무선 패킷데이터 전송을 위한 LMS기반의 반복결정 귀환 등화기

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LMS based Iterative Decision Feedback Equalizer for Wireless Packet Data Transmission

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요 약

최근의 무선 패킷데이터 시스템에서 짧은 버스트 데이터의 전송이 많이 사용되고 있고 훈련 심볼에 의한 오버헤드가 심각한 문제를 야기할 수 있다. 따라서 적응등화기의 설계에 있어서 짧은 훈련심볼과 빠른 수렴 알고리즘이 필수적인 문제라고 할 수 있다. 본 논문에서는 짧은 훈련심볼을 사용하는 MTLMS (multiple-training least mean square) 기반의 DFE (decision feedback equalizer) 의 성능을 향상시킬 수 있는 등화알고리즘을 제안한다. 제안된 알고리즘에서 DFE의 출력은 LMS (least mean square)기반의 적응 DFE 루프로 입력되고 확장된 훈련심볼로서 사용된다. 또한 전체적인 처리를 위하여 ML (maximum likelihood) 추정기를 사용하는 블록연산 대신에 낮은 복잡도의 적응 LMS연산이 사용된다. 시뮬레이션 결과에서 제안된 등화기는 반복귀환이 증가함에 따라 성능이 향상되고 시변 페이딩에 보다 강한 성능을 보여준다.

ABSTRACT

In many current wireless packet data system, the short-burst transmissions are used, and training overhead is very significant for such short burst formats. So, the availability of the short training sequence and the fast converging algorithm is essential in the adaptive equalizer. In this paper, the new equalizer algorithm is proposed to improve the performance of a MTLMS (multiple-training least mean square) based DFE (decision feedback equalizer)using the short training sequence. In the proposed method, the output of the DFE is fed back to the LMS (least mean square) based adaptive DFE loop iteratively and used as an extended training sequence. Instead of the block operation using ML (maximum likelihood) estimator, the low-complexity adaptive LMS operation is used for overall processing. Simulation results show that the performance of the proposed equalizer is improved with a linear computational increase as the iterations parameter increases and can give the more robustness to the time-varying fading.

키워드

iterative DFE, multiple-training, least mean square, adaptive equalizer, MTLMS

Ⅰ. Introduction

A key issue toward mobile multimedia communications is to create technologies for broadband signal transmission that can support high quality services. In such a broadband mobile communications system, the transmitted signal will be

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distorted severely by the time-varying multi-path fading channel [1]-[2]. Therefore, a desired system should be designed to reject the severe ISI (inter-symbol interference) and to be robust to time-varying fading, while providing high spectral efficiency and low power consumption. In many current mobile wireless systems as well as future mobile wireless packet data system, the short-burst transmissions are used to reduce end-to-end transmission delay, and to limit the time variation of wireless channels over a burst [3]. However, training overhead is very significant for such short burst formats. So, the availability of the short training sequence and the fast converging adaptive algorithm is essential in the system adopting the symbol-by-symbol adaptive equalizer.

In the equalizer or channel estimator, the length of the training sequence affects the system performance. Since the longer training sequence can provide the more sufficient channel information, the lower MSE (mean square error) performance and the better BER (bit error ratio) performance can be acquired [4], [5]. However, the longer training sequence gives lower spectral efficiency.

In this paper, the new method is proposed to improve the performance of a MTLMS based DFE using the short training sequence. In [6]-[7], it has been shown that the performance of the channel estimator can be improved if the output of the equalizer is feedback to the channel estimator and used as an extended training sequence. This idea is extended in a MTLMS based DFE and this method is called as 'LMS based iterative DFE'. In the proposed method, the output of the DFE is feedback to the LMS based adaptive DFE loop iteratively and used as an extended training sequence. Instead of the block operation using ML estimator in [6]-[7], the low-complexity symbol-by-symbol adaptive LMS operation is used for overall processing. Throughout the computer simulations, the performance of the proposed iterative DFE is investigated and compared to that of the conventional non-iterative DFE in mobile wireless channels. Simulation results show that the proposed iterative DFE can provide the performance improvement at moderate iterations. Consequently, the proposed method can achieve high spectral efficiency, and performance gain with a linear increase in computational complexity rather than exponential increase.

