A Pilot Aided Averaging Channel Estimator for DVB-T2

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Abstract—In DVB-T2 broadcasting applications, errors occur due to channel imperfections, reducing the system’s throughput. The use of a channel estimator and equalizer counteracts the channel induced distortions and thus improves the system’s Bit Error Rate (BER). In our study, a simple-to-implement channel estimator is proposed. It is based on averaging the sub-channel frequency responses of successive OFDM symbols in time, considering a very slow time-varying channel. In the frequency domain, spline interpolation is applied. The proposed estimator can be used in practical receivers with limited computation power and low memory capacity.

Index Terms—Channel estimation, DVB-T2, OFDM, scattered pilots

I. INTRODUCTION

D IGITAL Video Broadcasting Second Generation Terrestrial (DVB-T2) [1] was published in 2009 as an improvement to DVB-T which was published in 1997, and now is used in most countries for Digital Terrestrial Television (DTTV or DTT). DVB-T2 adopts many high-end technologies from DVB-T and Digital Video Broadcasting Second Generation Satellite (DVB-S2) [2]. The new standard is very flexible and thus it uses new additional technologies, such as Multiple Physical Layer Pipes (MPLPs), Alamouti coding, Constellation Rotation (CR), Extended Interleaving and Future Extension Frames (FEF). The combinations of QAM order, Guard Interval (GI), Forward Error Correction (FEC), and high FFT sizes up to 32k helps the system to achieve high bit rates up to 45.5Mbps [3]. These high bit rates are suitable for transmitting High Definition TV (HDTV) content.

Radio channel imperfections, such as attenuation, phase shifting and time delays, result in errors, reducing in this way the system’s throughput. In addition, due to multipath propagation, the channel suffers from distortion both in time and frequency domain causing in this way Inter-Symbol Interference (ISI). The use of a channel estimator and equalizer counteracts the channel’s induced distortions and thus improves the Bit Error Rate (BER).

The topic of channel estimation and equalization has thoroughly been studied in past years because of its importance. Several methods have been proposed to compensate the channel distortion. In the DVB-T2 case, pilots are used for frame, frequency and time synchronization, channel estimation, phase noise tracking and identification of the transmission mode [1].

The performance of a conventional Least Square (LS) estimator and a Least Mean Square (LMS) estimator is investigated in [4] for different pilot arrangements. In [5], an analysis of the effects of time-based pilot interpolation over time-varying channels is presented. Using different pilot patterns, a DVB-T2 channel estimator and an equalizer were modeled and evaluated in terms of performance in [6] and [7]. Also, an in-depth study of minimizing the number of the used pilots in terms of error probability can be found in [8]. A comparative study of channel estimation methods is provided in [9], where a LS estimator seems to be computationally superior compared to a Minimum Mean Square Error (MMSE) estimator. In [10], a blind channel estimation method is proposed where no pilots are used and thus the throughput is maximized.

In the method proposed here, a simple-to-implement channel estimator is proposed. It is based on averaging the channel frequency responses of successive OFDM symbols assuming a slow time-varying channel. In this study, as the channel varies slowly in time, the estimator treats the channel as constant in the time domain for the interval of the buffered OFDM symbols. In the frequency domain, the channel is assumed as selective. For the simulation needs, Rayleigh and Ricean models have been used. The proposed estimator can be used in practice by receivers that have limited computation power and reduced memory (RAM) capacity.

The rest of the paper is organized as follows: In Section II the architecture of DVB-T2 is described in brief. In Section III the Scattered Pilots (SP) used in DVB-T2 are discussed in detail. In Section IV the proposed method is explained and analyzed in depth. The simulation results for different channel environments as well as for different configurations of SP patterns, FFT sizes, QAM orders and mobile speeds are derived and analyzed in Section V. Finally, in Section VI the effectiveness of the averaging channel estimator is discussed and improvements of the estimator for fast time-varying
channels are proposed.

II. ARCHITECTURE OF DVB-T2 SYSTEM

The flowchart of a simplified DVB-T2 system is depicted in Fig.1. The Mode and Stream Adaptation (MSA) block is responsible to form the baseband frame (BBFRAME). The BBFRAME is inserted into Bit Interleaved Coding and Modulation (BICM) block. The outer coder (BCH) and the inner coder (LDPC) offer coding rates of 1/2, 3/5, 2/3, 3/4, 4/5, 5/6 and construct the forward error correction frame (FECFRAME). After a bitwise interleaving, each FECFRAME is de-multiplexed into parallel cell words and mapped into constellation values of Quadrature Amplitude Modulation (QAM). The QAM available orders are QPSK, 16QAM, 64QAM, and 256QAM. For robustness improvement, constellation rotation is optionally provided. These data cells are further interleaved in time to ensure uncorrelated interference and distortion along each FECFRAME.

