Analysis on Channel Estimation for the Equalization in ATSC DTV Receivers

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SUMMARY

This paper presents analysis results on finite-impulse response (FIR) channel estimation used for the equalization in Advanced Television Systems Committee digital television receivers. While channel estimation results have been effectively used for the equalization, the conditions of sufficient order and high signal-to-noise ratio (SNR) were assumed in most cases. To compensate for these unrealistic assumptions, we consider diverse probable conditions for channel estimation, such as reduced order and low SNRs, and then theoretically analyze each estimation case. The analysis shows that the adaptive FIR channel estimator provides an unbiased estimation and matches well its corresponding channel coefficients irrespective of the number of taps of the estimator and the non-causality of the unknown channel. Simulation results verify our analysis on the estimation of terrestrial DTV channels.

key words: channel estimation, equalization, DTV

1. Introduction

Advanced Television Systems Committee (ATSC) digital television (DTV) receivers have commonly used decision feedback equalizers (DFEs) to suppress intersymbol interference in the 8-vestigial sideband (VSB) signal [1]. The DFE used for ATSC DTV receivers, however, has suffered from error propagation phenomenon which causes the degradation of the convergence performance of the DFE because blind equalization or decision-directed equalization using slicer outputs has to be carried out in most received symbols [2]. As the symbol error rate (SER) can be as high as 0.2 and the training sequence is very short in the terrestrial DTV receiver, error propagation is unavoidable during the reception of unknown data symbols [2]. Many researchers have proposed novel equalization methods to solve the problem and thus effectively equalize the 8-VSB signal [3]–[6].

Among these methods, the equalization schemes presented in [4] and [6] are remarkable because they can deal with severe channels, for example, a channel having a near 0 dB ghost. Both the equalization schemes incorporated a channel-matched filter (CMF) obtained from channel estimation. The CMF makes the channel property more manageable in the DFE and thus reduces error propagation effects. In the two cases, however, channel estimations were performed under the conditions of the sufficient-order and the signal-to-noise ratio (SNR) values of more than 18 dB. Since terrestrial DTV channel conditions are very diverse, the assumptions may not be practical.

To compensate for these unrealistic assumptions and secure the application of the channel estimation results to the DFE in DTV receivers, this paper analyzes some properties of the channel estimation performed under probable channel conditions: reduced-order cases where the number of taps of a channel estimator is smaller than the order of an unknown channel to be estimated and relatively low SNR values of less than 18 dB. Furthermore, terrestrial DTV channels may be often non-causal because there exist both pre-ghosts and post-ghosts in real reception environments. The analysis on channel estimation under these situations is first performed based on theoretical results presented in [7] and [8]. Then, we show the validity of our theoretical analysis results through computer simulations.

2. Equalization for ATSC DTV Receivers

The 8-VSB signal is transmitted in “frames,” as shown in Fig. 1. Each data frame is composed of two data fields, each containing 313 “segments,” of which the first segment is the “field sync” segment, followed by 312 “data segments.” Each data segment is composed of 832 symbols, of which the first four symbols are the “segment sync” symbols (5, –5, –5, 5) and the remaining 828 symbols are Reed-Solomon-encoded, interleaved, and trellis-encoded symbols drawn from the 8 level pulse amplitude modulation constellation (±1, ±3, ±5, ±7) [9]. The field sync segment is used for a training sequence of the equalizer. The equalizer has been implemented with a decision-feedback equalizer (DFE). The DFE can be adapted by using the least-mean square (LMS) algorithm or the recursive-least square (RLS) algorithm in the training mode and using one of blind algorithms, such as the constant modulus algorithm (CMA), the stop-and-go (SAG) algorithm, or the SAG dual-mode CMA for the data segments [3]. However, the DFE with these adaptation algorithms has not dealt with severe channels, such as indoor, portable, and distributed transmission reception, where a near 0 dB ghost exists.

To handle such severe channels in the DFE, a channel-matched filter (CMF) is used for making the property of the channels more manageable [4], [6].

