REDUCED COMPLEXITY DOPPLER COMPENSATION FOR MOBILE DVB-T

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Abstract - The reception of the DVB-T signal on a mobile equipment is strongly affected by Doppler spread. Most of the current proposals require the use of two or more antennas at the receiver and a computationally expensive signal processing. In this paper for a DVB-T receiver with a single antenna we propose new schemes based on the interference cancellation principle. With respect to previously studied techniques, our solution is able to deliver DVB-T services at higher speeds, while a more flexible architecture allows a wider range of trade-off between complexity and performance. In order to improve the performance in terms of maximum achievable speed, an iterative scheme is proposed which performs both the interference cancellation and the channel estimation. A decoding and re-encoding of the inner convolutional code is inserted into the iteration process in order to boost the performance. Moreover, by applying the interference cancellation principle also for the estimation of the channel parameters, we obtain a scheme with a reasonable complexity for the DVB system. Simulations results for various DVB-T modes, constellations and code rates show that the proposed schemes allow the correct reception of DVB-T for moderate and high speeds.

Keywords - Doppler, DVB, interchannel interferece.

I. INTRODUCTION

The Digital Video Broadcasting (DVB) system has been proposed as a standard for the broadcast of video digital content through terrestrial wireless channels (DVB-T) [1]. Recently, a growing interest has been focused on the use of the DVB-T for mobile applications, ranging from portable devices with slow mobility to on-board equipment for cars. The use of the existing DVB-T standard in a mobile environment poses a number of challenges. In fact, since the transmission is based on the Orthogonal Frequency Division Multiplexing (OFDM), the symbol duration is long with respect to the coherence time of the used channel and hence the orthogonality among the subcarriers is disrupted. As a consequence, data on one subcarrier leak partially on other subcarriers thus yielding inter carrier interference (ICI). Also a more elaborate channel estimator is needed in order to track properly the channel variations inside and among OFDM symbols.

Various solutions to these problems have been proposed, i.e., longer time interpolators for a better estimate of the channel [2] and multiple antennas at the receiver yield a diversity gain [3]. However, both solutions have important drawbacks, since the interpolators require more silicon area for memory and the multiple antennas have an additional cost in extra equipment and deployment. On the other hand, current solutions based on one antenna are exceedingly complex for the DVB-T. In [4] a linear equalization scheme was proposed, which requires the adaptive inversion of many matrices. In order to reduce the complexity ICI cancellation schemes have been proposed [5,6], where data are first decided as for a conventional receiver and the decisions are used to remove interference. Lastly, another decision is taken on this "clear" data. Since the first decision is significantly affected by the Doppler interference, the performance of this ICI cancellation scheme only guarantees the required Quality of Service (QoS) for low speeds. In this paper we propose new schemes for both the interference cancellation and the channel parameter estimation. The basic idea is an iterative application of the interference cancellation principle that allows to increase progressively the reliability of the decided data. Moreover, a further increase of performance is also obtained by including in the iterative process the decoding and re-encoding of the inner convolutional code. In order to reduce the complexity we then derive an efficient scheme based on filters with a reduced numbers of taps. We also propose a decision-directed channel estimation scheme that successively estimates subset of the channel parameters. At each iteration a fixed filtering is applied to the received block and then a standard channel estimator for OFDM is used. Even if sub optimum with respect to previously studied schemes, this estimator has a significantly reduced complexity and a variety of well-known OFDM channel estimator may be included into this scheme, thus allowing a wide range of tradeoff between complexity and performance.

Simulations results are presented for various modes, constellation sizes and code rates of the DVB-T standard. By comparing the performance of the reduced complexity schemes with existing multiple antenna solution, we conclude that the new techniques allow the use of DVB-T for similar or higher speeds range with a limited complexity.

II. SYSTEM DESCRIPTION

The transmitted signal of DVB-T undergoes multipath propagation and, when mobility is considered, the Doppler spread phenomenon yields a time-variation of the channel. OFDM modulation is adopted for the transmission and according to the network deployment and the environment characteristics

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the broadcaster chooses the transmission mode and the number of subcarriers N: We assume that the guard interval is longer than the duration of the channel impulse response so that the received vector signal is not affected by intersymbol interference. After filtering by the receiver filter the signal is sampled at time T and the cyclic prefix is removed, then an FFT is appliedon the remaining N samples.