This paper is organized as follows: In Section II, we describe the scheme of proposed LMS based iterative DFE. In Section III, we compare the computational complexity of proposed and other equalization algorithms. In Section IV, for the performance analysis, channel and system parameters are modeled. Computer simulations are executed and their results are discussed in Section V. Finally, in Section VI, concluding remarks are presented.

Ⅱ. Proposed LMS based iterative DFE

The proposed LMS based iterative DFE method is largely composed of two steps, the initialization loop and the iteration loop. The proposed method and its schematic block diagram are depicted in Figs. 1 and 2, respectively. The processing steps can be described as follows:

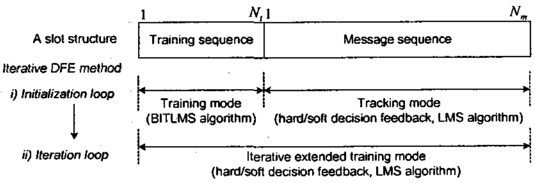

Fig. 1. The proposed LMS based iterative DFE method
그림 1. 제안된 LMS기반의 DFE 방식

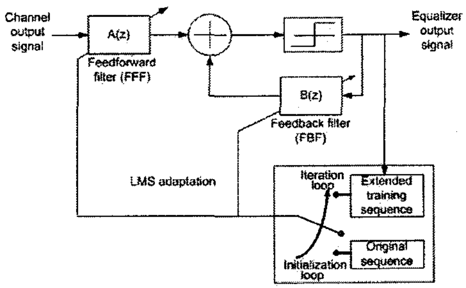

Fig. 2. Schematic block diagram of the proposed LMS based iterative DFE
그림 2. 제안된 LMS 기반의 DFE 의 블록 다이어그램구조

Initialization loop: The initial DFE tap coefficients are obtained by the training of the training sequence using the MTLMS algorithm. Using these initial coefficients, the tracking of the message sequence is performed using the LMS algorithm and hard/soft decision feedback. Then, the initial equalized data is acquired for the entire burst. Note that this MTLMS algorithm is used only for the convergence of the DFE initial tap coefficients.

Iteration loop: The initially equalized hard/soft decision output is fed back to the iterative equalization loop and concatenated with the original training sequence. This new concatenated sequence is called as an extended training sequence, which is used again to the DFE loop using the LMS algorithm. This iteratively equalized decision output makes another new extended training sequence, which is used again to the DFE loop. This iterative process may be repeated a number of times denoted by M. At k-th iteration, the n th DFE output can be written as

$$\hat{a}^k(n) = \sum_{i=0}^{N_f-1} g_f^k(n;i)x(n-i) + \sum_{j=0}^{N_b-1} g_b^k(n;j)d_{ext}^k(n-j)$$
$$\text{for } n = 1,2,\ldots,N_s,\ k = 1,2,\ldots,M \quad (1)$$

where $g_f^k(n;i)$ and $g_b^k(n;j)$ represents the FFF (feedforward filter) and FBF (feedback filter)tap coefficients, respectively. N_f and N_b are the length of FFF and FBF, respectively. $\mathbf{d}_{ext}^k$ represents the extended training sequence vector at k-th iteration and represented by $\mathbf{d}_{ext}^k = \left[\mathbf{d}^T, \left(\tilde{\mathbf{a}}^{k-1}\right)^T\right]^T$, $\tilde{\mathbf{a}}^{k-1} = [\tilde{a}^{k-1}(1), \tilde{a}^{k-1}(2), \cdots, \tilde{a}^{k-1}(N_m)]^T$ is the equalized hard/soft-decision output vector at (k-1)th iteration and $\mathbf{d} = [d(1), d(2), \ldots, d(N_t)]^T$ is the original training sequence vector. M, N_t, N_m and N_s denote the iterations parameter, the length of the original training sequence, the length of the message sequence, and the one slot length $(N_t + N_m)$, respectively. The equalizer tap coefficients are updated using the adaptive power-normalized LMS algorithm as