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2.1 Channel-Matched Filter

The channel-matched filter is obtained from the estimated channel \[4\]. Let \( \hat{H}_C(z) \) be an estimated channel impulse response with a length of \( N \), i.e.,

\[
\hat{H}_C(z) = \sum_{i=0}^{N-1} \hat{h}_i z^{-i}, \quad (1)
\]

Then, we can obtain the matched filter, or mirror filter of the above channel as

\[
H_M(z) = \sum_{i=0}^{N-1} \hat{h}^*_{N-i} z^{-i}, \quad (2)
\]

where the superscript \(*\) denotes complex-conjugation. The CMF changes the channel property and mitigates the incoming signals to maximize the output SNR. Figure 2 is an example showing the effect of the CMF. In this figure, \( T \) is the symbol duration of 0.093 \( \mu s \) corresponding to the ATSC DTV standard symbol rate of 10.76 MHz [9]. The combined response becomes a changed channel that the DFE has to equalize. It can be seen that the amplitude of the highest level (main path) is enhanced by about 10 dB. Actually, there is no “main path” in Fig. 2 plot (1), since there are two almost equal amplitude paths (0 dB echo). The CMF, Fig. 2 plot (3), changes the given severe channel comprised of high-level ghosts into the more manageable channel shown in Fig. 2 plot (4).

However, if the channel estimation is not precisely performed, the CMF may deteriorate the channel property and decrease the output SNR. Therefore, it is important to obtain the precise channel information.

2.2 Channel Estimation

We consider channel estimation with a finite-impulse response (FIR) filter as shown in Fig. 3 [6]. The training sequence of a field sync segment is used for channel estimation. As the length of the training sequence is not more than 832 symbols, fast algorithms are required to converge within the short duration. To achieve this, the RLS algorithm is appropriate and the adaptation in the channel estimation scheme is given by

\[
k[k+1] = \frac{\lambda^{-1} P[k] d[k]}{1 + \lambda^{-1} d^T[k] P[k] d[k]} \quad (3)
\]

\[
\hat{h}[k+1] = \hat{h}[k] - \lambda^{-1} k[k+1] e[k] \quad (4)
\]

\[
P[k+1] = \lambda^{-1} P[k] - \lambda^{-1} k[k+1] d^T[k] P[k], \quad (5)
\]

where \( \lambda \) is the forgetting factor and

\[
e[k] = r[k] - \hat{H}_C(z) d[k]
\]

\[
d[k] = \begin{bmatrix} d[k] d[k-1] \cdots d[k-N+1] \end{bmatrix}^T
\]

\[
\hat{h}[k] = \begin{bmatrix} \hat{h}_0[k] \hat{h}_1[k] \cdots \hat{h}_{N-1}[k] \end{bmatrix}^T.
\]

Here \( r[k] \) is the channel output at time \( k \), or the desired signal of the channel estimator and \( d[k] \) is the transmitted signal known as the training sequence, or the input of the channel estimator.

It is well known that the FIR channel estimation
scheme with the RLS works well in the sufficient-order case where the number of taps of the estimator is greater than or equal to the order of a channel to be estimated and under a causal channel which does not have pre-ghosts. In practical situations, however, there are many cases in which the order of a channel is unknown. In addition, communication channels often possess a non-causal property that both pre-ghosts and post-ghosts simultaneously exist. Therefore, it is necessary to analyze the estimation results performed in diverse situations in order to appropriately apply them to channel equalization.

3. Analysis of Adaptive Channel Estimation

This section deals with the estimation properties of reduced-order cases and low SNR cases. If the chosen order of the adaptive filter is inferior to that of an unknown channel, the channel estimation using adaptive FIR filtering results in reduced-order approximation.

3.1 Properties of the Training Sequence

Since some properties of the channel estimation may depend on the statistics of the input signal to the adaptive filter, it is needed to overview the training sequence of the ATSC DTV system. The training sequence is comprised of pseudo-random noise (PN) sequences and reserved bits as shown in Fig. 4 [9]. Note that PN63 is repeated three times and the middle PN63 is inverted on every other data field sync. Due to these repetitions and the length of not being $2^n$, the training sequence does not possess an optimal correlation property of m-sequence.

Despite the non-optimality, however, it is possible to approximate that the sequence is white by trying to compute its auto-correlation, which is shown in Fig. 5. The ratio of the second largest and the largest magnitudes is −17.8 dB. And the other magnitudes are less than −22.0 dB. Under that condition of the training sequence, the estimation properties approach to the ones derived from the white input sequence within the approximation-induced error of about less than −20 dB. The simulation results presented in Section 4 will verify the approximation of the whiteness of the training sequence.

