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페이딩 채널환경에서 CDFDM 시스템에 대한 채널 추정과 결합된 심볼검출 방법

(Symbol Decoding Schemes Combined with Channel Estimations for Coded OFDM Systems in Fading Channels.)

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요 약

본 논문에서는 페이딩 채널환경에서 COFDM 시스템에 대한 채널 추정과 결합된 심볼 검출 방법에 대해 제안하였다. 제안된 방법은 1) 심볼의 길이에 채널 부호화기의 코드워드 길이를 배합하는 기법과, 2) 세가지 채널 추정 기법과의 조합으로 나타내었다. 첫째 방법은 훈련 신호를 이용한 채널추정 기법과 제안한 심볼 검출 기법을 결합시킨 것이며, 둘째 방법은 첫째 방법에 결정지향 채널추정(Decision-Directed Channel Estimation) 기법을 결합시킨 방법이다. 마지막으로, 근접 부채널간의 AWGN(Additive White Gaussian Noise)의 영향을 줄이기 위해 디인터리브된 평균 채널추정(Averaged Channel Estimation) 기법을 둘째 방법에 결합시켰다. 제안한 3가지 방법들은 영 강압 등화(Zero Forcing Equalizing)방법과 비교하여 커다란 성능 개선 효과가 있음을 컴퓨터 시뮬레이션을 통하여 검증하였다.

Abstract

This paper proposes symbol decoding schemes combined with channel estimation techniques for coded orthogonal frequency division multiplexing (COFDM) systems in fading channels. sThe proposed symbol decoding schemes are consisted of a symbol decoding technique and channel estimation techniques. The symbol decoding based on Viterbi algorithm is achieved by matching the length of branch word from encoder trellis to the codeword length of symbol candidate on decoder trellis. Three combination schemes are described and their error performances are compared. The first scheme is to combine a symbol decoding technique with a training channel estimation technique. The second scheme joins a decision directed channel estimation technique to the first scheme. The time varying channel transfer functions are tracked by the decision directed channel estimation technique and the channel transfer functions used in the symbol decoder are updated every COFDM symbol. Finally, In order to reduce the effect of additive white Gaussian noise (AWGN) between adjacent subchannels, deinterleaved average channel estimation technique is combined. The error performances of the three schemes are significantly improved being compared with that of zero forcing equalizing schemes.

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Ⅰ. Introduction

According to increasing demand for multimedia wireless communication service, Orthogonal Frequency Division Multiplexing (OFDM) technique has been applied to various types of communication systems[1][2][3]. OFDM technique is known as combating the inter-symbol interference (ISI) that occurs in multipath propagation[4]. In OFDM system, generally the use of pilot symbols has been studied to estimate channels[5][6][7]. Decision-directed channel estimation (DDCE) scheme for OFDM system was studied to track the variation of channel transfer function in a Rayleigh-fading environment[8]. And an averaged decision-directed channel estimation (ADDCE) scheme was investigated to be obtained a improved channel estimation for wireless ATM system based on OFDM technique[9]. But OFDM system without channel coding is virtually unusable on multipath fading channels with deep notches occurring in the signal spectrum[1]. Therefore, OFDM technique requires channel coding to protect the system from transmission errors. Coded Orthogonal Frequency Division Multiplexing (COFDM) systems with channel coding have been studied to improve error performance[1][4][10][11]. And several approaches that decode the equalized symbols into the data sequence by maximum likelihood sequence detection (MLSD) have been investigated for COFDM systems[1][12][13]. But in[1][12][13], MLSE using Viterbi algorithm can be used only when the codeword length of channel encoder and the codeword length of symbol (modulated codeword) do correspond.

In this paper, we propose symbol decoding technique that is based on Viterbi algorithm and utilizes the channel transfer functions estimated by channel estimation techniques. We call it Symbol Viterbi Decoding (SVD) technique. Although the codeword length of channel encoder differs from the codeword length of symbol, this novel SVD technique is able to use Viterbi algorithm. In order to apply Viterbi algorithm to SVD decoder, we present how to match the branch words corresponding to the codewords of channel encoder into symbol branches on decoder trellis.

