Iterative Extrapolation for Channel Equalization in DVB-T Receivers

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***Abstract*—This paper proposes an iterative extrapolation method to improve the channel-equalization performance for DVB-T receivers. The coefficients of a one-tap equalizer are conventionally obtained by taking the reciprocals of the estimated channel coefficients. Inversion of the channel coefficients, however, may cause noise enhancement at those frequencies corresponding to spectral nulls. To reduce such undesirable noise amplification, iterative extrapolation based on the time-domain truncation is performed, resulting in smooth nulls. Simulation results show that the proposed method improves the channel-equalization performance compared to the conventional zero-forcing equalization method in terms of bit error rates in DVB-T receivers at all SNRs.**

***Index Terms*—DVB-T, equalization, extrapolation, noise enhancement, OFDM.**

I. Introduction

IN DIGITAL Video Broadcasting-Terrestrial (DVB-T), the European digital terrestrial television standard [1], orthogonal frequency division multiplexing (OFDM) has been adopted for signal transmission. In OFDM systems, the entire channel is divided into many narrow subchannels, which are transmitted in parallel. This results in the increase of the time duration corresponding to an OFDM symbol. With the insertion of cyclic prefix for the guard interval (GI), the increased symbol duration reduces the inter-symbol interference (ISI). Assuming that the orthogonality among the subcarriers is maintained and the multipath delay is shorter than the GI, a frequency domain one-tap equalizer could be used for each subchannel to correct the amplitude and phase distortions [2], [3].

While extensive researches on OFDM channel equalization have already been carried out [3]–[5], available equalizers, such as zero-forcing (ZF) and minimum-mean-square-error (MMSE) equalizers still have some drawbacks. The coefficients of a one-tap ZF equalizer are obtained by taking the reciprocals of the estimated channel coefficients. In case that a channel has deep nulls, however, the equalization performance may suffer from noise enhancement caused by a large coefficient of the equalizer. MMSE equalizers can be used to avoid the noise enhancement but require both previous knowledge of the system's noise variance and matrix inversion, resulting in high computational complexity [6].

There are several studies devoted to channel estimation and equalization for DVB-T receivers [7]–[10]. In [7] and [8], the channel estimation was performed based on block averaging to reduce the noise effect in frequency domain. These methods, however, are vulnerable to time-varying channels. In [9], a frequency-domain pilot time-domain correlation (FPTC) method was used for the estimation of the channel characterization. While the FPTC method is simple to implement and is robust to synchronization errors, its use is partially limited to the estimation of channels with the delay spreads less than 1/12 of the OFDM symbol duration. The channel-estimation scheme presented in [10] used an adaptive two-dimensional wiener filter implemented as a cascade of two one-dimensional filters. This method may achieve a good channel-estimation performance by adapting the bandwidth of the second filter according to a channel but it is highly dependent on the accuracy of the estimated maximum delay spread of the channel.

In this paper, we consider a trade-off between the practicability and the equalization performance and lead to an appropriate candidate for DVB-T receivers with a ZF equalizer robust to noise effect. To achieve this, a smoothing method of spectral nulls is proposed to reduce noise amplification in a one-tap ZF equalizer by iterative extrapolation.

This paper is organized as follows. In Section II, the pilot-based OFDM system, channel estimation, and an equalization problem are presented. Section III describes the proposed iterative extrapolation method for a one-tap ZF equalizer in DVB-T receivers. Simulation results given in Section IV show that the proposed method improves the bit-error-rate performance obtained after one-tap equalization, compared to the conventional equalization method in DVB-T receivers. Section V concludes the paper.

II. System Description

A. Baseband Model of a Pilot-Based OFDM System

Fig. 1 shows a baseband model of a typical pilot-based OFDM system [11]. The binary information data are grouped and mapped according to the modulation scheme, such as 16-QAM (quadrature amplitude modulation) and 64-QAM. After the pilot insertion, the modulated data are inputted to an IFFT (inverse fast Fourier transform) block and are transformed into a time-domain signal. To prevent possible ISI in OFDM systems, the guard interval (GI) is inserted with a cyclic prefix (CP) which contains a copy of the last part of the OFDM symbol. The transmitted signal is then passed a frequency selective multipath fading channel with additive white Gaussian noise (AWGN). At the receiver, the GI is removed and the received symbols are sent to an FFT block to demultiplex the multicarrier signal.

