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Iterative Interference Cancellation and Channel Estimation for Mobile OFDM

Stefano Tomasin, *Member, IEEE*, Alexei Gorokhov, *Member, IEEE*, Haibing Yang, and Jean-Paul Linnartz, *Senior Member, IEEE*

Abstract—In mobile reception, the reliability of orthogonal frequency division multiplexing (OFDM) is limited because of the time-varying nature of the channel. This causes intercarrier interference (ICI) and increases inaccuracies in channel tracking. In this paper, we model the ICI using derivatives of the channel amplitude. This allows us to design a relatively simple receiver scheme that iteratively cancels the ICI. The design of the canceler aims at maximizing the signal-to-noise-plus-ICI ratio at the detector input. We also propose a new channel estimator, and we show that it achieves reliable mobile reception in practical situations that are relevant to terrestrial Digital Video Broadcasting (DVB-T). Extensive simulations for a receiver with one or two antennas show that a small number of iterations between ICI cancellation and channel estimation allow a reliable reception at vehicle speeds above 100 km/h.

Index Terms—Channel estimation, interference cancellation, mobile communications, orthogonal frequency division multiplexing (OFDM).

I. INTRODUCTION

ORTHOGONAL frequency division multiplexing (OFDM) is widely recognized as an efficient modulation method for wireless communication. In OFDM, a stream of signals are modulated on many equally spaced parallel subcarriers [1]. Demodulation and modulation are efficiently implemented by means of a fast Fourier transform (FFT) and its inverse, respectively. In transmissions over a radio channel, the orthogonality of the signals is maintained only if the channel is flat and time-invariant. However, in a dispersive channel, self-interference occurs among successive symbols at the same subcarrier [intersymbol interference (ISI)], as well as among signals at different subcarriers [intercarrier interference (ICI)]. For a time-invariant but frequency-selective channel, ICI, as well as ISI, can effectively be avoided by inserting a cyclic

prefix before each block of parallel data symbols [2]. This paper addresses time-varying channels, as in mobile reception, where ICI occurs among the different subcarriers [3]–[5].

Currently, the use of terrestrial digital video broadcasting (DVB-T) is extended to deliver high-rate multimedia applications and services on the move; for instance, broadband Internet, provided that an alternative low-bandwidth cellular return channel is used. At the same time, both DVB-T spectrum efficiency requirements and the deployment of single-frequency networks (SFNs) [6] favor the use of the “8k” mode, i.e., using 8192 subcarriers. Regrettably, ICI is more severe for such systems using a large number of densely spaced subcarriers, and its effects become particularly disturbing at freeway speeds of the car.

Some solutions for ICI mitigation, such as precoding for self-cancellation [7], require a modification of the transmit format so that these are not suitable for existing standards. Other techniques are compatible with existing transmit schemes, but may not be suitable for high vehicle speeds or become too complex for consumer products. For instance, the conventional combining of multiple receive antennas [8], [9] loses effectiveness at high speeds. In fact, basic diversity methods treat ICI as additional additive white Gaussian noise (AWGN). Preferably, however, the receiver should exploit the fact that ICI is structured according to the transmit standard. This approach is used in this paper, but it was earlier also proposed in the adaptive matrix equalizer [10], although its complexity can be prohibitive for systems with a large number of subcarriers. In an effort to simplify the ICI cancellation, [11] proposed a two-stage decision feedback scheme (called DF-ICI). Initially, a standard equalizer, i.e., a one-tap filter per subcarrier, is applied. The tentatively detected data is used to regenerate the ICI, which is then subtracted from the received signal. A maximum likelihood (ML) estimator for the channel parameters has been derived for the special case when a dedicated training block is transmitted. Our paper relieves some of the drawbacks of the DF-ICI solution and mitigates the effect of errors in the initial estimate of the data. Channel estimation in a mobile setting deserves our special attention. Simulations in [9] revealed that relatively long interpolators of pilot signals are needed for the accurate estimation of time-varying channels.

Here, we address OFDM transmission using long FFT blocks and dense modulation constellations, but without a training sequence, with reference to mobile reception of 8k DVB-T. The objective is to allow reception at speeds above 100 km/h, while restricting ourselves to algorithms with acceptable complexity. To this end, we adopt an iterative (“turbo”) ICI cancellation and

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channel estimation scheme. A first conclusion will be that typical estimators based only on pilot symbols are inadequate. Instead, we resort to a decision-directed, iterative estimator and we will show that a reliable reception of DVB-T signals can be achieved at speeds beyond 100 km/h, using a single antenna.

This paper is organized as follows. Section II formulates a statistical model in the frequency domain for OFDM transmission over a mobile channel. This model will be used merely to design analytically our filters, whereas a time-domain model with a discrete number of rays will be used in the simulations. Section III presents the iterative ICI cancellation scheme. Channel estimation is discussed in two sections: Section IV focuses on a (quasi-) static channel, while Section V introduces rapid time variations. Section VI offers numerical results, including the combined use of multiple antennas and ICI compensation. Section VII concludes the paper.

II. SYSTEM MODEL

In OFDM modulation, a data stream with rate $R = 1/T$ is divided into N substreams of rate R/N , which are modulated with rectangular shape filters on different subcarriers, equally spaced at $f_s = 1/(TN)$. Let s_m denote the user data, with mean power E_s for subcarrier m ($m = 0, 1, \dots, N-1$). We define $j = \sqrt{-1}$. For a single block of data, the modulated signal at the q th instant of time is

$$S_q = \frac{1}{N} \sum_{m=0}^{N-1} s_m e^{2\pi j \frac{mq}{N}}, \quad q = 0, 1, \dots, N-1. \quad (1)$$

Following a common practice in OFDM systems, we insert a cyclic prefix, i.e., we start the transmit block $\tilde{\mathbf{S}}$ with the last L samples of \mathbf{S} ; thus

$$\tilde{S}_q = \begin{cases} S_{q+(N-L)}, & q = 0, 1, \dots, L-1 \\ S_{q-L}, & q = L, L+1, \dots, L+N-1. \end{cases} \quad (2)$$

The transmit and receive filters and synchronization reference t_0 are captured in the impulse response $g(\tau - t_0)$, where τ is the delay variable.