We assume a Jakes model [7] for the channel, where each channel tap has a frequency shift of $\{f_\ell\}$. For the m-th subcarrier we denote with s_m and n_m the data signal with power E_s and the additive white Gaussian noise with variance per dimension $N_0/2$, respectively. The received signal on the m-th subcarrier after the FFT can be written as

$$\begin{aligned} y_m &= \frac{\sqrt{E_s}}{N_0} \sum_{k=0}^{N-1} s_k \sum_{j=0}^{N-1} e^{-i2\pi(m-k)j/N} \\ &\quad \cdot \sum_{\ell} \tilde{H}_\ell[k] e^{-i2\pi f_\ell T_j} + n_m \end{aligned} \tag{1}$$

where $m = 0, 1, \ldots, N-1$ and $\tilde{H}_\ell[k]$ are coefficients depending on the static description of the channel (see [5, eq. (4)]). A more compact description of the effects of the Doppler phenomena has been derived in [5], based on the expansion into a Taylor series of the term $e^{-i2\pi f_\ell T_j}$, around $f_\ell = 0$. In particular, the channel parameters of the p-th term of the Taylor expansion are collected into the vector $\mathbf{H}^{(p)} = [H_1^{(p)}, H_2^{(p)}, \ldots, H_N^{(p)}]^T$ (see [5, eq. (5)]). By denoting with $\mathbf{F}$ the Fourier matrix with entries $[\mathbf{F}]_{k,m} = \frac{1}{\sqrt{N}} e^{-i2\pi km/N}$, $0 \leq k, m < N$, and by defining the matrices

$$\Xi^{(p)} = \mathbf{F}\mathbf{D}^{(p)}\mathbf{F}^*, \quad \mathbf{D}^{(p)} = \text{diag}\{(k/N)^p\}, \quad p \geq 0, \tag{2}$$

from (1) a matrix form is derived for $\mathbf{y} = [y_0, y_1, \ldots, y_{N-1}]$

$$\mathbf{y} = \sqrt{E_s}\left(\sum_{p\geq 0} \Xi^{(p)} \text{diag}\{\mathbf{H}^{(p)}\}\right)\mathbf{s} + \mathbf{n}, \tag{3}$$

where $\mathbf{s}$ and $\mathbf{n}$ are the data and the noise vector, respectively. A simplified model may be obtained from (3) by observing that the power of the coefficients $\text{diag}\{\mathbf{H}^{(p)}\}$ is rapidly decreasing and hence the series may be truncated to the first P terms, [5]. The resulting expression of $\mathbf{y}$ relies on $N(P+1)$ parameters to describe the channel. Since the eigenspectrum of the correlation matrix among different frequencies has an exponential profile, a reduced set of channel parameters can be used. By defining the $N \times \underline{N}$ truncated Fourier basis $\underline{\mathbf{F}}$ containing the first $\underline{N}$ rows of $\mathbf{F}$, the relation between $\underline{\mathbf{H}}$ and $\mathbf{H}$ is given by

$$\mathbf{H} = (\mathbf{I}_{P+1} * \underline{\mathbf{F}})\underline{\mathbf{H}}, \tag{4}$$

where $\mathbf{I}_{P+1}$ is an identity matrix of size $(P+1) \times (P+1)$ and $*$ denotes the Kroneker product [5].

A. *The ICI cancellation scheme*

In order to compensate for the Doppler effect, the received signal vector $\mathbf{y}$ may be multiplied by the inverse of the matrix $\left(\sum_{p\geq 0} \Xi^{(p)}\text{diag}\{\mathbf{H}^{(p)}\}\right)$, as proposed in [4]. However, when the number of OFDM subcarriers is large, this technique turns out to be exceedingly complex.

As an alternative, an interference cancellation scheme has been proposed in [5] and is shown in Fig. 1. The basic principle of the scheme is the cancellation from the received signal of the data-dependent interference by using an initial tentative decision.