In OFDM transmission schemes, there are two major types of pilot arrangement. The first kind is referred to a block type. The pilot signals are assigned to a particular OFDM symbol, which are sent periodically in time domain. The second one is

Manuscript received July 25, 2007; revised March 27. 2008. First published May 30, 2008; last published August 20, 2008 (projected). This work was supported by the 2nd-phase Brain Korea (BK) 21 Program funded by the Ministry of Education, Korea.

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Digital Object Identifier 10.1109/TBC.2008.2000465

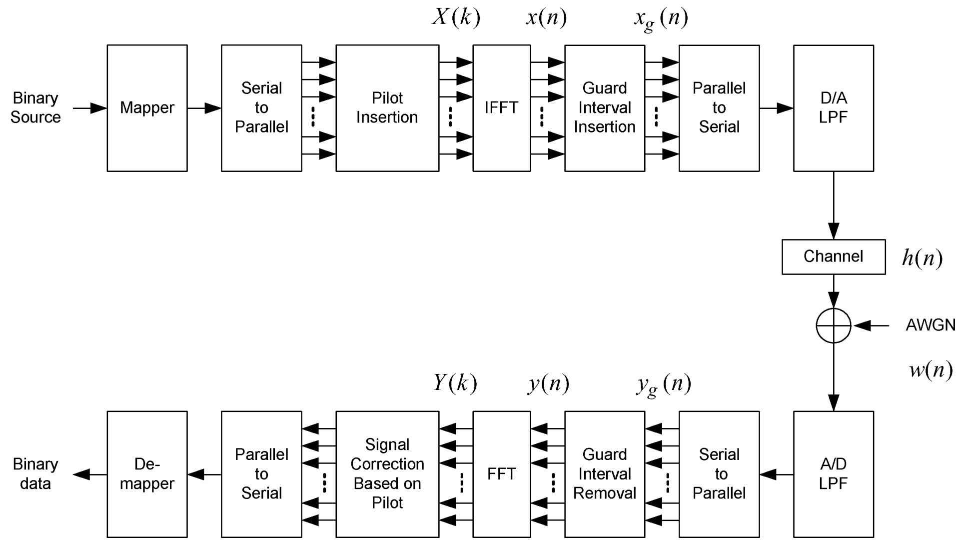

Fig. 1. Baseband model of a pilot-based OFDM system.

a comb type. The pilot signals are uniformly distributed within each OFDM symbol. In a comb-type arrangement, since only some subcarriers contain the pilots, the channel response corresponding to non-pilot subcarriers will be estimated by interpolating the coefficients of their neighbor pilot subchannels [11].

B. Channel Estimation in the DVB-T System

In the DVB-T system, pilot cells are inserted every four OFDM symbols at the same position and are uniformly distributed every twelve subcarriers within each OFDM symbol. Hence the pilot pattern in DVB-T can be viewed as a block type as well as a comb type. In this respect, the pilot-insertion type approximation and block-based channel estimation scheme may be efficient for estimating the channel coefficients [7].

When the duration of the channel impulse response is shorter than the GI, there is no ISI. Assuming also that there is no synchronization error, the FFT of the received signal $Y_p(k)$ can be formulated as

$$Y_p(k) = H_p(k) \cdot X_p(k) + W_p(k), \tag{1}$$

where $H_p(k)$ is the channel transfer function, $W_p(k)$ is the FFT of the time-domain AWGN [9], and the subscript p denotes a pilot. For the pilot subcarriers, the transmitted information $X_p(k)$ is known to the receiver. Therefore, the frequency response of the channel at the pilot frequencies can be estimated simply using

$$\hat{H}_p(k) = \frac{Y_p(k)}{X_p(k)} = H_p(k) + W'_p(k), \tag{2}$$

where $W'_p(k)$ is the noise effect existing at the estimated channel coefficients. The channel estimation scheme used in (2) is based on the Least-Squares (LS) method [12]. In order to estimate the channel coefficients at data subcarriers, we use the piecewise linear interpolation method due to its simplicity [13], which is given by

$$h_{Lin}(x) = \begin{cases} 1 - |x|, & 0 \le |x| < 1 \\ 0, & \text{elsewhere.} \end{cases} \tag{3}$$

Through the above interpolation, we can obtain all channel coefficients.