And we derive three combination schemes that are combined SVD technique with three channel estimation techniques. The channel estimating techniques for the proposed three combination schemes are Training Channel Estimation (TCE), DDCE, and Deinterleaved Average Channel Estimation (DACE). TCE technique uses training symbols periodically transmitted from transmitter. DDCE technique makes the estimated channel transfer functions be tracked a variation of channel on time. And DACE technique reduces the effect of AWGN between adjacent subchannels. We show that the combination schemes provide significant improvement in the error performance. The error performances of the schemes are compared with those of zero forcing equalizing schemes.

This paper is organized as follows. Section 2 describes the basic principle of proposed COFDM system. Section 3 derives SVD technique and channel estimation techniques, and presents three symbol decoding schemes combined with the channel estimation techniques. And section 4 shows computer simulation results to demonstrate the effectiveness of proposed schemes.

Ⅱ. System Model

In this paper, zero frequency offset and perfect symbol timing is assumed. Inter-channel interference (ICI) is ignored because it is assumed that the channel during one COFDM symbol is constant.

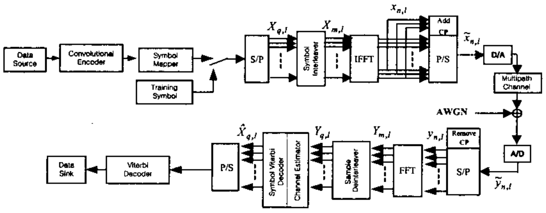


그림 1. 제안한 COFDM 시스템의 블록도
Fig. 1. A baseband block diagram for proposed COFDM system.

Fig. 1 shows the baseband block diagram of the proposed COFDM system. Training data are transmitted to estimate training channel transfer functions periodically. Information data are coded by convolutional encoder and modulated by a symbol mapper. After data are modulated to symbols, the symbols are scattered by a symbol interleaver in frequency domain. In this system, the interleaver and the deinterleaver have parallel inputs and parallel outputs, but their rearranging functions are the same functions as block interleaver or block deinterleaver. The deinterleaver rearranges samples of signal that are sampled by analog-to-digital (A/D) converter while the interleaver deals with symbols demodulated by symbol mapper. The COFDM symbols are generated by an inverse fast Fourier transfer (IFFT) and the *l-th* sample of the COFDM symbol can be expressed as follows

$$x_{n,l}=\sum_{m=0}^{N-1}X_{m,l}e^{j2\pi nm/N},\quad n=0,1,2,\cdots,N-1 \tag{1}$$

where $X_{m,l}$ represents a symbol transmitted on *m-th* subchannel of *l-th* COFDM symbol and N is the number of subchannels. To avoid ISI, caused by delay spread of multipath, cyclic prefix is inserted into $x_{n,l}$ as follows

$$\tilde{x}_{n,l}=\sum_{m=0}^{N-1}X_{m,l}e^{j2\pi nm/N},\quad n=-N_G,-N_G+1,\cdots,-1,0,1,\cdots,N-1 \tag{2}$$

where N_G is the number of samples for guard time. The inserting cyclic prefix simulates a channel performing cyclic convolution, which implies that orthogonality over dispersive channels can be kept when the cyclic prefix is longer than the impulse response of the channel[14].

After being converted by A/D converter at receiver, a received COFDM symbol is as follows

$$\tilde{y}_{n,l}=\sum_{m=0}^{N-1}X_{m,l}H_{m,l}e^{j2\pi nm/N}+w_{n,l},$$
$$n=-N_G,-N_G+1,\cdots,-1,0,1\cdots,N-1 \tag{3}$$

where $H_{m,l}$ represents a channel transfer function on m-th subchannel of *l-th* COFDM symbol. And $w_{n,l}$ is additive white Gaussian noise (AWGN). After being removed the cyclic prefix, the received COFDM symbol is performed on FFT. The samples, $Y_{q,l}$, rearranged by deinterleaver are decoded to symbols by SVD decoder, and the decoded symbols are decoded by Viterbi decoder.

We design the inter-subchannel space of the system is chosen large to void the ICI, compared to the maximal Doppler frequency.