A. Time-Varying Channel Model

The channel is modeled as a collection of $i = 1, 2, \dots, N_h$ resolvable multipath components, each with its delay τ_i and time-varying tap gain $\tilde{h}_i(t)$. So the time-varying impulse response becomes

$$h(t, \tau) = \sum_{i=1}^{N_h} \tilde{h}_i(t) \delta(\tau - \tau_i). \quad (3)$$

The channel is sampled at ℓT , although it continuously varies with t . To include the transmit and receive filters, the equivalent discrete-time channel taps are

$$c_\ell(t) = \sum_{i=1}^{N_h} \tilde{h}_i(t) g(t_0 - \tau_i + \ell T), \quad \ell = 1, 2, \dots, N. \quad (4)$$

The signal received at the m th subcarrier y_m includes the wanted signal plus additive white Gaussian noise n_m with zero mean and variance $N_0/2$ per dimension. Thus

$$y_m = \sum_{q=0}^{N-1} e^{-2\pi j \frac{qm}{N}} \sum_{\ell} c_\ell(qT) \tilde{S}_{(q-\ell+L)} + n_m \quad (5)$$

where $m = 0, 1, \dots, N-1$.

For our design of ICI cancellation filters, we further develop this channel model by expanding the path gains $\tilde{h}_i(t)$ into a Taylor series as a function of t around $t = (NT/2)$, having set $t = 0$ in correspondence of the first sample at the input of the FFT at the receiver. Hence, we expand the impulse response $h(t, \tau)$ only with respect to t , rather than to τ . Let $\beta_i^{(p)}$ indicate the p th derivative of $\tilde{h}_i(t)$ at $t = NT/2$

$$\beta_i^{(p)} = \left(\frac{d^{(p)} \tilde{h}_i(t)}{d^{(p)} t} \right)_{t=\frac{NT}{2}}. \quad (6)$$

Particularly, if $\tilde{h}_i(t)$ describes only a few reflections (i.e., N_j is small as we will discuss in Section II-B), these derivatives are finite and rapidly decrease in strength for higher orders p . To adopt an efficient vector-channel notation, we define

$$H_k^{(p)} = \frac{T^p}{p!} \sum_{i=1}^{N_h} \beta_i^{(p)} \sum_{\ell} g(t_0 - \tau_i + \ell T) e^{-2\pi j \frac{k\ell}{N}} \quad (7)$$

where $k = 0, 1, \dots, N-1$, $p \geq 0$, and we collect the p th derivative in the vectors $\mathbf{H}^{(p)} = [H_0^{(p)}, H_1^{(p)}, \dots, H_{N-1}^{(p)}]$. In fact, this allows us to write the received signal as

$$y_m = \frac{1}{N} \sum_{k=0}^{N-1} s_k \sum_{q=0}^{N-1} e^{-2\pi j \frac{(m-k)q}{N}} \sum_{p=0}^{\infty} H_k^{(p)} \left(q - \frac{N}{2} \right)^p + n_m. \quad (8)$$

Gathering $\{y_m\}$, $\{s_m\}$, and $\{n_m\}$ in the $N \times 1$ vectors \mathbf{y} , \mathbf{s} , and \mathbf{n} , respectively, the matrix notation for (8) becomes as compact as

$$\mathbf{y} = \left(\sum_{p=0}^{\infty} \mathbf{\Xi}^{(p)} \text{diag} \{ \mathbf{H}^{(p)} \} \right) \mathbf{s} + \mathbf{n} \quad (9)$$

where $\text{diag}\{\mathbf{H}^{(p)}\}$ is the diagonal matrix created from the column vector $\mathbf{H}^{(p)}$. The $N \times N$ matrices $\mathbf{\Xi}^{(p)}$ with entries

$$[\mathbf{\Xi}^{(p)}]_{m,k} = \frac{1}{N} \sum_{q=0}^{N-1} \left(q - \frac{N}{2} \right)^p e^{-2\pi j \frac{q(m-k)}{N}} \quad (10)$$

$m, k = 0, 1, \dots, N-1$, $p = 0, 1, \dots$, describe the *leakage* [12] from the transmit signal of subcarrier k to the receive subcarrier m . Interestingly, $\{\mathbf{\Xi}^{(p)}\}$ is a fixed matrix; that is, it acts as a system constant which neither adapts to the data nor to the channel $\mathbf{H}^{(p)}$.

The power of the ICI components, given the p th derivative, is

$$\sigma^{(p)2} = \text{E} \left[\left| \left(\mathbf{\Xi}^{(p)} \text{diag} \{ \mathbf{H}^{(p)} \} \mathbf{s} \right)_m \right|^2 \right] \quad (11)$$

where $(\mathbf{x})_m$ denotes the m th element of the vector \mathbf{x} . Ignoring some boundary effects, the power of ICI is equally distributed over all subcarriers.

In a typical fading channel model, $\tilde{h}_i(t)$, $i = 1, 2, \dots, N_h$ are complex zero-mean independent random processes, each with variance μ_i . Moreover, the statistical description of the time-variation of each tap is the same for all taps, i.e., $\tilde{h}_i(t) = \mu_i(\tau_i)\nu_i(t)$, where $\nu_i(t)$ are independent and identically distributed (i.i.d.) and do not depend on the delay τ_i . By indicating with T the transpose operator, from (7), the $(k_1 + pN, k_2 + qN)$ element of the correlation matrix of $\mathbf{H} = [\mathbf{H}^{(0)T}, \mathbf{H}^{(1)T}, \dots, \mathbf{H}^{(P-1)T}]^T$ can be written as (shown in the first equation at the bottom of the page) where H and $*$ denote Hermitian operator and complex conjugate, respectively.