$$g_f^k(n+1;i) = g_f^k(n;i) + \mu_f e^k(n)x^*(n-i),$$
$$\text{for } i = 0,1,\cdots,N_f - 1 \quad (2)$$

$$g_b^k(n+1;j) = g_b^k(n;j) + \mu_b e^k(n)d_{ext}^{k\,*}(n-j),$$
$$\text{for } i = 1,2\cdots,N_b \quad (3)$$

where μ_f and μ_b represent the FFF step size and the FBF step size, respectively. $x(n-i)$ is the ith power normalized output element of the received data and given in Eqn. (4) and (5). The error function of the n-th equalized output at k-th iteration is calculated by

$$e^k(n) = \hat{a}^k(n) - d_{ext}^k(n) \quad (4)$$

Note that the initial DFE tap coefficients at next iteration are the same as the last updated coefficients at previous iteration:

$$g_f^k(1;i) = g_f^{k-1}(N_s;i) \text{ for } i = 0,1,\cdots,N_f - 1$$
$$g_b^k(1;j) = g_b^{k-1}(N_s;j) \text{ for } j = 1,\cdots,N_b \quad (5)$$

Since overall processing is based on the low complexity LMS algorithm, the computational complexity of the proposed DFE increases linearly with the iterations parameter *M*.

Ⅲ. Computational complexity

The proposed iterative DFE has a computational complexity of the summation of $K(2N+1)$ complex multiplication for the initialization loop and $M(2N+1)$ complex multiplication for the iteration loop per input sample. The computational complexities of other algorithms are given in Table 1.

Table 1. Complexity comparisons
표 1. 복잡성의 비교

Algorithms	Complex multiplications	Complex division
MTLMS	$K(2N+1)$	N
RLS	$2.5N^2+4.5N$	2
FRLS	$20N+5$	3
LMS	$2N+1$	0
NLMS	$2N+1$	N

The comparisons of computational complexities per input sample during training mode are shown in Fig. 3.

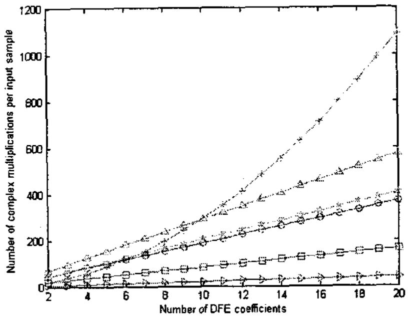

Fig. 3. Comparisons of the computational multiplications between the proposed iterative method (Circle: (K,M)=(4,5) Triangle-upward: (K,M)=(4,10)), MTLMS (Square: K=4), LMS (Triangle-right), RLS (Plus), and fast RLS (Star)
그림 3. 제안된 방식(원:(K,M)=(4,5), 삼각형: (K,M)=(4,10))과 MTLMS(네모 K=4) LMS(누운삼각형) 그리고 RLS(+표시)와 고속RLS(별표)방식의 곱셈 계산양의 비교

Note that the computational complexity of the proposed method increases linearly with the parameter *M*. With a moderate *M*, the competitive complexity can be achieved when compared to RLS algorithms for severely ISI distorted channel where the time span of channel is several tens of symbols.