C. Equalization Problem

Let $h(n)$ be the channel impulse response and $H(k)$ its Fourier transform, i.e., the channel transfer function. If the number of carriers is sufficiently large, the channel transfer function becomes virtually non-selective within the bandwidth of individual carriers. Focusing on one particular carrier, the influence of multipath fading reduces to attenuation and a phase rotation. For an OFDM system, assuming that the maximum multipath delay is shorter than the GI, a frequency-domain one-tap equalizer could be used for each subchannel to correct the amplitude and phase distortions [2], [3]. The reciprocals of the estimated channel coefficients are used for one-tap equalizer coefficients. This equalization scheme is based on the ZF criterion [14] which aims at canceling ISI regardless of noise level. The ZF criterion, however, may not provide a desirable solution when the channel transfer function has deep spectral nulls in the signal bandwidth. At the frequencies corresponding to spectral nulls, inversion of the channel transfer function requires a large gain, resulting in noise enhancement.

III. Iterative Extrapolation

As described in the previous section, the performance degradation of the one-tap ZF equalizer is inevitable due to the noise enhancement when deep nulls exist in the channel frequency response. We show one example of the noise enhancement case through the computer simulation. The channel profile used for this simulation is "Brazil channel D," which was the indoor channel used for the Laboratory Test in Brazil [15]. The channel information is given in Table I.

Fig. 2 shows the estimated channel coefficients in a DVB-T receiver with the guard band, where there exist deep spectral nulls. DVB-T mode is 2K and the constellation scheme is 16-QAM. Fig. 3 shows the coefficients of the one-tap ZF equalizer obtained by taking the reciprocals of the estimated channel coefficients. The equalizer has very large coefficients

TABLE I
MULTI-PATH PROFILE (BRAZIL CHANNEL D)

Delay (μs)	Amplitude (dB)
0.0	-0.1
+0.48	-3.9
+2.07	-2.6
+2.90	-1.3
+5.71	0.0
+5.78	-2.8

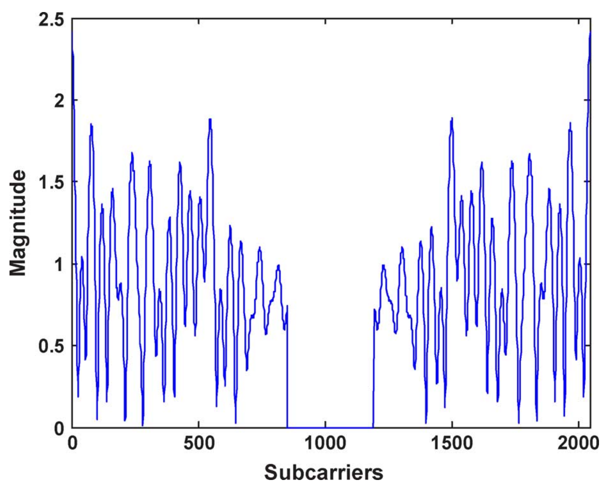

Fig. 2. Estimated channel coefficients in a DVB-T receiver with the guard band.

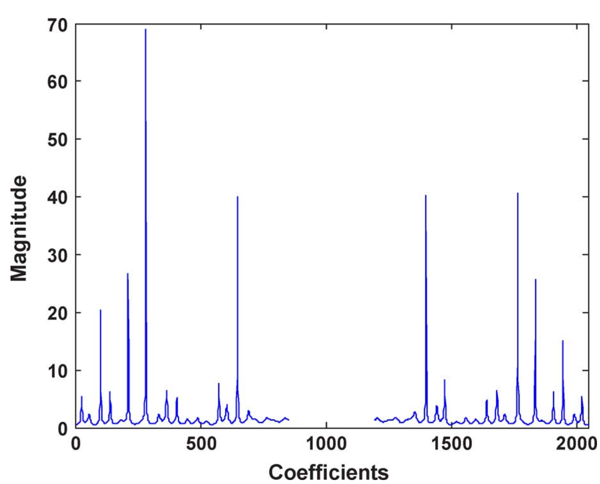

Fig. 3. Coefficients of the equalizer in the DVB-T receiver.

at those frequencies corresponding to spectral nulls. These large coefficients cause noise enhancement and thus degrade the system performance.