Ⅲ. Symbol Decoding and Channel Estimations

In this section, we derive SVD technique and three combination schemes. As Fig.2, the proposed combination schemes are combined the symbol decoding technique SVD using Viterbi algorithm with the channel estimation techniques, i.e., TCE, DDCE or DACE. These schemes are associated with a parallel sample deinterleaver. And those all are

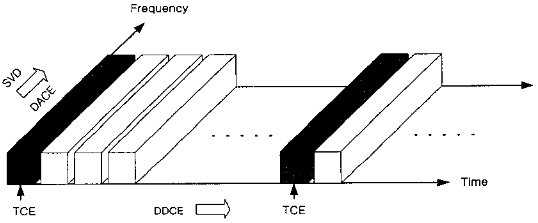


그림 2. 심볼검출 및 채널추정
Fig. 2. Symbol decoding and channel estimations.

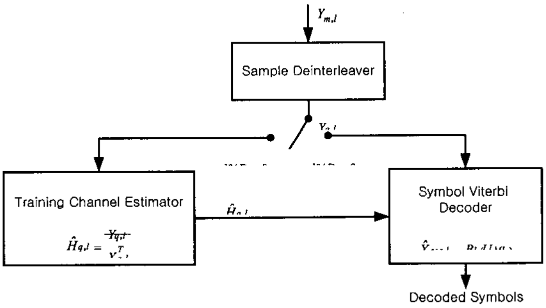


그림 3. 방안1 : SVD와 TCE가 결합된 방안의 블록도
Fig. 3. Scheme 1 : Functional block diagram of the scheme combined SVD with TCE.

performed in frequency domain, though TCE and DDCE techniques are executed on time axis while SVD and DACE technique are executed on frequency axis.

1. Scheme 1: Combining Symbol Decoding scheme with Training Channel Estimation technique.

After being deinterleaved, the q-th sample of the l-th received COFDM symbol is

$$Y_{q,l} = X_{q,l}H_{q,l} + W_{q,l},\quad q = 0, 1, 2, \cdots, N-1 \tag{4}$$

where q is subchannel number rearranged by the deinterleaver and N is the number of subchannels. And $X_{q,l}$, and $H_{q,l}$ represent sample of transmitted signal, channel trnsfer function and AWGN, corresponding to q-th rearranged subchannel of l-th COFDM symbol on frequency domain respectively. Therefore, the estimated channel transfer function is

$$\hat{H}_{q,l} = \frac{Y_{q,l}}{X_{q,l}} = H_{q,l} + \frac{W_{q,l}}{X_{q,l}} = H_{q,l} + W'_{q,l} \tag{5}$$

where $\hat{H}_{q,l}$ denotes the estimated value of $H_{q,l}$.

Let D be the repeating period of training symbol. Then, each training channel transfer function estimated by training symbol $X^T_{q,l}$ is descibed as follows

$$\hat{H}^T_{q,i} = \frac{Y_{q,l}}{X^T_{q,i}},\quad i = l \quad if\ (l\%D) = 0 \tag{6}$$

Here, % denotes modulo operator.

The training channel transfer functions are used by the SVD decoder to decode symbols and are updated periodically.

For example, if the polynomials of convolutional encoder are

$$\begin{aligned} g_0^{(0)}(D) &= 1 + D^2 \\ g_0^{(1)}(D) &= 1 + D + D^2 \end{aligned} \tag{7}$$

and 16-QAM modulator is used in the proposed system, we can derive the decoder trellis diagram in Fig. 4 and symbol candidates in Table 1.

Fig. 4 shows the trellis diagram for symbol decoding scheme that uses Viterbi algorithm associated with the convolutional encoder as Eq. (7). The outputs of the convolutional encoder are modulated to 16-QAM symbols. But SVD decoder is not able to use directly Viterbi algorithm to decode symbols at receiver because the codeword length of convolutional encoder, i.e., the length of branch word on decoder trellis differs from the codeword length of symbol. Therefore, branch words should be matched to symbol candidates on decoder trellis. As shown in Fig. 4, one symbol branch consists of 2 trellis branches and each state every trellis symbol step

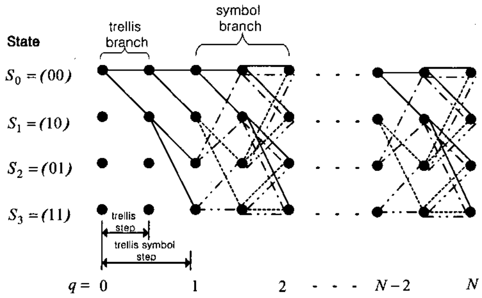


그림 4. 예시된 시스템의 심볼검출을 위한 디코더 트렐리스 블록도

Fig. 4. Decoder trellis diagram to decode symbols for the example system.