Ⅳ. Channel Model and System Parameters

4.1. Channel model

Mobile radio channels can be modeled as multipath Rayleigh fading channels having an impulse response $h(t;\tau)=\sum_{l=0}^{L-1}\alpha_l(t)\delta(\tau-\tau_l)$ where the coefficient $\alpha_l(t)$ is the *l*-th multipath gain which is modeled as complex Gaussian random processes with zero mean. Under the assumptions of wide sense stationary uncorrelated scattering (WSSUS) and a common time selective correlation function ρ across the delay profile, the autocorrelation function yields

$$E[h(t;\tau)h^*(t';\tau')]=P_h(\tau)\delta(\tau-\tau')\rho(t-t') \quad (6)$$

where the superscript * denotes the complex conjugation, P_h the delay power spectral density (PSD) or the power delay profile, and ρ the time selective correlation function, normalized to unit power, that is $\rho(0)=1$. A standard version of ρ, which is derived under the assumption of isotropic scattering, is used as

$$\rho(t-t')=J_0(2\pi f_d(t-t')) \quad (7)$$

where f_d is the one-sided Doppler spread and J_0 denotes the zeroth-order Bessel function of the first kind. It is assumed that the path delay is integer multiplies of T_s, i.e., $\tau_l=lT_s$, $l=0,1,\cdots,L-1$. In addition, the channel gain is normalized, i.e., $\sum_{l=0}^{L-1}E\left[|\alpha_l(t)|^2\right]=\sum_{l=0}^{L-1}P_l=1$. The minimum phase channel $(h_1(t,\tau))$ and the non-minimum phase channel $(h_1(t,\tau))$ models are considered and their power delay profiles are given in Table 2.

Table 2. Power delay profiles
표 2. 전력 지연프로파일

Path (l)	$h_1(t;\tau)$: 4-path minimum phase channel model		$h_2(t;\tau)$: 9-path nonminimum phase channel model	
	Delay τ_l	Average power, P_l	Delay, τ_l	Average power, P_l
0	0	0.575	0	0.213
1	T_S	0.362	T_S	0.405
2	$2T_S$	0.057	-	-
3	$3T_S$	0.006	$3T_S$	0.263
8	-	-	$8T_S$	0.119

For convenience of notation, the parameter τ is dropped in $h_i(t;\tau), i=1,2$. Note that these channel coefficients are similar to COST-207 channel coefficients used in [8] (some modification is done for urban model). The channel $h_1(t)$ has the time dispersion of $3T_s$ and the minimum phase property. The channel $h_2(t)$, which represents more severely distorted ISI channel, has the time dispersion of

$8T_s$ and the non minimum phase property. It is shown that $h_2(t)$ has the severe frequency selectivity and gives the worse channel characteristics than $h_1(t)$.

For representing the time-variation of the mobile radio channel, the symbol-normalized fade rate $f_d T_s$ is the useful index, where f_d is the maximum Doppler frequency and T_s is the symbol duration. However, in a channel structure which incorporates periodic transmission of training bits, $f_d T_s$ is an incomplete measure of the rate at which the channel varies. Since the training and tracking of the channel is performed during the transmission of the TDMA slot, the slot-normalized fade rate is used as a fading channel index and is defined as [9]

$$\kappa = f_d T_{slot} = f_d N_s T_s \tag{8}$$

where T_{slot} is the slot duration and N_s is the number of symbols per time slot. This number represents the rough average number of occurrences of deep fades per time slot.

4.2. System parameters

A QPSK signal is transmitted. Each transmitted burst contains the training sequence of variable length A and the message sequence of the length of 144 (only for the purpose of the simulation). The length of the training sequence are 12, 16, 20, 24, 28, 32 and64 which correspond to the overhead of about 7.7%, 10%, 12.2%, 14.3%, 16.3%, 18% and 31%, respectively. Note that the small overhead represents high spectral efficiency.

The carrier frequency is 5GHz and the channel bandwidth is 1MHz. The symbol interval is $1(\mu s)$. For 4-path channel model $h_1(t)$, the FFF length was set to be 7 and the FBF length was set to be 4. For 9-path channel model $h_2(t)$, the FFF length was set to be 11 and the FBF length was set to be 9. The FFF step size was 0.05 and the FBF step size was 0.005 for both channels. For the soft decision feedback DFE, $\gamma = 5dB$ is used [10].