One can separate the terms for $i_1 = i_2$ and those for $i_1 \neq i_2$. Moreover, it appears that the correlation matrix of $\beta_i^{(p)}$ can be factorized as the product of one term related only to the order of the derivative p and q [in (13) and (15) below] and another term depending on the tap number i and the frequency separation [indexes n and m in (14) and (16) below]. More precisely, the correlation matrix of \mathbf{H} can be written as

$$\mathbf{R}_H = \mathbb{E}[\mathbf{H}^H \mathbf{H}] = \mathbf{R}_{c1} \otimes \mathbf{R}_{f1} + \mathbf{R}_{c2} \otimes \mathbf{R}_{f2} \quad (12)$$

where \otimes denotes the Kronecker product. R_{f1} and R_{f2} are $N \times N$ matrices, and R_{c1} and R_{c2} are $P \times P$ matrices. Let us denote with $\nu_i^{(p)}$ the p th derivative of ν_i at $t = NT/2$. At position (p, q) , we see (13)–(16), found at the bottom of the page, where expectations are taken over all possible random channels according to the delay profile $\{\tau_i, \mu_i\}$.

In order to simplify the model of a general time-varying channel into a reduced set of parameters, we focus on the singular value decomposition (SVD) of the correlation matrices. Moreover, we truncate the infinite series of derivatives in (9) to P terms, ($p = 0, 1, \dots, P - 1$) to approximate \mathbf{y} with NP channel parameters.

Let $\mathbf{U}_1^{(p)}$ and $\mathbf{U}_2^{(p)}$ denote the eigenvector matrices of $[\mathbf{R}_{c1}]_{p,p} \mathbf{R}_{f1}$ and $[\mathbf{R}_{c2}]_{p,p} \mathbf{R}_{f2}$, respectively. Let us define $\mathbf{\Lambda}_1^{(p)}$ and $\mathbf{\Lambda}_2^{(p)}$ as diagonal matrices having as elements the eigenvalues of $[\mathbf{R}_{c1}]_{p,p} \mathbf{R}_{f1}$ and $[\mathbf{R}_{c2}]_{p,p} \mathbf{R}_{f2}$, respectively.

The singular value decomposition (SVD) of $[\mathbf{R}_{c1}]_{p,p} \mathbf{R}_{f1}$ and $[\mathbf{R}_{c2}]_{p,p} \mathbf{R}_{f2}$ yields

$$\mathbb{E} \left[\mathbf{H}^{(p)H} \mathbf{H}^{(p)} \right] = \mathbf{U}_1^{(p)H} \mathbf{\Lambda}_1^{(p)} \mathbf{U}_1^{(p)} + \mathbf{U}_2^{(p)H} \mathbf{\Lambda}_2^{(p)} \mathbf{U}_2^{(p)}. \quad (17)$$

Rearranging the SVD matrices, we can organize the eigenvalues of $\mathbf{\Lambda}_1^{(p)}$ and $\mathbf{\Lambda}_2^{(p)}$ in decreasing order from position (1,1) to $(\underline{N}_1 + 1, \underline{N}_1 + 1)$, respectively. We ignore the lowest eigenvalues beyond the $(N - \underline{N})$ largest ones in $[\mathbf{R}_{c1}]_{p,p} \mathbf{R}_{f1}$ and $[\mathbf{R}_{c2}]_{p,p} \mathbf{R}_{f2}$, respectively. We approximate each derivative vector $\mathbf{H}^{(p)}$ with the vector $\underline{\mathbf{H}}^{(p)}$ of size $\underline{N} = \underline{N}_1 + \underline{N}_2$ as

$$\mathbf{H}^{(p)} \approx \underline{\mathbf{F}}^{(p)} \underline{\mathbf{H}}^{(p)} \quad (18)$$

where $\underline{\mathbf{F}}^{(p)}$ is the $N \times \underline{N}$ matrix containing the eigenvectors associated with the \underline{N}_1 and \underline{N}_2 most significant eigenvalues.

The model (18) describes the channel with a reduced set of parameters, and N_1 and N_2 can be chosen according to the desired approximation. Note that for a time-invariant channel, (18) is reduced to the SVD channel description of [13].

B. Jakes' Simulation Model

In our simulations we follow Jakes' [14] proposal to approximate a Rayleigh-fading amplitude by a small number N_J of discrete components. This gives for the i th resolvable path

$$\tilde{h}_i(t) = h_i \frac{1}{N_J} \sum_{\ell=0}^{N_J-1} e^{2\pi j f_i N_J + \ell (t - \frac{TN}{2})}. \quad (19)$$

In this case, the Doppler shifts $\{f_\ell\}$ are i.i.d. random variables defined as

$$f_\ell = \left(\frac{v}{c} \right) f_c \cos \theta_\ell \quad (20)$$

where v is the vehicle speed, c the velocity of light, and f_c the carrier frequency. For an exponential delay profile with rms delay spread of τ_{rms} , we obtain $\mu_i = e^{-\tau_i/\tau_{rms}}$ and $\nu_i = (1/N_J) \sum_{\ell=0}^{N_J-1} e^{2\pi j f_i N_J + \ell (t - TN/2)}$.