표 1. 예시된 시스템의 각 트렐리스 상태상에서의 심볼들

Table 1. Symbol candidates on each trellis state for the example system.

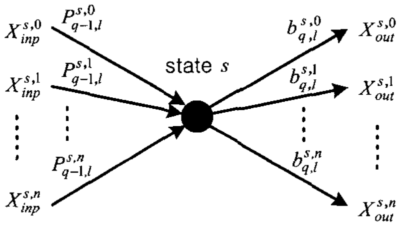
Trellis state		S_0		S_1		S_2		S_3	
Symbol candidate	t	Input ($X_{inp}^{0,t}$)	Output ($X_{out}^{0,t}$)	Input ($X_{inp}^{1,t}$)	Output ($X_{out}^{1,t}$)	Input ($X_{inp}^{2,t}$)	Output ($X_{out}^{2,t}$)	Input ($X_{inp}^{3,t}$)	Output ($X_{out}^{3,t}$)
	0123	0000	0000	0011	0111	1101	0001	1110	1011
		0111	0011	0100	0100	1010	0010	1001	1000
		1100	1101	1111	1010	0001	1100	0010	0110
		1011	1110	1000	1001	0110	1111	0101	0101

has 4 input symbol candidates and 4 output symbol candidates. The transmitted symbols can be one of sequences that are concatenated with one of 4 symbol candidates inputted to each state every trellis symbol step on decoder trellis. The 16-QAM symbol candidates on each state are presented with binary in Table 1. The number of trellis step is 2 by N but the number of trellis symbol step is *N*.

We can drive that a decoder trellis diagram of the proposed system associated with $1/k$ rate convolutional encoder and $Q-ary$ modulator has the property as follows

$$\log_k \sqrt{Q} = m \tag{8}$$

where k is the length of branch word on decoder trellis and m is the number of trellis branch corresponding to one symbol branch.



s : stste number
n : the number of input or output branches
$(n = 2^m)$

그림 5. 디코더 트렐리스 상에서의 입출력 심볼들에 해당하는 심볼 브랜치 및 패스
Fig. 5. Symbol branches and paths corresponding to input and output symbol candidates on *s-th* state in decoder trellis.

We can obtain the maximum likelihood symbol sequence by using Viterbi algorithm on the decoder trellis diagram. The number of the survival paths on each state is only one with the least value. This survival path value is added to the values of output symbol branches on the state in order to obtain next path values at next trellis symbol step. In Fig. 5, the path on *t-th* output symbol branch at *q-th* trellis symbol step has a value that is added the *(q-1)-th* survival path value to t-th output symbol branch value as follows.

$$P_{q,l}^{s,t} = b_{q,l}^{s,t} + min\,\mathbf{P}_{q-1,l}^{s}$$
$$t = 0,1,\cdots,r-1 \quad for\ s = 0,1,\cdots,z-1 \tag{9}$$

where

$$\mathbf{P}_{-1,} = \begin{Bmatrix} P_{-1,}^{,0} \\ P_{-1,}^{,1} \\ \vdots \\ P_{-1,}^{,-1} \end{Bmatrix}$$
$$b_{,}^{,t} = \left| Y_{,} - \hat{H}_{,} X_{t}^{,t} \right|^2, \qquad t = 0,1,\cdots,r-1 \quad for\ s = 0,1,\cdots,z-1 \tag{10}$$

where $P_{q,l}^{s,l}$ is the path value corresponding to *t-th* output symbol branch on *s-th* state at *q-th* trellis symbol step and $b_{q,l}^{s,l}$ denotes the symbol branch value with *t-th* output symbol $X_{out}^{s,t}$ on *s-th* state at *q-th* trellis symbol step. $X_{out}^{s,t}$ is the *t-th* output symbol of the output symbol candidates on *s-th* state in decoder trellis, and $\hat{H}_{q,l}$ is the channel transfer function estimated by TCE with Eq. (6). And t is the output symbol number on *s-th* state, s is the state number at *q-th* trellis symbol step, z is the number of states and r denotes the number of input or output branches, i.e., $r = 2^m$. In Eq. (10), $|\bullet|$ denotes absolute value.