The data cells of the FECFRAME are then inserted into the Frame Builder and OFDM generator. The Frame Builder constructs the T2-Frame by assembling the data cells into the P2, the Normal and the Frame closing symbols.

Table I depicts the type of pilots in each symbol. In this paper, without loss of generality, we focus only in normal OFDM symbols and therefore only SP are considered.

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Pilot Type</th>
<th>Scattered</th>
<th>Continual</th>
<th>Edge</th>
<th>P2</th>
<th>Frame Closing</th>
</tr>
</thead>
<tbody>
<tr>
<td>P1</td>
<td>√</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>P2</td>
<td>√</td>
<td>√</td>
<td>√</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Normal</td>
<td>√</td>
<td>√</td>
<td>√</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Frame Closing</td>
<td>√</td>
<td></td>
<td></td>
<td></td>
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<td></td>
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</tbody>
</table>

The IFFT block is responsible to transform the data and the pilot cell information into an equivalent signal in time domain. The subcarriers of each frequency cell are orthogonal to each other to prevent Inter Carrier Interference (ICI). Equation (1) describes mathematically the OFDM system.

\[ s(t) = \sum_{g=-\infty}^{\infty} \sum_{k=0}^{N-1} c_{n,k} \cdot g_k(t-nT_s) \]  

with

\[ g_k(t) = \begin{cases} e^{j2\pi f_k t}, & t \in [0, T_s] \\ 0, & \text{elsewhere} \end{cases} \]

and

\[ f_k = f_0 + \frac{k}{T_s}, \quad k = 0, \ldots, N-1 \]

where \( c_{n,k} \) is the symbol of the \( k \)-th subcarrier, \( N \) is the number of total subcarriers, \( f_k \) is frequency of the \( k \)-th subcarrier and \( f_0 \) is the lowest frequency used. The available numbers of subcarriers are: 1k, 2k, 4k, 8k, 16k and 32k.

A Guard Interval (GI) is made up as a prefix of a cyclic continuation of the useful part of the OFDM symbol used to prevent ISI. The available GI values are: 1/4, 19/128, 1/8, 19/256, 1/16, 1/32, and 1/128.

III. SCATTERED PILOTS IN DVB-T2

DVB-T2 uses 8 different SP patterns, named PP1 to PP8, to compensate for the variation of the channel in the time and in the frequency domain. The position of the pilots onto a subcarrier in the OFDM symbol satisfies the following condition:

\[ k \cdot \text{mod}(D_x \cdot D_y) = D_x \cdot (\ell \mod D_y) \]  

(2)

where \( D_x \) defines the separation of pilots bearing carriers in

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Fig. 1. DVB-T2 system.
each OFDM symbol, $D_x$ defines the number of OFDM symbols forming one SP sequence, $k \in [K_{\text{min}}, K_{\text{max}}]$ is the index of subcarrier into the OFDM symbol, and $\ell \in [1, S]$ is the index of the OFDM symbol into the T2 frame that contains S symbols in total.

The available values of $D_x$ and $D_y$ given in Table II theoretically support fluctuations in time and frequency up to the Nyquist limit. Thus, the spacing in time and frequency domain should not exceed the respective limit given below:

$$N_T \leq \frac{1}{2 f_D T}, \quad N_f \leq \frac{1}{2 \Delta f \tau_{\text{max}}}$$

(3)

where $f_D$ is the Doppler frequency, $\Delta f$ is the spacing between subcarriers, $T$ is the duration of the OFDM symbol, and $\tau_{\text{max}}$ is the multipath delay. The maximum GI fraction should never exceed $1/D_x$. From (3) the Doppler limit is proportional to $1/D_y$. The capacity of transmission, neglecting all other types of pilots (cf. Table I), is a fraction of $1/D_x D_y$. From Table II, the overhead is derived to be equal to 8.33% for PP1 and 1.04% for PP7 and PP8. It is obvious that as the overhead increases, higher values are obtained for the Nyquist limit. With $D_y = 2$, PP2, PP4 and PP6 provide higher Nyquist limit for Doppler speed, [12].