Ⅴ. Simulation results

Since the proposed iterative DFE is much affected by the iterations number, the performance is investigated in terms of the iterations parameter M. The effects of hard or soft decision feedback are also investigated. In the computer simulations, SNR was 18dB, the FFF step size was 0.05, the FBF step size was 0.005, the MTLMS parameter K was 4, the length of the training sequence was 16, and the overhead was 10%. The channel is assumed to be slow fading with the normalized Doppler frequency of 0.00012.

Fig. 4 shows the BER performance of the proposed iterative DFE as a function of the parameter M . As M increases, the performance becomes better but the computational complexity increases linearly with the iterations parameter M. So the parameter M and the performance are tradeoffs. When the value of M varies from 1 to 5, the large improvement is achieved and the performance is saturated with further increase. When the performance gain and the increase of the computational complexity are considered, the value of $M=5$ will be a proper choice. When a soft decision feedback is used, the performance improvement is achieved. This performance improvement is more obvious in $h_2(t)$.

Fig. 5 and 6 show the BER performance of the proposed iterative DFE as a function of the SNR (E_b/N_0) in $h_1(t)$ and $h_2(t)$, respectively. The MTLMS parameter K was 4 and the soft decision feedback was used. The slow fading with a normalized Doppler frequency of 0.00012 was assumed. It is shown that the performance of the proposed DFE is improved with the increase of the parameter M when the 16 training symbols were used. Furthermore, at high SNR, the performance of the proposed method is better than that of the non-iterative DFE with the 32 or 64 training symbols. However, at low SNRs, the performance improvement is very slight. From simulation results, it is shown that the proposed method can shorten the length of the required training sequence effectively with a linear increase in computational complexity. So, the spectral efficiency can be achieved.

In Fig.7 and 8, the BER performance of the proposed iterative DFE is shown as a function of the normalized Doppler frequency in $h_1(t)$ and $h_2(t)$, respectively. As the normalized Doppler frequency increases, overall performance becomes worse since the time slot with a deep fade, where the reliable equalizer outputs cannot be achieved, is increased during given time. It is shown from simulation results that the proposed method can give more robustness to the time-varying fading.

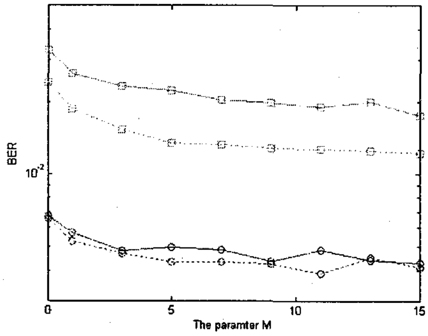

Fig. 4. BER performance of the proposed iterative DFE. E_b/N_0=18dB, K=4, 16 training symbols, Solid line: hard decision; Dotted line: soft decision, Circle: $h_1(t)$ Square: $h_2(t)$.

그림 4. 제안된 DFE방식에 대한 BER 성능. E_b/N_0=18dB, K=4, 16개의 훈련심볼, 실선: 경판정, 점선:연판정, 원표시: $h_1(t)$ 네모표시: $h_2(t)$

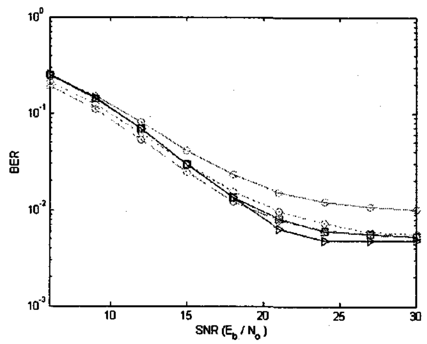

Fig. 5. BER of the proposed iterative DFE in $h_1(t)$. K=4, Solid line: 16 training; Dotted line: 32 training; Dashed line: 64 training, Circle: M=0 Square: M=5 Triangle-right: M=10.

그림 5. $h_1(t)$채널에서의 제안된 DEF의 BER 성능. K=4, 실선: 16개 훈련심볼, 점선: 32개 훈련심볼, 데쉬선: 64개 훈련심볼, 원표시: M=0 네모표시: M=5 삼각형: M=10.