To reduce such a noise enhancement problem, we propose an iterative extrapolation method based on time-domain constraint which prevents the equalizer from taking large coefficients by smoothing spectral nulls. Fig. 4 shows a block diagram of the proposed extrapolation method. First, we estimate channel coefficients using the received signal after FFT. Then, zeros are inserted in the guard band as the same procedure at the

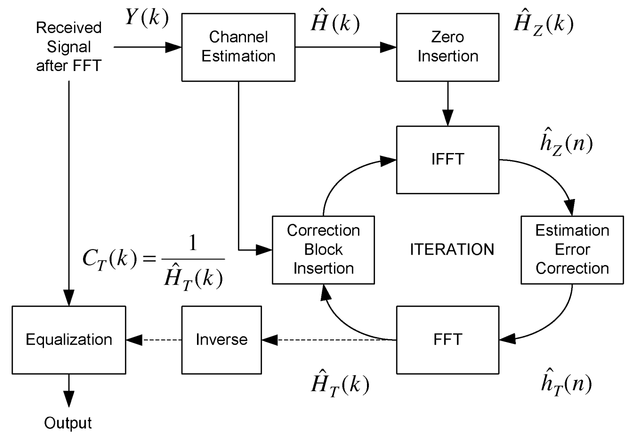

Fig. 4. Block diagram of the proposed method.

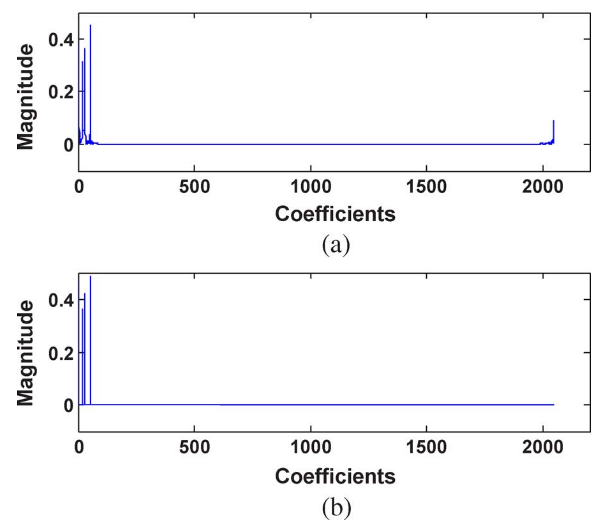

Fig. 5. Channel impulse responses: (a) the estimated channel impulse response, (b) the original channel impulse response.

transmitter. The estimated channel coefficients with the inserted zeros $\hat{H}_Z(k)$ are transformed into time-domain coefficients as:

$$\hat{h}_Z(n) = \sum_{k=0}^{N-1} \hat{H}_Z(k) \exp\left(\frac{j2\pi nk}{N}\right), \tag{4}$$

where N is the number of subcarriers. In time-domain transformed channel coefficients, some undesirable error terms may appear due to the guard band which is inserted to prevent the intercarrier interference by adjacent channels.

Figs. 5(a) and 5(b) show the estimated channel impulse response and the original channel response, respectively. The length of the channel response is shorter than that of the GI. This means that the estimated channel has no response over the GI. Comparing Fig. 5(a) with 5(b), however, some residual coefficients are appeared at the end of the estimated channel impulse response. Since it is desirable that those coefficients should be removed for the precise channel estimation, we suppress them to zero at the estimation-error-correction block.

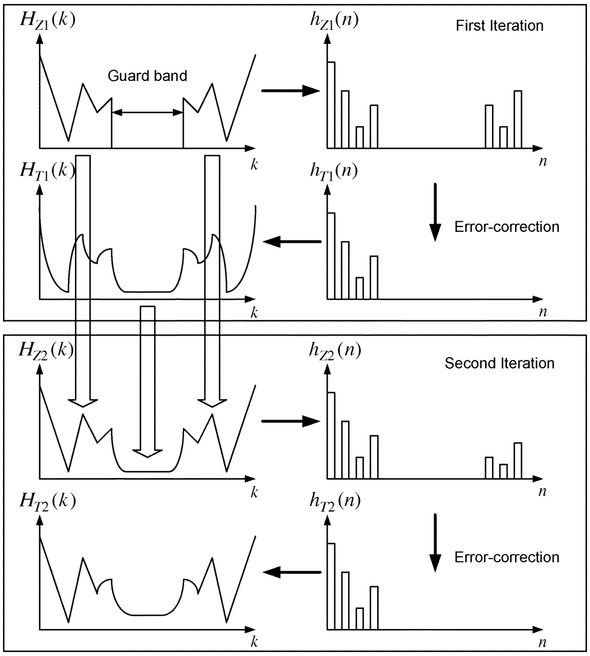

Fig. 6. An example of iterative processes of the proposed method.