Let's define the diagonal matrix

$$\mathbf{C} = \sum_{p=0}^{P} \text{diag}\{\Xi^{(p)}\}\text{diag}\{\mathbf{H}^{(p)}\}. \tag{5}$$

In the ICI cancellation scheme the received block is first equalized by multiplying $\mathbf{y}$ by $\mathbf{C}^{-1}$ and then taking a decision with a slicer to obtain $\tilde{\mathbf{s}}$. Then, these decisions are used to regenerate the interference and delete it from the received signal. In particular, the new data estimates are given by

$$\hat{\mathbf{s}} = \mathbf{C}^{-1}\left(\mathbf{y} - \left(\sum_{p=1}^{P} \Xi^{(p)}\text{diag}\{\hat{\mathbf{H}}^{(p)}\}\right)\tilde{\mathbf{s}}\right). \tag{6}$$

The maximum likelihood (ML) parameters estimator for the ICI cancellation scheme was also derived, [5]. By defining the matrix

$$\underline{\Xi} = [\Xi^{(0)}\text{diag}\{\mathbf{s}\}\underline{\mathbf{F}}, \ldots, \Xi^{(P)}\text{diag}\{\mathbf{s}\}\underline{\mathbf{F}}], \tag{7}$$

the ML estimator for the reduced set of parameters is given by

$$\hat{\underline{\mathbf{H}}} = \frac{\sqrt{E_s}}{N_0}\left(\frac{E_s}{N_0}\underline{\Xi}^*\underline{\Xi} + (\mathbf{R}_c^{-1} \otimes \mathbf{R}_f^{-1})\right)^{-1} \underline{\Xi}^*\mathbf{y}, \tag{8}$$

where $\mathbf{R}_f$ and $\mathbf{R}_c$ are the correlation matrices among the subcarriers and the derivatives, respectively (see [5, eq. (8)–(10)]). The $\underline{N}(P+1) \times \underline{N}(P+1)$ matrix inverse is computed off-line if a fixed training sequence is used to estimate the channel. When the training sequence is not available and a decision directed estimation is performed, the matrix should be inverted for each instance of the data, with a considerable amount of complexity.

On the other hand, the product $\underline{\Xi}^*\mathbf{y}$ can be computed with a complexity of $\underline{N}(P+1)$ complex multiplications [5].

III. ITERATIVE ICI CANCELLATION FOR DVB-T

The OFDM-DFE scheme needs an estimation of the data in order to properly cancel the interference. However, due to the interference and the noise, the first decisions have a poor reliability and the errors on the decided data propagate to the next decisions so that the ICI cancellation process yields an additional

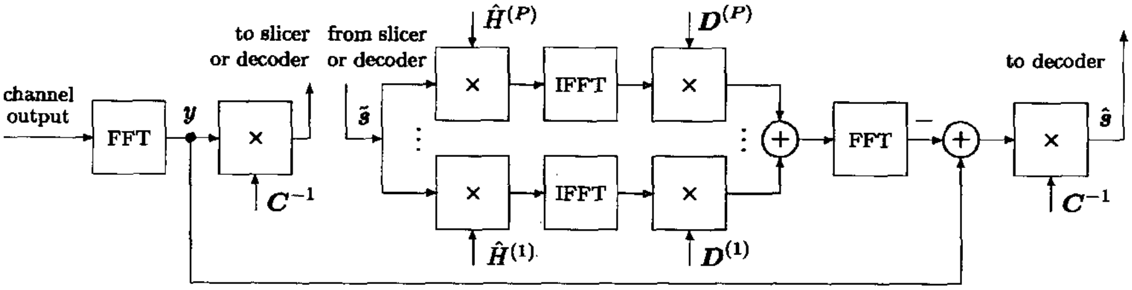

Fig. 1. The DFE-OFDM receiver with ICI removal.

disturbance to the received signal. Moreover, since a training sequence is not available in the DVB-T standard, a decision-directed estimation of the time-varying channel should be performed.

We propose here a new iterative scheme, which is shown in Fig. 2. In this receiver, a first decision on the data is taken after a standard equalization of the channel. Then the decided symbol is used to perform the channel estimation and to delete the interference. However, since the initial estimated data are significantly affected by Doppler interference and noise the resulting signal after the ICI cancellation may not satisfy the QoS requirements of DVB, even having a better reliability than the first estimation. Therefore these decisions are fed back to the channel estimator and ICI cancellation block to perform a more reliable ICI cancellation in a repetitive manner. The iteration process can be repeated many times since at each stage more reliable results are available.