<table>
<thead>
<tr>
<th>Pilot Pattern</th>
<th>$D_x$</th>
<th>$D_y$</th>
</tr>
</thead>
<tbody>
<tr>
<td>PP1</td>
<td>3</td>
<td>4</td>
</tr>
<tr>
<td>PP2</td>
<td>6</td>
<td>2</td>
</tr>
<tr>
<td>PP3</td>
<td>6</td>
<td>4</td>
</tr>
<tr>
<td>PP4</td>
<td>1</td>
<td>2</td>
</tr>
<tr>
<td>PP5</td>
<td>1</td>
<td>4</td>
</tr>
<tr>
<td>PP6</td>
<td>2</td>
<td>2</td>
</tr>
<tr>
<td>PP7</td>
<td>2</td>
<td>4</td>
</tr>
<tr>
<td>PP8</td>
<td>6</td>
<td>1</td>
</tr>
</tbody>
</table>

To reduce the noise on channel estimation, the pilots are boosted considering that all the symbols have approximately the same power. Table III shows the amplitudes of each SP pattern.

<table>
<thead>
<tr>
<th>SP pattern</th>
<th>Amplitude</th>
<th>Equivalent Boost in dB</th>
</tr>
</thead>
<tbody>
<tr>
<td>PP1,PP2</td>
<td>4/3</td>
<td>2.5</td>
</tr>
<tr>
<td>PP3,PP4</td>
<td>7/4</td>
<td>4.9</td>
</tr>
<tr>
<td>PP5,PP6,PP7,PP8</td>
<td>7/3</td>
<td>7.4</td>
</tr>
</tbody>
</table>

The modulations of the pilots depend on the carrier index and the symbol number. The real and imaginary parts are given below:

$$\text{Re}\{c_{m,\ell,k}\} = 2 A_{\text{sp}} \left( 1 - r_{\ell,k} \right)$$

$$\text{Im}\{c_{m,\ell,k}\} = 0$$

(4)

where $A_{\text{sp}}$ is the SP amplitude in Table III, $r_{\ell,k}$ is the reference sequence generated by XOR-ing a Pseudo Random Binary Sequence (PRBS) with a Pseudo-Number, $m$ is the index of the T2 frame, $k$ is the frequency index of the carriers, and $\ell$ is the index of the OFDM symbol. Finally, the arrangement for two Pilot Patterns (PP) in SISO mode for DVB-T2 including Edge pilots are shown in Fig. 2 (a) for PP1 and (b) for PP6.

![Fig. 2. SP patterns in SISO mode.](image)

### IV. CHANNEL ESTIMATION

After serial to parallel transformation and pilot insertion depending on specific SP pattern, the data sequence $\{X(k)\}$ is transformed from a frequency domain into a time domain signal $\{x(n)\}$ by the IFFT block. Thus, the transmitted signal is transformed by the following expression:

$$x(n) = \text{IFFT}\{X(k)\} = \sum_{k=0}^{N-1} X(k)e^{\frac{-2\pi nk}{N}}$$

(5)

Thus, the received signal is as described below:

$$y(n) = \sum_{\ell=0}^{L-1} h(n, \ell) \cdot x(n-\ell) + w(n)$$

(6)

where $0 \leq n \leq N-1$, $L$ is the number of multipath versions of the original signal $x(n)$, $h(n, \ell)$ is the channel impulse response of the $n$-th OFDM symbol from the $\ell$-th path and $w(n)$ is the Additive White Gaussian Noise (AWGN) with zero mean and variance $\sigma^2$. At the receiver side, after cyclic prefix (CP) removal and FFT, it is derived that:

$$Y(k,n) = \text{FFT}\{y(n)\} = X(k,n) \cdot H(k,n) + I(k,n) + W(k,n)$$

(7)
where: $H(k,n)$ is the channel frequency response, $W(k,n)$ is the noise signal in the frequency domain, and $I(k,n)$ denotes the introduced Inter Carrier Interference (ICI) caused by the time-varying channel. In this study, a very slowly time-varying channel is assumed and thus the $I(k,n)$ term can be neglected.

If $P$ is the vector of the position of the pilots in the OFDM symbol, then the estimated channel frequency response $\hat{H}_e(k)$, $k \in P = \{p_0, p_1, \ldots, p_P\}$ undergoes Spline interpolation in order to obtain the channel response $H_e(k)$, $k \in [0,k_{\text{max}}]$ for every carrier of the OFDM symbol.