Our simulation follows the Jakes' model (19), by taking equally spaced angles of arrival θ_ℓ in the interval $[0, 2\pi)$.

$$\left(\mathbb{E}[\mathbf{H}^H \mathbf{H}] \right)_{(k_1+pN, k_2+qN)} = \frac{T^{p+q}}{q!p!} \sum_{i_1} \sum_{i_2} \mathbb{E} \left[\frac{d^{(p)} \tilde{h}_{i_1}}{d^{(p)} t} \frac{d^{(q)} \tilde{h}_{i_2}^*}{d^{(q)} t} \times \sum_{\ell_1} \sum_{\ell_2} g(t_0 - \tau_{i_1} + \ell_1 T) e^{-2\pi j \frac{(k_1 \ell_1 - k_2 \ell_2)}{N}} g(t_0 - \tau_{i_2} + \ell_2 T) \right]$$

$$[\mathbf{R}_{c1}]_{p,q} = \frac{T^{p+q}}{p!q!} \mathbb{E} \left[\nu_i^{(p)} \right] \mathbb{E} \left[\nu_i^{(q)*} \right] \quad (13)$$

$$[\mathbf{R}_{f1}]_{n,m} = \sum_{i_1} \sum_{i_2 \neq i_1} \sum_{\ell_1} \sum_{\ell_2} \mathbb{E} [g(t_0 - \tau_{i_1} + \ell_1 T) \times g^*(t_0 - \tau_{i_2} + \ell_2 T) \mu_{i_1} \mu_{i_2}^*] \times e^{-2\pi j \frac{(\ell_1 k - \ell_2 m)}{N}} \quad (14)$$

$$[\mathbf{R}_{c2}]_{p,q} = \frac{T^{p+q}}{p!q!} \mathbb{E} \left[\nu_i^{(p)} \nu_i^{(q)*} \right] \quad (15)$$

$$[\mathbf{R}_{f2}]_{n,m} = \sum_i \sum_{\ell_1} \sum_{\ell_2} \mathbb{E} \left[|g(t_0 - \tau_i + \ell_1 T)|^2 |\mu_i|^2 \right] \cdot e^{-2\pi j \frac{(\ell_1 k - \ell_2 m)}{N}} \quad (16)$$

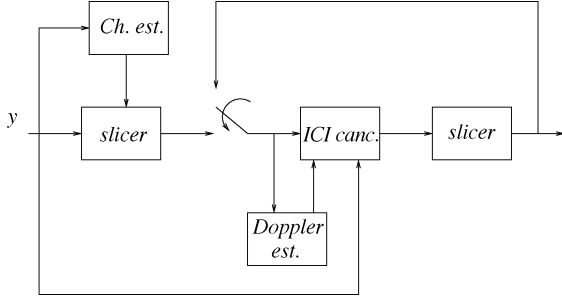


Fig. 1. Doppler compensation scheme.

III. ITERATIVE ICI CANCELLATION

In [10], a feedforward linear filter was proposed to compensate the ICI, but it requires the adaptation of a matrix equalizer to the channel variations, which is relatively complex. Instead, we consider a decision feedback equalizer (DFE), which performs an iterative ICI cancellation. Iterative detection and cancellation was previously used to reduce the multiuser access interference (MAI) in the parallel interference cancellation (PIC) for code division multiple access (CDMA) [15]. Also, iterative cancellation was recently proposed to alleviate the ISI in single carrier transmission [16].

A. Iterative DFE Scheme

In the first round (iteration $n = 1$) of our DFE, the received OFDM block \mathbf{y} is equalized by using a single tap, i.e., one complex multiplication per subcarrier. Tentative symbol decisions $\hat{\mathbf{s}}^{(1)}$ are used to regenerate an estimate of the ICI, which is subtracted from \mathbf{y} in the next iteration. As shown Fig. 1, for iteration $n = 2, 3, \dots, N_I$, the previously detected data $\hat{\mathbf{s}}^{(n-1)}$ is used to partially cancel the interference using the feedback matrix filter $\mathbf{B}^{(n)}$ according to

$$\mathbf{y}^{(n)} = \mathbf{y} - \mathbf{B}^{(n)}\hat{\mathbf{s}}^{(n-1)}, \quad n = 2, 3, \dots, N_I. \quad (21)$$

For $n = 1$, we set the initial conditions to $\mathbf{B}^{(1)} = \mathbf{0}$ and $\hat{\mathbf{s}}^{(0)} = \mathbf{0}$. Various design criteria can be adopted for the feedback filter $\mathbf{B}^{(n)}$. For example, with $N_I = 2$ and

$$\mathbf{B}^{(2)} = \sum_{p=1}^{P-1} \Xi^{(p)} \text{diag} \left\{ \mathbf{H}^{(p)} \right\} \quad (22)$$

the DF-ICI of [11] is recovered. In addition, the channel parameters $\mathbf{H}^{(p)}$ need to be estimated in the presence of noise and ICI. In iterative schemes, the estimation filter preferably is optimized (per iteration) to match the reliability of the previously detected data (see Section III-B). In summary, we propose the following iterative algorithm.

- 1) Set $n = 1$.
- 2) Perform data detection, de-interleaving, decoding, and re-encoding and re-interleaving according to the DVB-T standard.
- 3) Compute $\mathbf{B}^{(n)}$ and perform ICI cancellation.
- 4) Increment n by 1.
- 5) If $n < N_I$, go to point 2.

B. Design of Feedback Filter

The feedback filters $\mathbf{B}^{(n)}$ intend to maximize the signal-to-interference-plus-noise ratio (SINR) at the input of the decoder. So, we first analyze this SINR as a function of the iteration step n . Let us assume that errors on the data are independent of both the subcarrier and the data symbols. We approximate the errors at the n th iteration $\mathbf{e}^{(n)} = \hat{\mathbf{s}}^{(n)} - \mathbf{s}$ as a random vector of i.i.d. random complex Gaussian variables, with zero mean and uncorrelated elements. By defining the matrix

$$\mathbf{E}^{(n)} = \sum_{p=1}^{P-1} \Xi^{(p)} \text{diag} \left\{ \mathbf{H}^{(p)} \right\} - \mathbf{B}^{(n)} \quad (23)$$