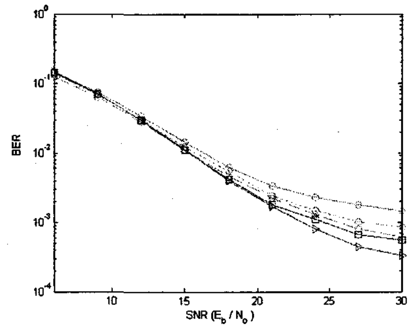

Fig. 6. BER of the proposed iterative DFE n $h_2(t)$. K=4, *Solid line: 16 training; Dotted line: 32 training;* Dashed line: 64 training, Circle: M=0 Square: M=5 Triangle-right: M=10.

그림 6. $h_2(t)$채널에서의 제안된 DEF의 BER 성능. K=4, 실선: 16개 훈련심볼, 점선: 32개 훈련심볼, 데쉬선: 64개 훈련심볼, 원표시: M=0 네모표시: M=5 삼각형: M=10.

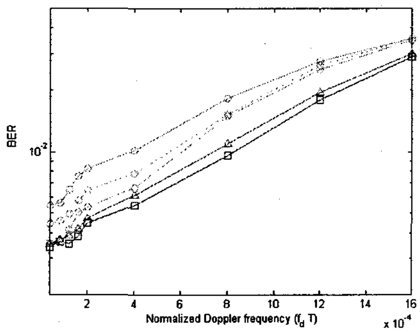

Fig. 7. BER of the proposed iterative DFE in $h_1(t)$ according to Doppler frequency. E_b/N_0=18dB, K=6, Solid line: 16 training; Dotted line: 32 training; Dashed line: 64 training, Circle: M=0 Triangle-upward: M=5 Square: M=10.

그림 7. $h_1(t)$채널에서 도플러 주파수에 따른 제안된 DEF의 BER 성능. E_b/N_0=18dB, K=4, 실선: 16개 훈련심볼, 점선: 32개 훈련심볼, 데쉬선: 64개 훈련심볼, 원표시: M=0 삼각형표시: M=5 사각형표시: M=10

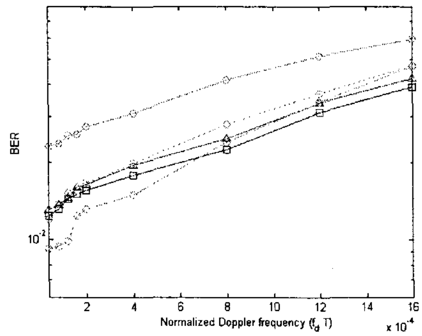

Fig. 8. BER of the proposed DFE in $h_2(t)$ according to Doppler frequency. E_b/N_0=18dB, K=4, Solid line: 16 training; Dotted line: 32 training; Dashed line: 64 training, Circle: M=0 Triangle-upward: M=5 Square: M=10

그림 8. $h_2(t)$ 채널에서 도플러 주파수에 따른 제안된 DEF의 BER 성능. E_b/N_0=18dB, K=4, 실선: 16개 훈련심볼, 점선: 32개 훈련심볼, 데쉬선: 64개 훈련심볼, 원표시: M=0 삼각형표시: M=5 사각형표시: M=10.

Ⅵ. Conclusion

The LMS based iterative DFE scheme, where the output of the DFE was feedback to the LMS based adaptive DFE loop iteratively and used as an extended training sequence, was proposed to overcome the performance degradation of a conventional DFE using a short training sequence. Simulation results show that the performance of the proposed method is improved with a linear computational increase as the iterations parameter (*M*) increases. In addition, it is shown that the proposed method can give the more robustness to the time-varying fading. Since the length of the training sequence can be reduced effectively, high spectral efficiency can be acquired. When the performance gain and the increase of the computational complexity are considered, the value of *M*=5 will be a proper choice.

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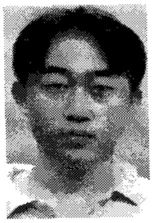

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