3.2 Reduced-Order Estimation

Some properties of the reduced-order estimation using the adaptive infinite-impulse response (IIR) filter were found in [7] and [8]. Since the FIR filter can be considered as a special case of the IIR filter, having no feedback section, or denominator filter, we may infer the properties of the reduced-order FIR channel estimation from those of the reduced-order IIR estimation.

Let $G(z) = \sum_{i=0}^{\infty} g_i z^{-i}$ be an unknown IIR system and $\hat{G}(z)$ be an adaptive IIR filter for estimating the unknown system. This is approached by constructing an adjustable rational model:

$$\hat{G}(z) = \sum_{i=0}^{\infty} \hat{g}_i z^{-i}. \quad (7)$$

One of the reduced-order estimation properties presented in [7] and [8] says that in case the input is white, the estimated coefficients become equal to the system parameters within the number of taps of the numerator filter $B(z)$, that is,

$$\hat{g}_i = g_i, \quad i = 0, 1, \ldots, N. \quad (8)$$

To consider a non-causal system with $M$ anti-causal coefficients, we have to change $G(z)$ into $z^M G(z)$. Then, we may decompose the non-causal system into the anti-causal and causal parts:

$$z^M G(z) = \sum_{i=0}^{\infty} g_i z^{-(i-M)} = \sum_{i=0}^{M-1} g_i z^{-(i-M)} + \sum_{j=0}^{\infty} g_{j+M} z^{-j} \quad (9)$$

\[Fig. 4\] VSB data field sync (or the training sequence) [9].

\[Fig. 5\] Auto-correlation of the training sequence.
Even in this non-causal case, since the adaptive IIR filter still remains to be causal, the anti-causal part of the unknown system cannot be estimated. Then, we may know that the adaptive filter estimates only the causal part. Therefore, from (8) and (9) we have
\[
\tilde{g}_i = g_{i+M}, \quad i = 0, 1, \cdots, N. \tag{10}
\]

If an unknown channel \( H(z) \) has a longest-delayed pre-ghost of \( M \)-symbol time and a longest-delayed post-ghost of \( K \)-symbol time, then \( H(z) \) is given by
\[
H(z) = \sum_{i=-M}^{K} h_i z^{-i}. \tag{11}
\]

In the reduced-order FIR case \((N < K \text{ and } L = 0)\), from (10) we may find the adaptive FIR filter \( \hat{H}_c(z) \) given in (1), which is equal to the IIR filter \( \tilde{G}(z) \) with \( A(z) = 1 \), has the following estimation property
\[
\hat{h}_i = h_i, \quad i = 0, 1, \cdots, N. \tag{12}
\]

This result means that in adaptive FIR channel estimation, the tap coefficients of the estimator converge to their corresponding channel coefficients irrespective of the number of taps of the estimator and the causality of an unknown channel.

3.3 Noise Effect

Adaptive system estimation may be affected by noise, in which case the estimator is called a biased estimator. To check the unbiasedness of the estimator, we can use the fact that the statistical average of the update term is zero at the convergent point. Applying this fact to (4), we obtain
\[
E\{k[k + 1]e[k]\} = 0. \tag{13}
\]

For the value of \( \lambda \) close to 1, \( P[k] \) varies slowly, so we can assume that \( P[k] \) is uncorrelated with \( d[k] \), which enables (13) to be rewritten as
\[
E\{d[k]e[k]\} = E\{d[k]\} \{r[k] - \hat{h}^T[k]d[k]\} = 0. \tag{14}
\]

Note that \( r[k] \) is the channel output which is given by
\[
r[k] = h^T d[k] + w[k], \tag{15}
\]

where \( h \) is a channel-coefficient vector and \( w[k] \) is zero-mean uncorrelated additive noise. Using (15) and rearranging all the terms, (14) becomes
\[
E\{d[k]d^T[k]\} h - E\{d[k]d^T[k]\} E\{\hat{h}[k]\} + E\{d[k]w[k]\} = 0. \tag{16}
\]

Since \( d[k] \) and \( w[k] \) are uncorrelated, the last term of (16) becomes a zero vector, i.e., \( E\{d[k]w[k]\} = 0 \). Therefore, we get to the final result
\[
E\{d[k]d^T[k]\} E\{h - \hat{h}[k]\} = 0. \tag{17}
\]

From (17) we can know that \( E\{h - \hat{h}[k]\} \) is zero if the inverse matrix of \( E\{d[k]d^T[k]\} \) exists. This means that if the transmitted symbols \( d[k] \) are persistently exciting, the channel estimator becomes an unbiased estimator.