Finally the channel equalization will be done using the Zero Force (ZF) equalizer or the Least Square Error (LSE) equalizer, described respectively by the following two expressions:

$$X_e = \frac{Y(k)}{H_e(k)}, \quad k = 0,1,\ldots,K_{\text{max}}$$

$$X_e = \frac{H^*}{H_e \cdot H_e} Y_e$$

where $(\cdot)^*$ denotes the conjugate of a complex number or a complex matrix.

V. PROPOSED ESTIMATOR

Assuming a slowly time-varying channel, the estimation of $H_e^P(k)$ can be calculated by simply averaging the values of the last $B$ received OFDM symbols. The value of $B$ is derived as follows: Let $T_B$ be the time interval between the last received OFDM symbol $S$ and the $S-(B+1)$ OFDM symbol. Then:

$$T_B = B \cdot T_u$$

where $T_u$ is the duration of an OFDM symbol, [1]. In order to ensure that $T_B < T_c$, where $T_c$ is the coherence time, we set:

$$T_B = \frac{1}{50} T_c$$

The coherence time can be set equal to $T_c = 0.5/f_d$, where $f_d$ is the Doppler spectrum, [13], and is extracted by the expression:

$$f_d = f_c \frac{v}{c}$$

where $c$ is the speed of light, $f_c = 800\text{MHz}$ is the upper limit of the carrier frequency of the analog transmitted signal in DTV, and $v(m/s)$ the mobile speed. Combining (10), (11) and (12), it is derived that:

$$B = \text{floor} \left( \frac{c}{100 v f_c T_u} \right)$$

If $B<1$ in (13), then we set $B = 1$ and actually no averaging is performed. That is why the proposed algorithm is suitable for low mobile speeds or equivalently for large coherence time $T_c$ where the channel varies very slowly in time. However, the performance of the proposed estimator is tested also for fixed values of buffer size, in order to clearly demonstrate its capabilities and limitations.

The averaging estimator operates for the last $B$ received OFDM symbols as follows. First, the symbols are stored in a matrix $A$ of size $B \times N$. Then a vector $\text{avPilots}$ is constructed as given below:

$$\text{avPilots}(n) = \frac{1}{P_n} \sum_{i=1}^{B} A(i,n)$$

where $P_n$ is the number of pilots in the $n$th subcarrier. It must be noted that if the $n$-th subcarrier does not carry pilots, then $P_n = 1$. The averaging process is depicted in Fig. 3.

After the construction of the vector $\text{avPilots}$, with spline interpolation, the channel frequency response $\hat{H}(n)$ is calculated and by using a LSE estimator the received data symbols are equalized.

It is important to note that the proposed estimator can be used in conjunction with more sophisticated estimators, such as MMSE and Kalman filters, where the knowledge of the channel is an important factor for their good performance. As the averaging estimator rejects the noise, (8) can be rewritten as:

$$H = \frac{Y(k)}{X(k)} + W(k)$$

Since $H_{\text{av}} = \bar{Y}(k)/\bar{X}(k)$, the previous equation can be written in following form:

$$\hat{H} = \hat{H}_{\text{av}} + \hat{W}$$
where $\hat{H}_{av}$ is the averaged estimation of the channel, $\hat{H}$ is the conventional channel estimation and $\hat{W}$ is the noise. So, this could be used for noise estimation and then to pass this information to a more sophisticated estimator.

VI. SIMULATION AND RESULTS

In this section, the performance of the averaging estimator is tested under various configurations of QAM order, FFT size, channel model, speed of the receiver, and size of the buffer.

For convenience in the comparison of the different configurations, the bandwidth of the RF signal is set equal to 8 MHz, which is the usual bandwidth in DVB-T. In order to use one of the highest possible frequencies, which suffer the most of the Doppler effect, the central carrier frequency is set to be equal to 800 MHz. Two fading channel models are used, the Rayleigh and Ricean ones, with 6 taps each. No source or channel encoding is used and none of the available interleaving schemes are used, in order to focus only into the improvement offered by the estimator.

The first simulation tests the performance of the estimator in a pure AWGN channel with no multipath components. A 4-QAM modulation is used (QPSK), the FFT size is 4k and the buffer size $B$ is set equal to 20. Fig. 4 depicts the performance of the proposed and the conventional estimator. It is clear that the proposed estimator overcomes the conventional one by about 3.5 dB for the same BER.

More realistic configurations are used below. The first one is displayed in Table IV. The buffer size calculation is based on (13). Fig. 5 depicts the improvement because of using the averaging estimator which effectively reduces the noise.