the SINR before decoding for the m th subcarrier is defined as

$$Z_m^{(n)} = \frac{E_s \left| H_m^{(0)} \right|^2}{\mathbb{E} \left[\left| \mathbf{E}_m^{(n)} \mathbf{s} - \mathbf{B}_m^{(n)} \mathbf{e}^{(n-1)} \right|^2 \right] + N_0} \quad (24)$$

where $(\mathbf{X})_m$ denotes the m th row of \mathbf{X} . For a large N and for sufficiently selective channels, the numerator and denominator can be treated as independent variables, so we may write

$$Z^{(n)} = \frac{E_s \left\| \mathbf{H}^{(0)} \right\|^2}{\mathbb{E} \left[\left\| \mathbf{E}^{(n)} \mathbf{s} - \mathbf{B}^{(n)} \mathbf{e}^{(n-1)} \right\|^2 \right] + N_0} \quad (25)$$

where $\|\mathbf{X}\|^2 = \sum_r \sum_c |X_{r,c}|^2$.

Let us indicate the average SINR *after* decoding of the n th iteration as

$$\Gamma_n = \frac{\mathbb{E} \left[\|\mathbf{s}\|^2 \right]}{\mathbb{E} \left[\left\| \mathbf{e}^{(n)} \right\|^2 \right]}. \quad (26)$$

A feedback filter that minimizes the denominator of (25), maximizes $Z^{(n)}$, and in particular

$$\mathbf{B}^{(n)} = \frac{1}{1 + \frac{1}{\Gamma_{n-1}}} \sum_{p=1}^{P-1} \Xi^{(p)} \text{diag} \left\{ \mathbf{H}^{(p)} \right\}. \quad (27)$$

In order to find the relation between Γ_n and the symbol error probability $P_e^{(n)}$ at the n th iteration, we observe that the most probable errors occur when a constellation point is mistaken for one of its nearest points. So, by indicating with d_{\min} the minimum distance between any two constellation points, an approximate expression of Γ_n is given by

$$\Gamma_n \approx \frac{E_s}{P_e^{(n)} d_{\min}^2}. \quad (28)$$

A closed-form expression for $P_e^{(n)}$, related to the statistical properties of the channel, in particular to $Z_m^{(n)}$, is derived in [17].

IV. DESIGN OF AMPLITUDE ESTIMATOR

Similar to most OFDM systems, DVB-T uses a subset of all subcarriers to transmit a reference signal (*pilot tones*) to estimate the channel [18], [19]. However, the MSE of the pilot estimate increases exponentially with the speed, so that DVB-T

reception on a moving car becomes inaccurate, as confirmed also in the MOTIVATE reports [9]. Consequently, we investigate other estimation strategies, which include the time-varying nature of the channel.

A. Predictive Estimator

To reduce the MSE of the initial amplitude estimate, we rely on a predictive method based on the simplified model (18). It predicts the complex amplitude of each subcarrier, using the estimate of the zero and first derivative ($P = 2$).

A prediction of the derivative for the $(k + 1)$ OFDM block is obtained from the derivatives estimated at time k as

$$\mathbf{H}'^{(0)}(k + 1) = \bar{\mathbf{H}}^{(0)}(k) + T' \bar{\mathbf{H}}^{(1)}(k). \quad (29)$$

In order to reduce the estimation error, the predicted estimate is a weighted average of the pilot-based estimate and a previous estimate

$$\hat{\mathbf{H}}^{(0)}(k) = \alpha_k \hat{\mathbf{H}}^{(0)}(k) + (1 - \alpha_k) \mathbf{H}'^{(0)}(k) \quad (30)$$

where initially, the weight factor $\alpha_0 = 1$.

The weights α_k are chosen in order to minimize the MSE of the estimate. In particular, we indicate with γ_0^2 and γ_k^2 , the MSE of the estimate $\hat{\mathbf{H}}^{(0)}(k)$ and $\mathbf{H}'^{(0)}(k)$, respectively. The variance of the error of the estimate of the zero-order derivative at time k is then

$$\begin{aligned} \sigma_k^2 &= \mathbb{E} \left[\left\| \hat{\mathbf{H}}^{(0)}(k) - \mathbf{H}^{(0)} \right\|^2 \right] \\ &= (1 - \alpha_k)^2 \gamma_{k-1}^2 + \alpha_k^2 \gamma_0^2 \end{aligned} \quad (31)$$

and minimization of σ_k^2 , with respect to α_k , yields

$$\alpha_k = \frac{\gamma_{k-1}^2}{\gamma_0^2 + \gamma_{k-1}^2}. \quad (32)$$

To simplify the estimator, a fixed value $\alpha = \bar{\alpha}$ can be used instead of (32). For the considered simulation scenario, we empirically optimize the weight to $\bar{\alpha} = 0.6$.

B. Iterative Estimator for Zero-Order Derivative

A refinement of the estimation is obtained by iterating the estimation of the zero-order derivative and the data detection. The resulting technique is shown in Fig. 2. In particular, the data of all subcarriers are initially detected, to obtain $\hat{\mathbf{s}}^{(0)}$, based on the pilot tones. Then, $\hat{\mathbf{s}}^{(0)}$ is used to perform the ML estimation, which indeed is the minimum mean square error estimation of the reduced set of parameters for the zero order derivative, i.e.,

$$\begin{aligned} \hat{\mathbf{H}}^{(0)} &= \left(\frac{E_s}{N_0} \mathbf{F}^{(0)H} \text{diag} \{ \hat{\mathbf{s}}^{(0)} \}^H \text{diag} \{ \hat{\mathbf{s}}^{(0)} \} \mathbf{F}^{(0)} + \mathbf{R}^{(0)-1} \right)^{-1} \\ &\quad \times \mathbf{F}^{(0)H} \text{diag} \{ \hat{\mathbf{s}}^{(0)} \}^H \mathbf{y} \end{aligned} \quad (33)$$

where $\mathbf{R}^{(p)} = \mathbb{E}[\mathbf{H}^{(p)H} \mathbf{H}^{(p)}]$. Note that for a channel response with equally distant taps $\tau_\ell = \ell T$, the product

$$\mathbf{F}^{(0)H} \text{diag} \{ \hat{\mathbf{s}}^{(0)} \}^H \text{diag} \{ \hat{\mathbf{s}}^{(0)} \} \mathbf{F}^{(0)} \quad (34)$$