Note that the iterative scheme fails to improve the performance when two or more subcarriers are strongly interfering each other. In this case no initial reliable decision is available and the cancellation turns out to be ineffective. Since the Doppler interference is stronger on adjacent subcarriers due to the correlation of their gains, it turns out that the badly affected data are grouped into clusters. By using interleaving, decoding and re-encoding the clusters of errors are efficiently spreaded and corrected and the ICI cancellation scheme turns out to be more effective. In Fig. 2 the slicer is replaced by a block that performs deinterleaving, decoding, re-encoding and interleaving.

IV. REDUCED COMPLEXITY SCHEMES

The DVB-T standard uses a large number of subcarriers and therefore the complexity of the ICI cancellation scheme and of the channel estimation is excessive. Note that, since no training sequence is available the matrix inversion in (8) must be performed for each OFDM symbol. Moreover, a large number

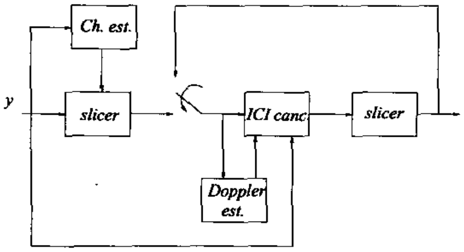

Fig. 2. The iterative ICI cancellation scheme.

of complex multiplications are required to perform the matrix multiplication $\Xi^*\Xi$ in (8). Therefore we consider some techniques to simplify the Doppler compensation scheme.

The ICI estimator proposed in [5] has a reduced complexity only when a training sequence is available for the channel estimation. In the DVB-T standard, only pilot symbols are provided for this purpose, which are not suitable for the estimation of the channel and its derivatives. The mutual interference among pilots turns out to be negligible compared to the interference on these pilots caused by the data transmitted on their adjacent carriers. Therefore the estimation cannot be simply based on only pilots. In this paper we investigate various decision driven channel estimation solutions to solve this problem. A first option is the use of the ML estimator on the entire symbol in a decision-directed fashion. In this case (8) is used to estimate the channel parameters and $\Xi^*\Xi$ must be adaptively computed, as well as the inverse at the denominator. Since the complexity of this scheme is exceedingly high we investigate a new iterative channel estimator. The iterative channel estimation scheme is shown in Fig. 3.

In this scheme the correlation among the different derivatives is neglected, this yields a suboptimal solution with respect to the ML estimator. But the complexity is significantly reduced because the matrix multiplication $\underline{\Xi}^*\underline{\Xi}$ is not required. First an estimation of $\hat{\mathbf{H}}^{(0)}$ is performed using the ML estimator of (8) and by assuming $P = 0$, i.e.

$$\begin{aligned}\underline{\hat{\mathbf{H}}}^{(0)} \;&=\; \frac{\sqrt{E_s}}{N_0}\left(\frac{E_s}{N_0}\underline{\mathbf{F}}^*\mathrm{diag}\{\mathbf{s}\}^*\mathrm{diag}\{\mathbf{s}\}\underline{\mathbf{F}}+\right.\\ &\left.(\mathbf{R}_c^{-1}\otimes\mathbf{R}_f^{-1})\right)^{-1}\underline{\Xi}^*\mathbf{y}\,. \qquad (9)\end{aligned}$$

Note that now $\underline{\mathbf{F}}^*\mathrm{diag}\{\mathbf{s}\}^*\mathrm{diag}\{\mathbf{s}\}\underline{\mathbf{F}}$ is a Toeplitz matrix and its first row and column are the first $\underline{N}$ samples of the FFT of the vector with entries $\{|s_k|^2\}$. By using the estimator (9) we are neglecting the correlation of noise and interference, as well as their correlation with the zero-derivative parameters. We observe that in this case the ML channel estimator is limited to the static part of the channel and reduced complexity estimators are available in literature [8, 9], this allows a trade-off between complexity and performance. In particular by substituting the product $\mathrm{diag}\{\mathbf{s}\}^*\mathrm{diag}\{\mathbf{s}\}$ with its averaged value, the matrix inversion can be computed off-line.