The truncated coefficients $\hat{h}_T(n)$ are obtained with the use of a time-domain window as:

$$\hat{h}_T(n) = \hat{h}_Z(n) \cdot w_T(n), \tag{5}$$

where $w_T(n)$ is a rectangular window function with the length of a maximum channel delay. When the length of a channel is not available, the length of the window can be determined by the guard interval.

After the error correction, the coefficients $\hat{h}_T(n)$ are transformed into frequency-domain coefficients at the FFT block as:

$$\begin{aligned}\sum_{n=0}^{N-1} \hat{h}_T(n) \exp\left(-\frac{j2\pi nk}{N}\right) &= \sum_{n=0}^{N-1} \hat{h}_Z(n) \cdot w_T(n) \exp\left(-\frac{j2\pi nk}{N}\right) \\ &= \hat{H}_Z(k) * W_T(k) \\ &= H_{T1}(k). \end{aligned} \tag{6}$$

As shown in (6), the frequency-domain coefficients of $\hat{H}_z(k)$ are smoothened by the convolution with the sync function of $W_T(k)$ and thus the spectral nulls become milder through these repeated processes.

Fig. 6 illustrates the results of each step of the iterative extrapolation method, where for the convenience of explanation frequency samples are continuously represented. The first step is to take the IFFT of the estimated channel coefficients with the inserted zeros and remove the residual coefficients appeared at the end of the channel impulse response. Then, the modified impulse response is transformed into a frequency-domain signal. Since the removal in time-domain affects all subcarriers, the zeros in the guard band are extrapolated as shown in $H_{T1}(k)$ of Fig. 6. In the next iteration, the LS estimated channel coefficients are used again in the in-band but the extrapolated coefficients obtained by the first iteration replace the zeros in the guard band. These iterative procedures yield smoothed frequency-domain coefficients as shown in Fig. 7 where all the depths of the spectral nulls become shallow.

Fig. 8 shows the channel coefficients obtained by the first iterative process, the ideal channel coefficients, and the estimated channel coefficient by the conventional LS method. Comparing the coefficients obtained by iteration with the LS-estimated coefficients, we can see that the difference between the adjusted coefficients by iteration process and the ideal coefficients are larger than that of the conventional method. In other words, we do not reach the desirable estimation results with only one iterative process.

This can be easily overcome by increasing the number of iterations. Fig. 9 shows the adjusted channel coefficients after the fifth iteration, the ideal channel coefficients, and the LS-estimated channel coefficients. Comparing with the estimation error between the adjusted coefficients and the ideal coefficients in Fig. 8, the corresponding one in Fig. 9 remarkably decreases. The relation between the estimation error and the iteration number can be clearly shown in the zoomed version of channel coefficient plots in Fig. 10. Although the difference between the error-corrected coefficients by iteration process and the ideal coefficients seems to be larger than that of the conventional method, the coefficients appeared in the guard band do not involve the system performance. Through extensive simulations, we get to the result that in most cases "five-iteration" is satisfactory though the iteration number may be varied with the channel condition and the required estimation performance.

IV. Simulation Results

We performed computer simulations to verify the performance of the proposed method applied to the DVB-T system. The channel profile was "Brazil channel D," which was already described in Section III. "Brazil channel A" was used for verifying the robustness to channel characteristics of the propose method. The channel information of "Brazil channel A" is described in Table II [15]. To consider the effect of the Doppler spread, we also performed the simulation under COST 207 TU6 channels [16] with the vehicle speeds of 60 km/h and 120 km/h, respectively. The profile of the COST 207 TU6 channel is indicated in Table III. The DVB-T system parameters used in the simulation are shown in Table IV. We assumed that there was no synchronization error, since the purpose of the proposed method lies in enhancing the performance of the channel equalization. The code rate was 1/2 and the generator polynomials of the mother code are $G_1 = 171_{\mathrm{OCT}}$ and $G_2 = 133_{\mathrm{OCT}}$ for output, respectively. DVB-T mode was decided to fix onto the 2K and 8 MHz derivative of the standard. Moreover, we chose the GI to be greater than the maximum delay spread in order to avoid ISI. As shown in Table I, the maximum channel delay is $+5.78\ \mu\mathrm{s}$ and the elementary symbol period T is

