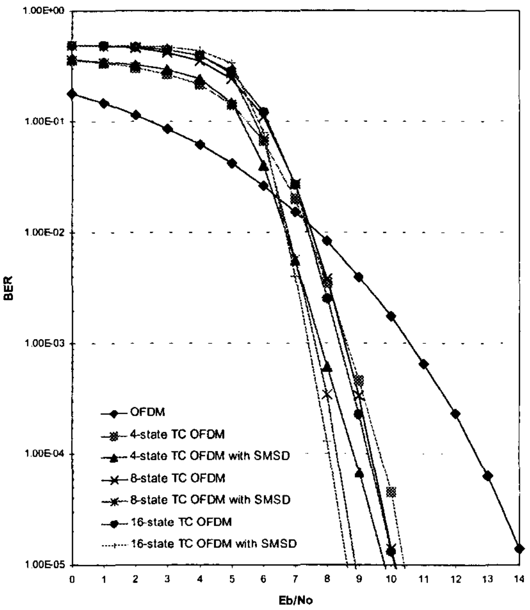

Fig. 5 Performance of the OFDM and the TC π/8 shift 8PSK-OFDM with SMSD for Rician channel(subcarrier number=1024).

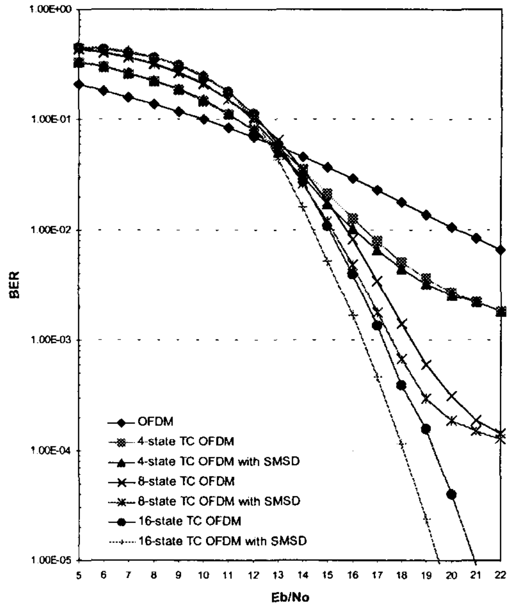

Fig. 6 Performance of the OFDM and the TC π/8 shift 8PSK-OFDM with SMSD for Rayleigh channel(subcarrier number=1024).

In the TC π/8 shift 8PSK-OFDM, neither the ideal channel state information nor pilot symbols are required. We extract the phase difference of the successive channel symbols for information. Compared with multiple symbol detection method [9], our method utilizes a sliding window instead of block-by-block processing to perform continuous multiple symbol detection.

V. Conclusions

We have presented the TC π/8 shift 8PSK-OFDM with sliding multiple symbol detection. At the receiver, we design the differential demodulator extracting multiple symbol phase difference. In addition, the Viterbi decoder containing new branch metrics of the squared Euclidean distance of the L phase differences is introduced in order to improve the bit error rate (BER) in the differential detection of the trellis-coded π/8 shift 8PSK-OFDM. The proposed Viterbi decoder is based on a sliding multiple symbol detection method that uses the branch metric with the first up to the L-th order phase differences. Also, we describe the Viterbi algorithm in order to use such branch metrics.

Through this method, a performance improvement of 3.3-4.3 dB at a 10^{-5} BER is obtained over the uncoded π/4 shift QPSK-OFDM for AWGN. We also reveal the fact that the BER performance for Rician channel can be improved up to 4.4-5.7 dB. For Rayleigh channel, the uncoded OFDM and 4 or 8-state TC π/8 shift 8PSK-OFDM with or without SMSD yield the error floor. But for the 16-state TC π/8 shift 8PSK-OFDM, error floor doesnt occur and the performance increases to 1.6 dB with the SMSD. These improvements are achieved without sacrificing any bandwidth efficiency (data rate) or power efficiency.

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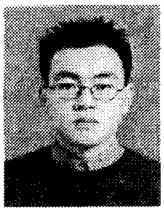
저 자 소 개

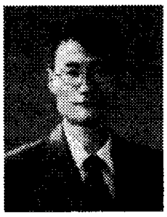


안필승(Phil-Seung Ahn)

1994년~2001년 관동대학교 전자공학과 공학사

2001년~현재 관동대학교 전자공학과 공학석사

※관심분야 : 이동통신, TCM, OFDM 등



김종일(Chong-Il Kim)

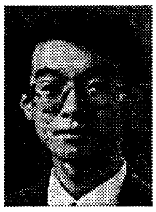
1987년 서강대학교 전자공학과 공학사

1989년 연세대학교 전자공학과 공학석사

1993년 연세대학교 전자공학과 공학박사

1993년~현재 관동대학교 정보기술공학부(전자정보전공) 교수.

※관심분야 : 이동통신, 부호이론, CDMA, TCM, OFDM 등



김한종(Han-Jong Kim)

1986년 한양대학교 전자공학과 공학사

1988년 연세대학교 전자공학과 공학석사

1994년 연세대학교 전자공학과 공학박사

1994년~현재 한국기술교육대학교 정보기술공학부(정보통신전공) 부교수.

※관심분야 : 이동통신, 방송통신, TCM, OFDM, Space-Time Codes 등

Zhengyuan Xu

Zhengyuan Xu received both the B.E. and M.E. degrees in electronic engineering from Tsinghua University, Beijing, China, in 1989 and 1991 respectively, and the Ph.D. degree in electrical engineering from Stevens Institute of Technology, Hoboken, NJ, in 1999. From 1991 to 1996, he worked as an engineer and department manager at the Tsinghua Unisplendour Group Corp. of Tsinghua University. From 1996 to 1999, he was a research assistant and research associate at Stevens Institute of Technology. In 1999, he joined the Department of Electrical Engineering, University of California, Riverside, as an assistant professor. His current research interests include advanced signal processing, multirate communication, multiuser detection and system identification. Dr. Xu received the Outstanding Student Award and the Motorola Scholarship from Tsinghua University, and the Peskin Award from Stevens Institute of Technology. He also received the Academic Senate Research Award and the Regents' Faculty Award from University of California, Riverside.