After $\underline{\hat{\mathbf{H}}}^{(0)}$ has been estimated, its contribution is deleted from the received signal, i.e., assuming correct decisions on the data and the channel

$$\mathbf{y}^{(1)} = \mathbf{y} - \hat{\mathbf{H}}^{(0)}\tilde{\mathbf{s}} = \sum_{p=1}^{P}\Xi^{(1)}\mathrm{diag}\{\mathbf{H}^{(p)}\}\mathbf{s} + \mathbf{n}\,. \qquad (10)$$

Then, by considering only the contribution of the first term of the sum in (10), the MMSE estimate of the product $\hat{\mathbf{H}}^{(1)}\tilde{\mathbf{s}}$ is obtained as follows

$$\tilde{\mathbf{y}}^{(1)} = \tilde{\Xi}^{(1)}_{MMSE}\mathbf{y}^{(1)}\,, \qquad (11)$$

where, from (2),

$$\tilde{\Xi}^{(1)}_{MMSE} = \mathbf{F}\mathrm{diag}\left\{\frac{N\sqrt{E_s}k}{E_s k^2 + N_0 N^2}\right\}\mathbf{F}^*\,. \qquad (12)$$

Lastly, $\mathbf{H}^{(1)}$ is estimated from $\tilde{\mathbf{y}}^{(1)}$ by using (9). Note that also in this case we are neglecting the correlation of noise and interference among different subchannels.

If more terms of the Taylor expansion are considered, the procedure is repeated by successively subtracting from the received signal the contribution of the previous Taylor terms and, after the multiplication by the matrix

$$\tilde{\Xi}^{(p)}_{MMSE} = \mathbf{F}\mathrm{diag}\left\{\frac{N^p\sqrt{E_s}k^p}{E_s k^{2p} + N_0 N^{2p}}\right\}\mathbf{F}^*\,, \qquad (13)$$

(9) is applied.

V. PERFORMANCE COMPARISON FOR THE DVB-T STANDARD

Simulation on the DVB-T standard considered both the 2K and the 8K mode, with various constellations. All simulations were performed on channel CH40 (626 MHz) with a bandwidth of 8 MHz. Note that the overall band for analog TV is between 400 and 790 MHz, so that the considered channel is roughly in the middle of the spectrum.

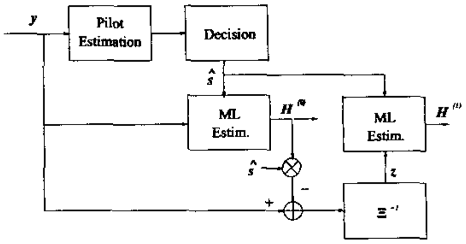

Fig. 3. The iterative estimation scheme.

The channel model is a Rayleigh channel with an exponential decaying power profile with root mean square delay spread of $\tau = 1.1\mu$s and a maximum delay of 7μs. This model approximates the Typically Urban (TU) defined by the COST 207 project for GSM [10].

The lines in Fig. 4 show the maximum speed that can be reached by the DVB system in different modes with different constellations. As a performance criterion to decide if the DVB system was still properly working we relayed on the test lab results obtained for the Motivate project [3]. According to the conclusions of this work we obtained a target bit error rate at the input of the Viterbi decoder for each code rate. For code rates of 1/2, 2/3 and 3/4 we obtained a target BER of $7\cdot10^{-3}$, $4\cdot10^{-3}$ and $2\cdot10^{-3}$, respectively. The maximum speed is the speed at when this BERs were reached.

A training OFDM block was used for synchronization purposes. The number of Taylor coefficients is $P = 1$, the number of iteration is 2 and the number of reduced parameters is $\underline{N} = 60$.