$$T = \frac{7}{64}\ \mu\mathrm{s} = 0.109375\ \mu\mathrm{s} \tag{7}$$

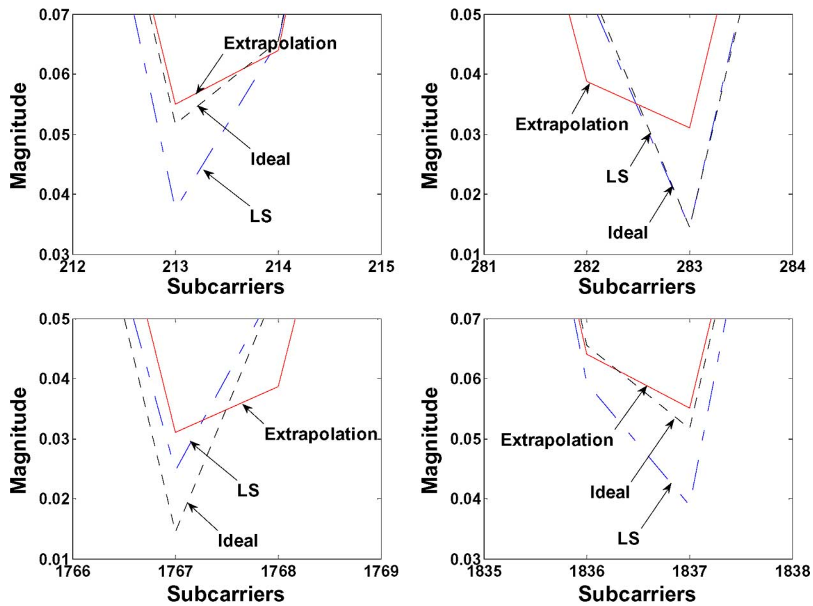

Fig. 7. Channel coefficients obtained by the fifth iteration corresponding to the spectral nulls.

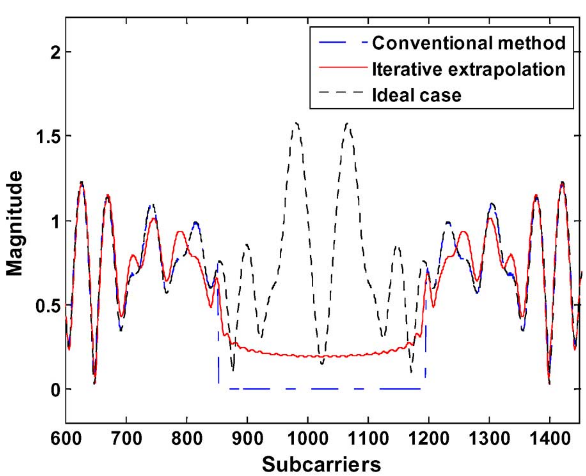

Fig. 8. Channel coefficients obtained by the one iteration.

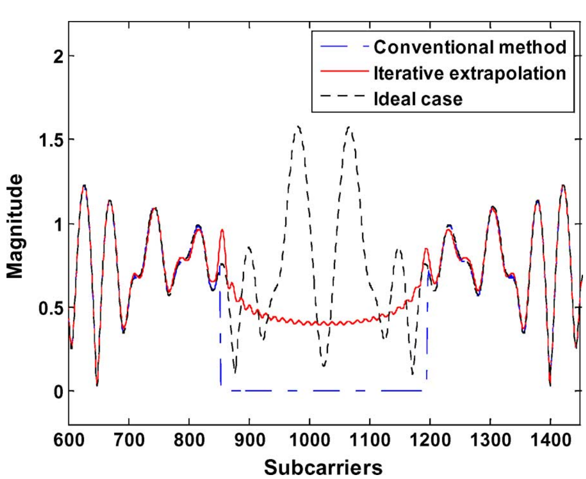

Fig. 9. Channel coefficients obtained by the five iterations.

for an 8 MHz channel [1]. The maximum channel delay corresponds to about 53 elementary symbols. Since the length of