The input symbol candidate $X_{out}^{s,j}$ corresponding to the survival path $P_{q-1,l}^{s,j}$ on the state s should be memorized in order to decode symbol sequence. We can decode a symbol sequence $\hat{X}_{seq}$ by tracing back at *(N-1)-th* trellis symbol step as follows.

$$_{seq,l} = B_l(\ _{ML}) = B_l\left(\min \begin{Bmatrix} \min \mathbf{P}_{N-2,l}^{0} \\ \min \mathbf{P}_{N-2,l}^{1} \\ \vdots \\ \min \mathbf{P}_{N-2,l}^{z-1} \end{Bmatrix} \right) \tag{11}$$

where $B_l(\bullet)$ denotes the function that decodes symbol sequence with input symbol candidates on maximum likelihood survival path U_{ML}.

The SVD technique improves significantly the performance because it corrects error symbols by decoding ML symbol sequence with Viterbi algorithm.

2. Scheme 2: Combining Symbol decoding scheme with Training Channel Estimation technique and Decision Directed Channel Estimation technique.

In time varying channel environments, it is necessary to track adaptively a variation of channels at receiver. We use DDCE technique[9] as follows to track the variation of channel transfer function.

$$\hat{H}_{q,l} = \rho \frac{Y_{q,l}}{\hat{X}_{q,l}} + (1-\rho)\hat{H}_{q,l-1} \tag{12}$$

where ρ is the update factor for channel tracking and $\hat{X}_{q,l}$ denotes a symbol decoded by SVD decoder.

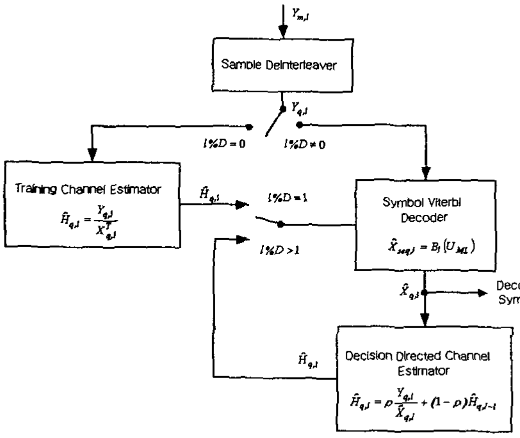


그림 6. 방안 2 : SVD, TCE 및 DDCE가 결합된 방안의 블록도

Fig. 6. Scheme 2 : Functional block diagram of the scheme combined SVD with TCE and DDCE.

Fig. 6 is shown a combined symbol decoding scheme with DDCE technique. This scheme can adaptively obtain accurate channel transfer functions from DDCE estimator because of using the symbols corrected by SVD decoder every OFDM symbols. And the obtained channel transfer function is feedback into SVD decoder to decode symbol sequence for next OFDM symbol. Therefore, this scheme can obtain excellent performance of the system, especially over higher signal-to-noise rate.

3. Scheme 3: Combining Symbol decoding scheme with Training Channel Estimation technique, Decision Directed Channel Estimation technique and Deinterleaved Average Channel Estimation technique.

In system with the interleaver /deinterleaver pair, two symbols transmitted at two adjacent subcarriers on transmission channels are separated by the number of columns at receiver. Thus receiver has need to deinterleave a receiving signal. After deinterleaving, two adjacent subcarriers on transmission channels are separated by the number of column of the deinterleaver. Therefore, in order to average the channel transfer functions for adjacent subcarriers after deinterleaving, we obtain the averaged channel transfer functions by rearranging again the deinterleaved channels as follows