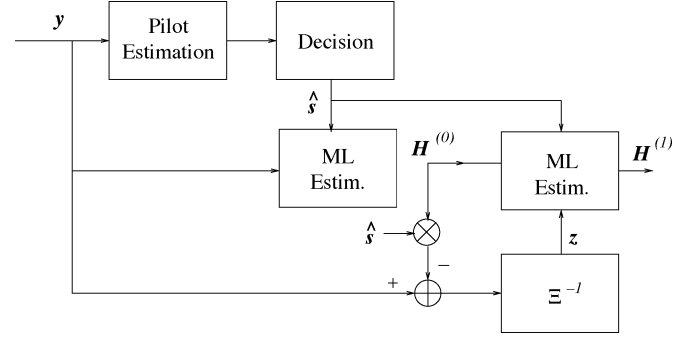


Fig. 2. ICI cancellation with iterative zero-derivative estimation.

is a Toeplitz matrix. The first row and first column both consist of \underline{N} samples of the DFT of the square of the data samples, i.e.,

$$\left[\mathbf{F}^{(0)H} \text{diag} \{ \hat{\mathbf{s}}^{(0)} \}^H \text{diag} \{ \hat{\mathbf{s}}^{(0)} \} \mathbf{F}^{(0)} \right]_{0,\ell} = \sum_k \left| \hat{s}_k^{(0)} \right|^2 e^{-2\pi j \frac{\ell k}{N}}$$

where $\ell = 0, 1, \dots, \underline{N} - 1$. Moreover, estimators based on the average ML estimate [20], [21], having a reduced complexity, could be used.

V. HIGHER ORDER DERIVATIVE ESTIMATION

For the estimation of the derivatives $\{\mathbf{H}^{(p)}\}$, the ML estimator was introduced in [11]. Since matrices $\Xi^{(p)}$ are not orthogonal, the ML estimator can not be divided into P parallel ML estimators, one for each derivative, but it operates jointly on all of them. In particular, by defining the matrix

$$\Xi = \left[\Xi^{(0)} \text{diag} \{ \mathbf{s} \} \mathbf{F}^{(0)}, \dots, \Xi^{(P-1)} \text{diag} \{ \mathbf{s} \} \mathbf{F}^{(P-1)} \right] \quad (35)$$

the ML estimator that uses \underline{N} rather than N parameters is

$$\hat{\mathbf{H}} = \left(\frac{E_s}{N_0} \Xi^H \Xi + N_0 \mathbf{R}^{-1} \right)^{-1} \Xi^H \mathbf{y} \quad (36)$$

where $\mathbf{R} = \mathbb{E}[\mathbf{H}^H \mathbf{H}]$ and $\mathbf{H} = [\mathbf{H}^{(0)}, \mathbf{H}^{(1)}, \dots, \mathbf{H}^{(N-1)}]$. In order to compute $\Xi^H \Xi$, a matrix of size $P\underline{N} \times \underline{N}$ should be multiplied by its Hermitian conjugate, and the multiplication has a complexity of $\mathcal{O}(N^2 \underline{N} P)$. If a training sequence is available, the multiplication can be precomputed offline, and the resulting matrix can be stored. When no training sequence is available (as in DVB-T), the multiplication should be performed on the received data $P\underline{N} \times \underline{N}$. Hence, ML estimation is infeasible for the 8 k mode of DVB-T.

A. Multistage Channel Estimator

In order to simplify the estimation process of the ICI derivatives, we consider here a suboptimal approach. In the multistage scheme, we estimate first the derivatives with higher power by modeling the contribution of lower power derivatives as white noise. For the channel model [11, (19)], it has been shown that the power of the derivatives is decaying faster than exponentially. At each step, the contribution of the previously estimated derivatives is removed from the received signal, and the next order of derivatives is estimated.

The estimator works as follows. From the signal model (9), at the p th stage, we compute the vector

$$\mathbf{r}^{(p)} = \mathbf{y} - \sum_{v=0}^{p-1} \Xi^{(v)} \text{diag} \left\{ \hat{\mathbf{H}}^{(v)} \right\} \mathbf{s} = \Xi^{(p)} \text{diag} \left\{ \hat{\mathbf{H}}^{(p)} \right\} \mathbf{s} + \mathbf{w}^{(p)} \quad (37)$$

where $\mathbf{w}^{(p)}$ includes residual interference and noise. By performing a minimum MSE inversion of $\Xi^{(p)}$, we obtain an approximated version of $\text{diag} \left\{ \mathbf{H}^{(p)} \right\} \mathbf{s}$

$$\begin{aligned} \mathbf{G}^{(p)} &= \left(\Xi^{(p)*} \Xi^{(p)} + \tilde{\Gamma}^{(p)} \mathbf{I}_P \right)^{-1} \Xi^{(p)*} \mathbf{r}^{(p)} \\ &= \left(\Xi^{(p)*} \Xi^{(p)} + \tilde{\Gamma}^{(p)} \mathbf{I}_P \right)^{-1} \Xi^{(p)*} \Xi^{(p)} \text{diag} \left\{ \mathbf{H}^{(p)} \right\} \mathbf{s} + \mathbf{q}^{(p)} \end{aligned}$$

where $\mathbf{q}^{(p)} = \left(\Xi^{(p)*} \Xi^{(p)} + \tilde{\Gamma}^{(p)} \mathbf{I}_P \right)^{-1} \Xi^{(p)*} \mathbf{w}^{(p)}$, and $\tilde{\Gamma}^{(p)}$ is the SINR factor that includes the ICI due to the derivatives higher than p , i.e.,

$$\tilde{\Gamma}^{(p)} = \frac{\sigma^{(p)2}}{N_0 + \sum_{k>p} \sigma^{(k)2}}. \quad (38)$$