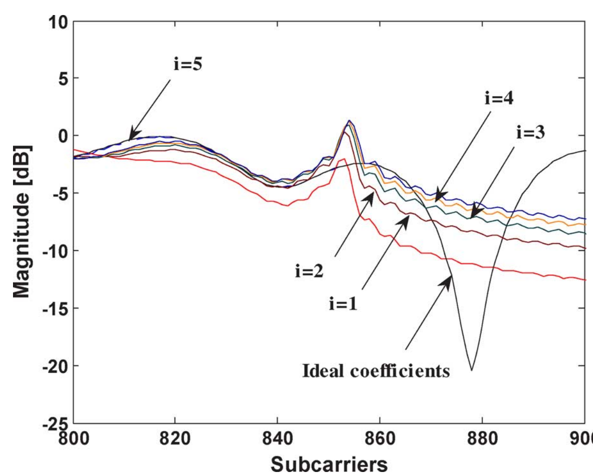

Fig. 10. Estimated channel coefficients for the iteration number of $i = 1, 2, 3,$ 4 and 5.

TABLE II
MULTI-PATH PROFILE (BRAZIL CHANNEL A)

Delay (μs)	Amplitude (dB)
0.0	0.0
+0.15	-13.8
+2.22	-16.2
+3.05	-14.9
+5.86	-13.6
+5.93	-16.4

the GI is longer than that of the maximum channel delay, there is no ISI. Transmitted data were mapped based on 16-QAM. We used the 7-point piecewise linear interpolation method for channel estimation.

At first, we obtained a mean-square error (MSE) in dB between the modified coefficients by the proposed method and the ideal channel coefficients to find a proper number of iterations for performance enhancement of the DVB-T receiver with the "Brazil channel D." The adopted SNR values were 5 dB, 10 dB,

TABLE III
MULTI-PATH PROFILE (COST 207 TU6)

Delay (μs)	Amplitude (dB)
0.0	-3.0
+0.2	0.0
+0.5	-2.0
+1.6	-6.0
+2.3	-8.0
+5.0	-10.0

TABLE IV
SIMULATION PARAMETERS

Parameters	Specifications
DVB-T mode	2K
Number of carriers	1705
OFDM symbol duration	224 μs
Guard Interval	1/4 (512)
Signal Constellation	16-QAM
Channel Model	Brazil channel A and D COST 207 TU6

TABLE V
MEAN-SQUARE ERRORS

Iteration	SNR: 5 dB (-7.07)	SNR: 10 dB (-12.07)	SNR: 15 dB (-17.01)	SNR: 20 dB (-21.87)
1	-9.65	-13.83	-16.80	-18.47
2	-9.95	-14.51	-18.38	-21.05
3	-9.96	-14.77	-19.08	-22.35
4	-10.01	-14.89	-19.34	-23.08
5	-10.01	-14.89	-19.56	-23.54

15 dB and 20 dB, respectively. The values in the parentheses are the MSE results between channel coefficients obtained by the LS estimation and the ideal channel coefficients. As shown in Table V, which described the MSE results, the five iterations are sufficient to draw a good performance of the equalizer in the DVB-T system.

Fig. 11 shows the BER performance of the proposed channel equalization method using the linear interpolation with 16-QAM. The legends of "Conventional method" and "Ideal case" denote the equalizer with the coefficients obtained by taking the reciprocals the estimated channel coefficients based on LS and that with the ideal channel coefficients, respectively. Under the "Brazil channel A," the proposed equalization method improves the BER performance of the conventional ZF equalizer about 2 dB at all signal-to-noise ratios (SNRs). When the "Brazil channel D" is considered, the iterative extrapolation method still improves the system performance by more than 1.5 dB even though the channel condition becomes extremely severe.

Fig. 12 shows the BER performance of the proposed channel equalization compared to the conventional one-tap ZF equalization. The tested vehicle speeds were 60 km/h and 120 km/h, respectively. At high SNRs, the BER performance of the system is highly degraded caused by Doppler spread. However, the proposed iterative method improves the system performance by more than 1.2 dB because the spectral nulls become milder due

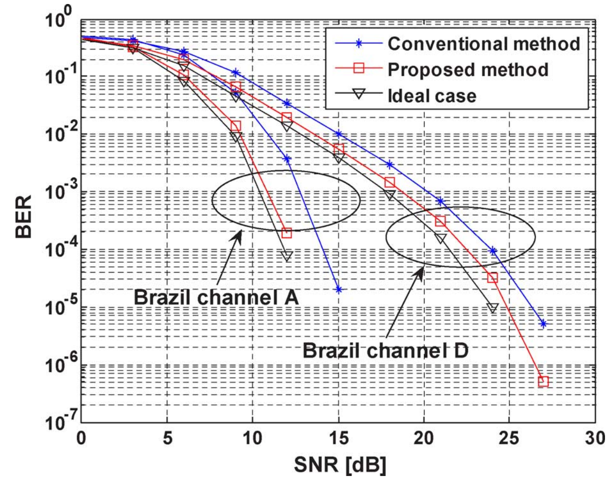

Fig. 11. BER performance of the proposed method with the five iterations under the Brazil channel A and D using 16-QAM.