$$\begin{aligned}
&\hat{H}_{q,l} = \alpha\hat{H}_{(q+N-1)-c,l} + (1-2\alpha)\hat{H}_{q,l} + \alpha\hat{H}_{q+c,l} && \mathit{for}\ \ 0<q<c \\
&\hat{H}_{q,l} = \alpha\hat{H}_{q-c,l} + (1-2\alpha)\hat{H}_{q,l} + \alpha\hat{H}_{q+c,l} && \mathit{for}\ \ c\le q<N-c \\
&\hat{H}_{q,l} = \alpha\hat{H}_{q-c,l} + (1-2\alpha)\hat{H}_{q,l} + \alpha\hat{H}_{(q-N+1)+c,l} && \mathit{for}\ \ N-c\le q<N-1
\end{aligned} \tag{13}$$

where N is the number of subcarriers and c is the number of columns of the interleaver/ deinterleaver. The weighting factor is determined by the following basic rules. First, the value of α needs to be decreased when the channel transfer function varies rapidly over adjacent subchannels. Second, the value needs to be increased as Eb/N0 is decreased.

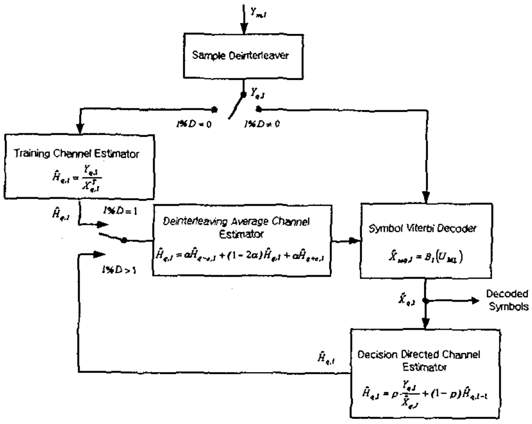


그림 7. 방안 3 : SVD, TCE, DDCE 및 DACE가 결합된 방안의 블록도

Fig. 7. Scheme 3: Functional block diagram of the scheme combined SVD with TCE, DDCE and DACE.

Fig. 7 shows a functional block diagram of the scheme combined SVD with TCE, DDCE and DACE for the proposed system. DACE estimator rearranges again the channel transfer functions by order transmitted on communication channels and averages them as Eq. (13). Then SVD decoder performs the symbol decoding with Eq. (11).

DACE estimator results accurate training channel transfer functions every repeating period of training symbol, or calculates more precisely the tracked time varying channel transfer functions every OFDM symbol period.

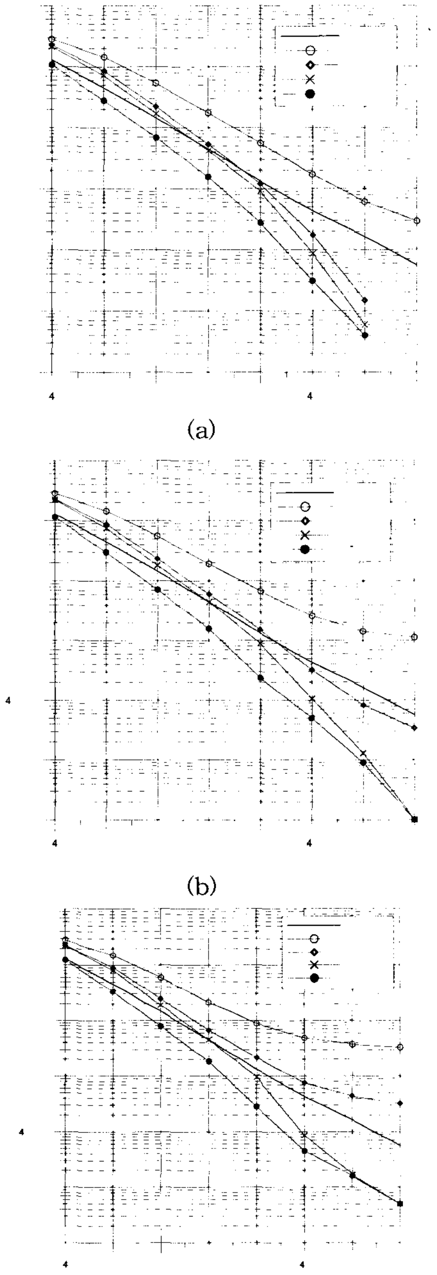
Ⅳ. Simulation Results

In the proposed system, 16-QAM modulation scheme is applied to COFDM system with 64 subchannels on fading environment.