4. Simulation Results

We performed computer simulations to verify the analysis results on the channel-estimation properties in ATSC DTV receivers. The channel profiles used for the simulations were the Brazil channel D which is the indoor channel used for the Laboratory Test in Brazil [10] and the Advanced Television Technology Center (ATTC) channel D which is the outdoor channel including anti-causal part.[11]. The channel information are given in Tables 1 and 2, respectively. The received SNR was obtained from the baseband equivalent VSB channel model presented in [12]. Note that the SNR includes echo power for the signal power while carrier-to-noise ratio (CNR) refers only to the main signal power, not including echo power. In our simulations based on the VSB channel model, we considered VSB modulation and passband-related effects such as phase information and a carrier frequency under Korean DTV CH 14 for which the center frequency is 473 MHz. The RLS algorithm with the forgetting factor of \( \lambda = 1.0 \) was used for all channel estimations.

To verify the reduced-order channel estimation property, we used the Brazil channel D under the SNR of 20 dB which corresponds to the CNR of 12.7 dB. More than 80 taps are required to perform the estimation in the sufficient-order condition. Estimation with an adaptive filter of less than 80 taps corresponds to the reduced-order estimation. Figure 6 shows that the tap coefficients of the estimator converge to their corresponding channel coefficients irrespective of the number of taps of the estimator.

Under low SNR conditions, the channel estimation property is shown in Figs. 7 and 8, the SNRs of which are 10 dB and 0 dB, respectively. In terms of CNR, those SNR values correspond to 2.7 dB and –7.3 dB. Comparing the estimation results with the same taps in Figs. 7 and 8, we can find that the estimated coefficients were not affected by noise. In addition, like under high SNR conditions such as the results in Fig. 6, the reduced-order channel estimation property still holds even in low SNR values.

Figure 9 verifies the non-causal channel estimation property. For this simulation, the ATTC channel D was used because it has both pre- and post-ghosts. For this channel, the SNR was 20 dB and the CNR was 19.5 dB. In spite of the existence of the anti-causal part generated by the pre-ghosts, the channel estimator matches well the causal part within its tap limit.

<table>
<thead>
<tr>
<th>Table 1</th>
<th>Multi-path profile: Brazil Channel D.</th>
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<tbody>
<tr>
<td>Delay (µs)</td>
<td>0.0</td>
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<tr>
<td>Amp. (dB)</td>
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</table>

<table>
<thead>
<tr>
<th>Table 2</th>
<th>Multi-path profile: ATTC Channel D.</th>
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<tr>
<td>Delay (µs)</td>
<td>–1.75</td>
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<tr>
<td>Amp. (dB)</td>
<td>–20</td>
</tr>
<tr>
<td>Phase (deg.)</td>
<td>45</td>
</tr>
</tbody>
</table>
Fig. 6 Reduced-order channel estimation (SNR=20.0 dB, CNR=12.7 dB). (a) Original channel. (b) Estimated channel (80 taps). (c) Estimated channel (60 taps). (d) Estimated channel (30 taps).

Fig. 7 Reduced-order channel estimation (SNR=10.0 dB, CNR=2.7 dB). (a) Estimated channel (80 taps). (b) Estimated channel (60 taps). (c) Estimated channel (30 taps). (d) Estimated channel (20 taps).
Fig. 8  Reduced-order channel estimation (SNR=0 dB, CNR=−7.3 dB). (a) Estimated channel (80 taps). (b) Estimated channel (60 taps). (c) Estimated channel (30 taps). (d) Estimated channel (20 taps).

Fig. 9  Estimation of a non-causal channel (SNR=20.0 dB, CNR=19.5 dB). (a) Original channel. (b) Estimated channel (220 taps). (c) Estimated channel (100 taps). (d) Estimated channel (50 taps).
5. Conclusions

Channel estimation are very useful for the DFE in ATSC DTV receivers, especially, under severe channels, such as a channel with 0 dB ghost. We analyzed the estimation properties to go over the applicability of the estimation to the DFE in real fields. For such situations, reduced-order cases and low SNR cases were considered. Through simple mathematical derivation, we showed that the FIR channel estimator may match an unknown channel within its tap limit irrespective of the number of taps and the non-causality of the unknown channel. In addition, the estimator possesses an unbiased property. Each case was checked by computer simulations. Our analysis results are expected to provide much information in the field of enhancement of the equalization performance in ATSC DTV receivers.

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