An estimate of $\mathbf{H}^{(p)}$ is then carried out under the assumption that the vector $\mathbf{q}^{(p)}$ is zero-mean Gaussian distributed by using (33). Indeed, in the special case of a Rayleigh channel, $\mathbf{q}^{(p)}$ is a jointly Gaussian vector. The procedure reduces the joint estimation of the P derivatives into the separate estimation of a set of zero-order derivatives. Note also that the matrix multiplication in (33) is carried out once per symbol.

If the multistage channel estimator is included into the iterative scheme, the derivatives' estimation $\left\{ \hat{\mathbf{H}}_q^{(p)} \right\}$ at iteration q can be used to delete the ICI contribution, including all derivatives of order different from q .

VI. NUMERICAL RESULTS

We evaluated the receiver performance for a DVB-T system in 2k and 8k mode, thus having $N = 2048$ and $N = 8192$ subcarriers, respectively [18]. Various lengths of the cyclic prefix, constellation sizes, and code rates were considered. Synchronization has been acquired by applying an algorithm based on the correlation of an OFDM block used as training. We used a Rayleigh fading channel with an exponentially decaying power profile: a mean square delay spread of $\tau_{\text{rms}} = 1.1 \mu\text{s}$ truncated at $7 \mu\text{s}$. These parameters are in accordance with the Typically Urban (TU6) model defined by the COST 207 project for GSM [22], which was shown to give an accurate description of the DVB-T mobile radio channel in [8]. All simulations were performed on channel CH40 (626 MHz) with a bandwidth of 8 MHz. Note that analog TV channels span the frequencies between 400 and 790 MHz, so that the considered carrier is roughly in the middle of the broadcast spectrum, which will be shared between analog and digital transmission.

We show the performance of the receiver architectures described in Table I, where the number 2 in the abbreviation denotes the use of two antennas, with maximum ratio combining (MRC).

A. BER Performance

Iterative ICI cancellation. For the ideal (ID) scheme, Fig. 3 shows the bit error rate (BER) at the output of the ICI cancellation scheme, as a function of the speed for various numbers

TABLE I
RECEIVER ARCHITECTURES

Rec. name	Channel estimation	ICI canc.
STD	linear pilot (2x8, time x frequency), [8]	–
2STD	pilot advanced [8]	–
IMI (2IMI)	multistage	iterative
ML	ML estimator	iter., $N_I = 1$
ID	ideal	iterative

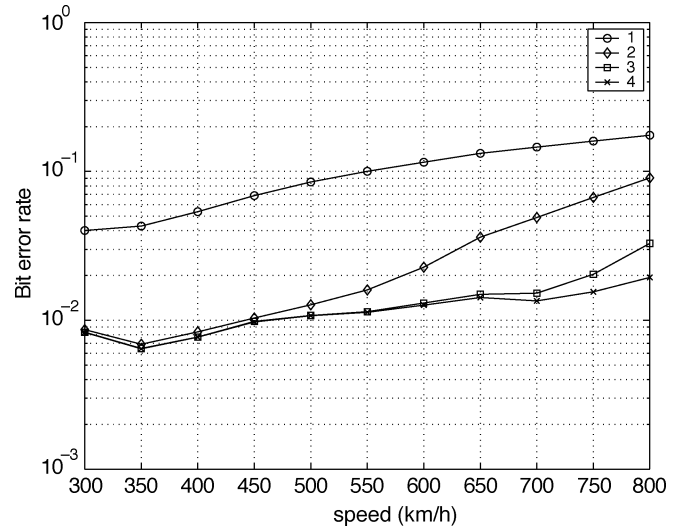


Fig. 3. BER as a function of the speed for iteration $N_I = 1, \dots, 4$ for the ID scheme.

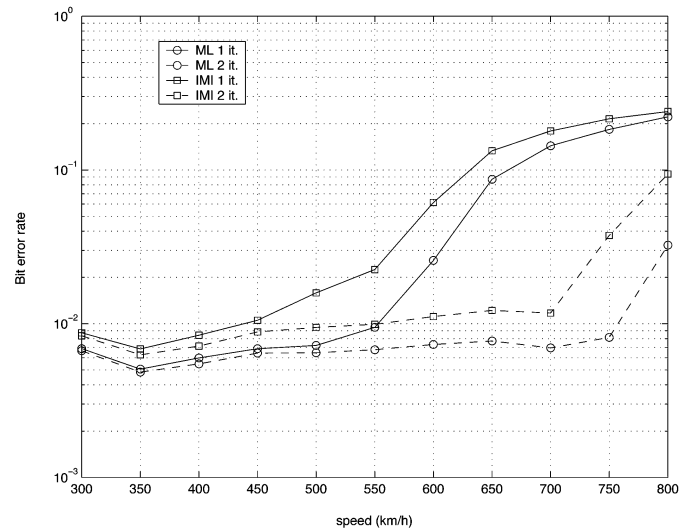


Fig. 4. BER as a function of the speed for two methods of estimation of the higher derivatives parameters: ML and IMI scheme, for $N_I = 1$ (solid lines) and $N_I = 2$ (dashed lines).

of iterations of the ICI cancellation loop. The number of derivatives of ICI cancellation scheme is $P = 2$. We observe that there is a significant improvement if one iterates beyond $N_I = 1$, i.e., with $N_I = 2$ and $N_I = 3$. Further iterations do not significantly enhance the performance.