The lines in Fig. 4 show the following equalization/estimation modes:

- *Dashed line, filled circles* – One antenna receiver, standard equalization and channel estimation methods (no ICI cancellation), compare [3, Fig. 8].
- *Dashed line, empty circles* – One antenna receiver, standard equalization and channel estimation methods (no ICI cancellation), simulation results.
- *Continuous line, empty circles* – Two antennas receiver with maximum ratio combining, standard equalization and channel estimation methods (no ICI cancellation), compare [3, Fig. 8].
- *Dashed line, filled squares* – One antenna receiver with ICI cancellation, no error correction, Taylor coefficients estimation by training sequence and genie aided ICI cancellation. The data used to delete the interference are those actually transmitted. This curve represents an asymptotic bound on the performance.

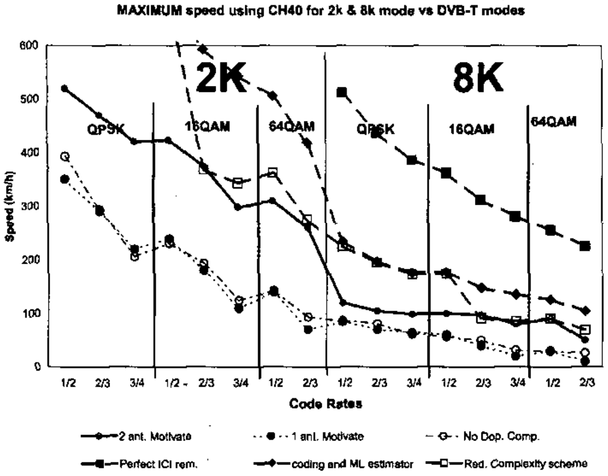

Fig. 4. Maximum speed for various estimation-equalization methods and various DVB modes.

- *Dashed line, empty diamonds* – One antenna receiver with ICI cancellation, error correction and ML estimator.
- *Dashed line, empty squares* – One antenna receiver with ICI cancellation, error correction and reduced complexity schemes, as reported in the previous section.

From Fig. 4 we see that the proposed ICI cancellation scheme has comparable performances as the two antennas system. Moreover, a significant potential is still available for the ICI cancellation scheme provided that the data fed into the ICI cancellation block is more accurate.

We observe that the reduced complexity scheme has comparable performances than the two antenna schemes. Moreover, if a scheme with no reduced complexity is considered, which includes the decoding and encoding feature, the performance is better. In particular, for the interesting configuration of 8k mode and code rate 2/3 the maximum speed with the ICI cancellation scheme is about 100 km/h while the two antenna system is about 50 km/h.

Fig. 5 shows the BER before Viterbi decoder as a function of the speed and of the number of iteration for the reduced complexity iterative interference cancellation scheme. For this figure the DVB-T mode is 2k, the code rate is 2/3, the constellation is 64-QAM. We note that the first iteration yields a significant improvement of the performance, while with further iterations the gain is decreasing.

VI. CONCLUSIONS

We have proposed reduced complexity interference cancellation schemes for a mobile use of the DVB-T system. Iterative interference cancellation and decoding has been considered.

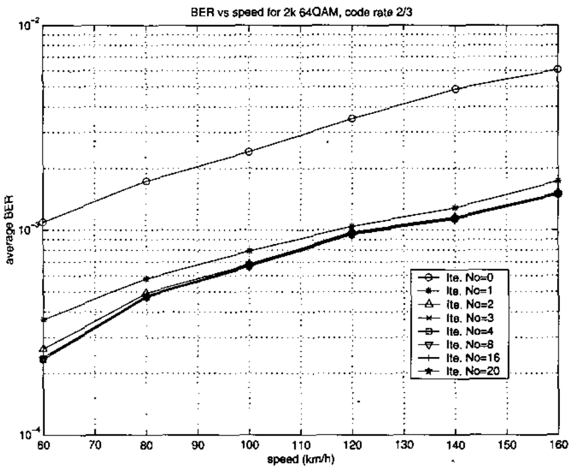

Fig. 5. Bit error rate as a function of the speed for different numbers of iterations.

In order to reduce the complexity of the resulting system, a new channel parameter estimator has been proposed. The basic building block of the channel estimator can be chosen among a variety of existing solutions, thus allowing a trade off between complexity and performance. As a result the reduced complexity scheme allows to fulfill the quality requirements for DVB-T at speeds that are similar to a system employing with two antennas at the receiver and maximum ratio combining.

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