COFDM symbol period including guard interval (*880 nsec*) becomes *4.4 μsec*. The effective system period is *3.52 μsec*. The subchannels are spaced at *0.284 MHz*, and the total bandwidth required by 64 subchannels is about *18.2 MHz*. Sampling period becomes *55 nsec*. One COFDM symbol in this system is composed of 64 data samples and a circular prefix of length 16. All simulations are performed with $\alpha=0.15$, $\rho=0.2$. And it is assumed that one OFDM training symbol is sent periodically and the *64-th* subchannels in every OFDM information symbols carry zero data at transmitter in order to decode data sequence by COFDM symbol unit at receiver. That is, two zero data for *127-th*, *128-th* source binary data are inserted to the convolutional encoder to clear the state memories of the encoder whenever OFDM symbol is started. The code rate ½ convolutional encoder and 8×8 symbol interleaver/deinterleaver are used, and the decoding depth of Viterbi decoder is designed with 32.

Jake's model with 2 paths of same power is considered as the channel model with maximum Doppler frequencies, i.e., from *52 Hz* to *520 Hz*.

Fig.8 shows performances for the proposed schemes and zero forcing schemes that the repeating period of training symbol has 20, 60 or 100 each. The zero forcing scheme is performed with the



(a)

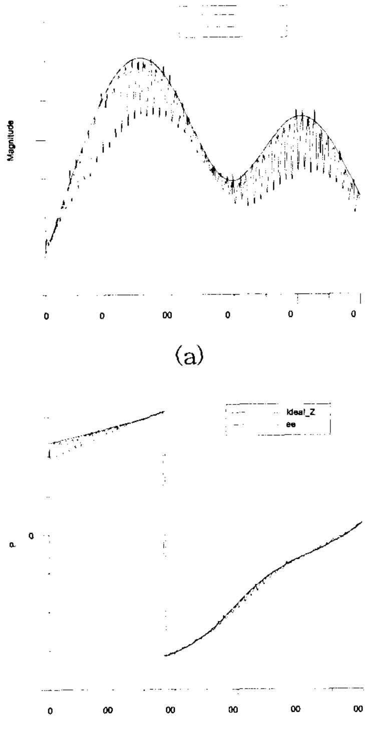
(b)

(c)

그림 8. 도플러주파수 *52 Hz*에서 훈련심볼 주기 *D* 를 갖는 시스템에서의 성능비교
(a) *D=20*. (b) D=60. (c) *D=100*.

Fig. 8. Performance comparisons with maximum Doppler frequency *52 Hz* and repeating period of training symbol *D*.
(a) *D=20*. (b) *D=60*. (c) *D=100*.

received sample, $Y_{q,l}$, and the channel transfer function estimated by TCE. And the ideal zero forcing is assumed that channel information is perfectly known. In fig.8, the performances of the proposed schemes are significantly better improved than that of the zero forcing schemes. Especially scheme 2 has better performance than the ideal zero forcing scheme when SNR is more than 16 dB, and scheme 3 has significantly improved performance compared with the ideal zero forcing scheme over the whole SNR period. The reason that the performances of the proposed schemes are better than those of conventional scheme is because the proposed schemes perform not only channel estimation but also symbol error correction by Viterbi algorithm. In Fig.9, we see that the proposed scheme 3 tracks magnitude response and phase response of a channel. This results are performed with a SNR of 16 dB and with the repeating period of the training symbol *D* = *100* at *32-th* subchannel. Through the responses are not accurately the same responses as the ideal zero forcing scheme, the excellent error performance results from the symbol error correction in SVD decoder and the accurate channel estimation by DDCE estimator and DACE estimator.



(a)

(b)

그림 9. 최대 도플러 주파스 *52Hz* 와 훈련심볼주기 100인 시스템에서의 채널 응답
(a) 크기 응답 (b) 위상 응답

Fig. 9. Channel response with maximum Doppler frequency *52 Hz* and repeating period of training symbol 100.
(a)magnitude response. (b) phase response.