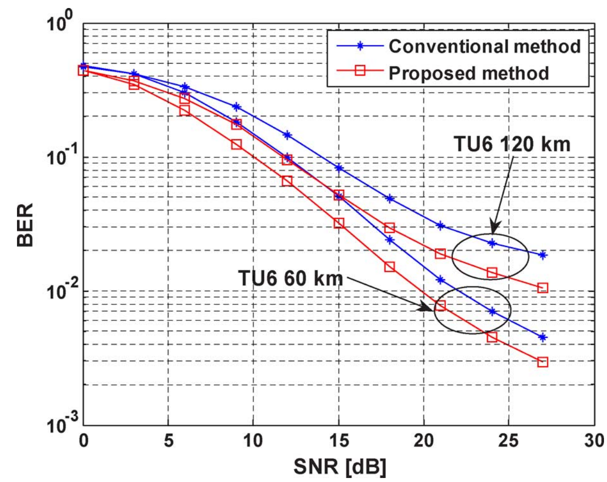

Fig. 12. BER performance of the proposed method with the five iterations under the COST 207 TU 6 channel using 16-QAM.

to the time-domain low-pass filtering effect obtained by the use of the time-domain window.

To verify the robustness of the proposed iterative extrapolation method to severe frequency-selective channels, we performed simulations with a modified "Brazil channel A." The corresponding results are described in Fig. 13. The longest delay of the channel, as given in Table II, was changed to $+16.4$ μs and the GI was 1/8 of 28 μs. As shown in Fig. 13, even when the channel has a long maximum delay spread and the short GI of 1/8, the proposed method achieves a better BER performance than the conventional ZF method.

It should be noted that the proposed method would be very useful for some OFDM applications using a time-domain equalizer, such as an equalization on-channel repeater. The coefficients of the time-domain equalizer are usually obtained by taking an IFFT of the frequency-domain equalizer's coefficients, which are the reciprocals of the estimated channel coefficients. Since the zeros are inserted in the guard band, we can not directly obtain the frequency-domain coefficients of the equalizer corresponding to the guard band because the inverse of the zero is infinite. However, the proposed method

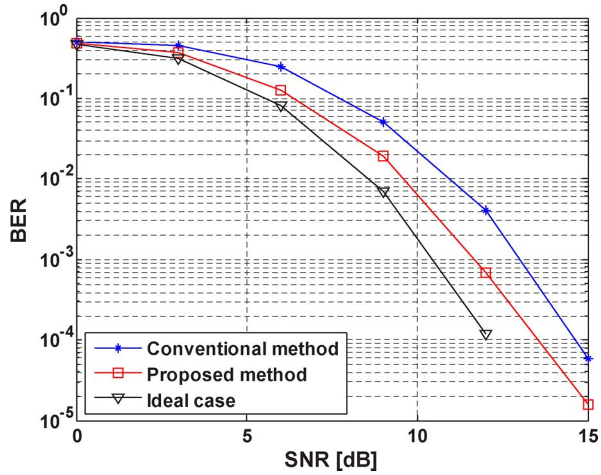

Fig. 13. BER performance of the proposed method with the five iterations under the modified Brazil channel A.

enables to use a time-domain equalizer based on the full-band frequency-domain equalizer by extrapolating the zeros in the guard band.

V. Conclusion

We proposed an iterative extrapolation method for channel equalization in DVB-T receivers. The proposed method effectively improved the channel equalization performance at all SNRs in terms of the BER by smoothing deep nulls of the channel. It is expected that the iterative extrapolation method contributes to the performance improvement of the equalizer in any OFDM systems irrespective of channel conditions. In addition, the proposed method will be especially useful for OFDM applications requiring a time-domain equalizer by removing the frequency leakage of the guard band.

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