Higher order derivative estimation. Fig. 4 shows the BER as a function of the speed for the ML and iterative multistage implementation (IMI) scheme, for iterations $N_I = 1$ and 2. The DVB-T mode is 8k with code rate 1/2 and a constellation of 64-QAM. We observe that the IMI scheme yields a reduction of

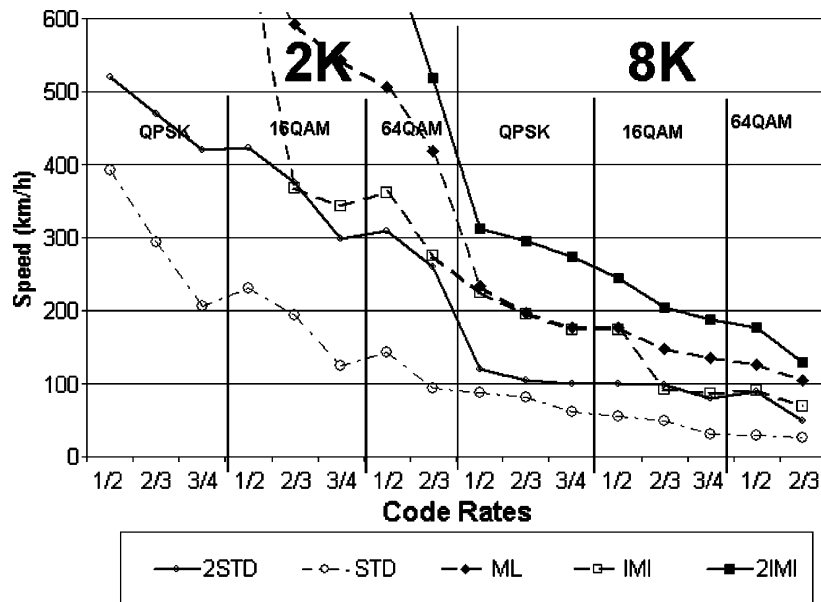


Fig. 5. Maximum achievable speed for various DVB-T modes, constellation sizes, code rates, and receiver techniques.

the achievable speed of about 50 km/h, compared to the performance of the ML scheme.

B. Maximum Achievable Speed

A relevant measure of performance of a DVB-T receiver is the maximum speed at which the reception of the television signal is perceptually satisfying. Laboratory experiments [8] allowed the derivation of each error correcting code, a target bit error rate for the bit stream at the input of the Viterbi decoder that yields good reception for DVB-T. From these results, we obtained the maximum speed for various receiver architectures, and we set as target BERs $7 \cdot 10^{-3}$, $4 \cdot 10^{-3}$, and $2 \cdot 10^{-3}$ for code rates 1/2, 2/3, and 3/4, respectively.

Fig. 5 shows the maximum speed for various architectures, constellation sizes, code rates, and FFT sizes. The standard receiver apparently achieves good performance for the 2k mode, as well as for the lower constellations sizes of the 8k mode. The 8k mode with 64-QAM gives the maximum bit rate (150 Mbit/s) but it requires more sophisticated ICI compensation techniques to achieve speeds beyond 100 km/h. We note that the multistage estimator with advanced zero-derivative estimate has a performance that is similar to the ML estimator. We also see that the use of multiple antennas improves performance, both for the standard (STD) scheme and for the IMI scheme, due to the additional space diversity gain achieved by the receiver.

Comparing the performance of the different receiver structures, we can conclude that the combination of two antennas and IMI provides the best performance, since it benefits from both an appropriate channel model and the space diversity. A good tradeoff between complexity and performance is achieved by the IMI scheme with a single antenna. In this case, speeds of up to approximately 100 km/h are achievable, without the requirement of multiple antennas.

TABLE II
IMPLEMENTATION COMPLEXITY OF VARIOUS ARCHITECTURES

	# Antenna	Complexity (# FFT)	Storage (# OFDM blocks)	Eq. Silicon (# SA)
STD	1	2	5	9
2STD	2	4	50	60
ML	1	1700	7	4300
IMI	1	15	10	50

C. Implementation Complexity

Complexity is evaluated in terms of silicon area needed for the implementation of the algorithms, including both the required storage and the computational complexity. As unit measure, we consider the silicon area required for the storage of one OFDM symbol (1 SA). For the computation of one FFT, which requires $N \log_2(N)$ complex multiplications, an area of 2.5 SA was considered. For the inversion of a $L \times L$ matrix, we assumed a complexity of L^3 complex multiplications. Based on evidence from practical DVB chip implementations, decoding and re-encoding have been accounted for one FFT.

The implementation complexity of various schemes is shown in Table II. We observe that by employing the multistage estimator scheme, a significant drop in complexity is achieved, compared to the ML estimator. Even in this case, the impact of the complexity of the channel estimation is still substantial, and further improvements in chip design are needed before this technique can be actually implemented. Fig. 5 and Table II show that STD does not guarantee the appropriate performance for speeds above 60 km/h. Therefore, they are not suitable for the reception of the DVB-T signal on freeways outside the city center. When one combines multiple antennas and advanced channel estimation, the main contribution to the silicon area is due to the extra storage needed for the longer time interpolation. The combination of ICI cancellation and receiver diversity requires some more signal processing for the MRC, but it merges the advantages of the two architectures.

VII. CONCLUSION

A new iterative interference cancellation scheme for the reception of an OFDM signal in a mobile environment has been introduced. In our scheme, the interference is canceled iteratively, whereby at every stage the reliability of the data increases. A critical part of the scheme is the estimation of the channel parameters. Using a multistage scheme, estimation is reduced to a set of independent problems with a lower overall complexity. The resulting architecture allows the reception of high rate DVB-T signals at speeds of about 100 km/h. The use of the iterative ICI cancellation scheme, together with the multistage estimator on a receiver equipped with multiple antennas, allows the reception of the signal at higher speeds.

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