The same configuration as in Table IV is used below, except the mobile speed which is set equal to 3 km/h. The buffer size is equal to 16 which is a suitable value. Thus the performance is similar to the previous one and the averaging estimator still works acceptably. The results are depicted in Fig. 6.

Finally, the mobile speed is set equal to 10 km/h. According to (13), the buffer becomes equal to 4 and thus degradation is expected. The curves in Fig. 7 evidently show the predicted degradation. For SNR $< 20$ dB no improvement occurs. For higher SNR values, the estimator works again as in the previous simulations. Further increase in the mobile speed leads to $B = 1$ and thus no averaging is performed.

<table>
<thead>
<tr>
<th>TABLE IV CONFIGURATION OF SIMULATION</th>
</tr>
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<tbody>
<tr>
<td>Radio Environment: Rural Area</td>
</tr>
<tr>
<td>Radio Channel Type: Rayleigh</td>
</tr>
<tr>
<td>QAM-Order: 4</td>
</tr>
<tr>
<td>FFT size: 4k</td>
</tr>
<tr>
<td>Speed of Mobile (Km/h): 1</td>
</tr>
<tr>
<td>Buffer Size: 49</td>
</tr>
</tbody>
</table>

Fig. 4. Comparison between the proposed and the conventional estimator in AWGN channel.

Fig. 5. Comparison between the proposed and the conventional estimator with speed equal to 1 Km/h.

Fig. 6. Comparison between the proposed and the conventional estimator with speed equal to 3Km/h.

Fig. 7. Comparison between the proposed and the conventional estimator with speed equal to 10Km/h.
Fig. 7. Comparison between the proposed and the conventional estimator with speed equal to 10 Km/h.

Then, the performance of the proposed estimator is examined for different QAM orders. The configuration of Table IV is used again and only the QAM order is changed to 16-QAM. The proposed estimator again succeeds to improve the reception up to 3dB as shown in Fig. 8. Although the estimator performs better than the non-averaging one for higher QAM orders with improvement up to 6 dB, as depicted in Fig. 9, the overall performance is not acceptable.

Next, the FFT size is tested. The configuration of Table IV is used again but the FFT is set equal to 1k, so the buffer size becomes $B = 196$. The curves of BER vs. SNR for the two investigated estimators are depicted in Fig. 10. The superiority of the averaging estimator is obvious. The estimators are also examined for FFT = 8k which corresponds to $B = 24$. Fig. 11 illustrates the performance of the estimators and shows that while the conventional one converges to a plateau, the averaging estimator leads to lower values of BER.

It is interesting to see how the proposed estimator behaves for another type of fading channel model. Therefore, the configuration of Table IV is used again except the radio channel type which is considered to be Ricean with certain values of the radio channel factor $K$ (i.e., the ratio between the power in the line of sight and the power of the multipath components). For $K=10^5$ (very strong line of sight component) the averaging estimator outperforms the conventional one, as shown in Fig. 12. For $K=10$ (weak line of sight component) the performance of the averaging estimator is even better, as shown in Fig. 13.
Finally, the proposed estimator is tested for different SP patterns. The same configuration of Table IV is used. The PP3 and PP7 are tested and the results are depicted respectively in Figs. 14 and 15. It is clear that the proposed estimator performs acceptably even in cases where the available pilots are fewer than PP1 such as in PP3. The conventional estimator results in a flat BER curve for SNR > 18dB due to noise enhancement. The averaging estimator succeeds to discard the noise and thus it achieves lower values of BER. In the case of PP7, the pilot arrangement is sparser (see Fig. 2b) and the conventional estimator performs even worse. The averaging estimator is also affected but still performs better than the conventional one.

VII. CONCLUSION

The performance of an averaging estimator is thoroughly investigated. Initially, the estimator is tested in a frequency flat channel suffering only of AWGN and the performance of the estimator is found to be 3.5 dB better than a non-averaging one. The proposed estimator is also tested in frequency selective channels and for different sizes of FFT, QAM order, radio channels, mobile speeds, and pilot patterns. In all cases, for very low mobile speeds, the proposed estimator gives better results than the conventional one. As expected, the proposed estimator fails for higher mobile speeds due to the Doppler Effect. As the speed increases the channel response becomes time-varying and the coherence time gets smaller. So, fewer OFDM symbols can be buffered and thus the averaging procedure is less effective. The proposed estimator can also be used for estimation of the noise variance. This information can be utilized by more accurate and complicated estimators based on channel statistics in order to make a more accurate channel estimation.

REFERENCES