Performance comparison depending on the repeating period of training symbol for scheme 3 is shown in Fig.10. This results were performed with the repeating period of the training symbol *D* = *20*, *100*, *200* and *300*. As known these results, the proposed scheme 3 is able to be obtained more excellent performance only with 0.5% overhead for training symbol than the ideal zero forcing scheme.

Fig.. 11 shows the performance comparisons depending on Doppler frequencies. It is shown that the simulation result with *D* = *100* is close to the result of ideal zero forcing scheme when maximum Doppler frequency is the same frequency as the space between subchannels, i.e. *284 Hz*. This result is shown that the proposed scheme can be used satisfactorily in fading environment to move receiver by about *16.4* m/s.

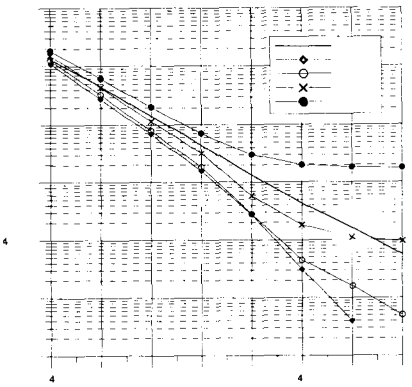


그림 10. 최대 도플러 주파수 *52 Hz* 인 방안 3의 시스템에서 훈련심볼의 주기에 따른 성능 비교

Fig. 10. Comparison of performances according to repeating period (D) of training symbol for the scheme 3 with maximum Doppler frequency *52 Hz*.

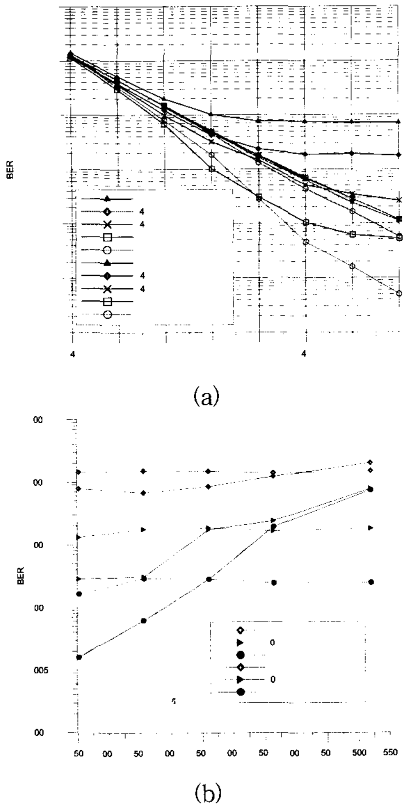


그림 11. 훈련심볼 주기 100인 시스템에서 최대 도플러 주파수에 따른 성능비교
(a) SNR에 따른 성능 (b) 최대 도플러 주파수에 따른 성능

Fig. 11. Comparison of performances being depended on maximum Doppler frequency (Df) with repeating period of training symbol 100. (a) Performance according to SNR. (b) Performance according to maximum Doppler frequencies.

V. Conclusions

In this paper, We have derived the symbol decoding scheme using Viterbi algorithm that makes the output codewords of convolutional encoder be matched to symbol branch corresponding to symbol candidates on decoder trellis. Therefore, we are able to apply directly the symbol decoding scheme using Viterbi algorithm to the systems that the output codeword of channel encoder and the modulated symbol codeword have different codeword lengths each other.

And combination schemes that are combined the symbol decoding technique with channel estimation techniques have been newly proposed for COFDM system on fading channels. We have studied the performances of the combination schemes depending on the repeating periods of training symbol and maximum Doppler frequencies. And we found that the proposed schemes are able to obtain significantly improved performance. Scheme 1 that uses only SVD decoder with TEC estimator has the performance close to that of ideal zero forcing scheme. And it is shown that the performance of Scheme 2 is significantly improved by tracking accurately time-varying channel as well as by decoding ML symbol sequence. Especially, this scheme obtains better performance than that of ideal zero forcing scheme during over SNR 16 dB. Scheme 3 has the best performance in the proposed schemes because of reducing the effect of AWGN between adjacent channels. Therefore, the proposed schemes are suitable to receiver of wireless multimedia system that accurate data transmission is required on fading channels, though the system is more complicated than the system with only zero forcing.

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