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MMSE Estimation of Time-Varying Channels for DVB-T Systems with Strong Co-Channel Interference*

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ABSTRACT

We present a minimum mean-square error (MMSE) channel estimator for a multi-antenna DVB-T receiver in the presence of strong co-channel interference. Based on the scattered pilot symbols contained in the DVB-T transmit signal, our method estimates the time-varying vector channel in an off-line, blockwise fashion. An implementation in the time-delay domain is used to reduce computations and enhance estimation performance. We also propose algorithms for estimating the channel statistics required for the design of the channel estimator. Simulation results show that in the case of strong co-channel interference, the proposed channel estimator achieves a significant performance improvement over a reference method.

1 Introduction

Terrestrial digital video broadcasting (DVB-T) [1–3] is an OFDM-based [4] communications scheme that is currently being deployed in several European countries as a successor to analog terrestrial television systems [5]. DVB-T is furthermore being considered for multimedia content provision via dense cellular unicast networks [6, 7]. Since strong co-channel interference will occur in these cellular networks, network operators will have to analyze and resolve service-limiting interference situations. For this purpose, a prototype measurement device employing an antenna array and off-line signal processing is being developed in the framework of the IST project ANTIUM. To gather information about the strength and origin of interfering co-channel DVB-T signals, this device will have to separate the signals of DVB-T transmitters with highly unequal power levels and to decode their cell identifications that are contained in the transmission parameter signalling (TPS) data stream [3]. To this end, the receiver signal processing comprises synchronization (discussed in [8]), channel estimation (considered here), and decoding.

Several different MMSE estimators of time-varying channels that exploit the scattered pilot symbols contained in the DVB-T transmit signal have been proposed previously (e.g., [9–11]). The channel estimator presented here differs from previous methods in that it is based on block processing, accurately estimates the second-order statistics of the time-varying channel, and uses an implementation in the time-delay domain. Together, these features allow to eliminate a substantial part of the interference and thus yield good performance even in the case of strong co-channel interference.

This paper is organized as follows. After a description of the system model in Section 2, the MMSE channel estimator is developed in Section 3. Estimators for the prior knowledge required (channel delay and Doppler profile, interference/noise variance) are proposed in Section 4. Finally, Section 5 provides simulation results.

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2 System Model

Transmitters. We consider an equivalent discrete-time baseband system with I interfering DVB-T transmitters. The symbols of the ith transmitter are denoted $a_i[n,k]$, where $n \in \mathbb{Z}$ is the OFDM symbol (time) index and $k \in [0, K-1]$ is the subcarrier (frequency) index. Of the K subcarriers, only those with $k \in [K_{\min}, K_{\max}]$ are used to transmit actual data [3]. Furthermore, scattered pilot symbols (a BPSK encoded pseudo-random binary sequence) are transmitted at the locations [3]

$$\mathscr{P} = \big\{(n,k) \,\big|\, k = K_{\min} + 3\,(n+n_0) \bmod 4 + 12\,p,\;\; p \in [0, P-1]\big\}.$$

Here, $P = \lfloor (K_{\max} - K_{\min})/12 \rfloor$ is the number of scattered pilot symbols at each n, and the fixed parameter $n_0 \in \{0, \ldots, 3\}$ accounts for the four possible locations of the scattered pilots. We assume n_0 to be known in the following.

The nth OFDM symbol (in the signal domain) of the ith transmitter is the inverse discrete Fourier transform (IDFT) of the $a_i[n,k]$, preceded by a cyclic prefix of length L_{cp},

$$s_n^{(i)}[m] = \begin{cases} \frac{1}{\sqrt{K}} \sum\limits_{k=0}^{K-1} a_i[n,k]\, e^{j2\pi mk/K}, & m = -L_{\text{cp}}, \ldots, K-1, \\ 0, & \text{else.} \end{cases}$$

Thus, the OFDM symbol duration is $N_{\text{s}} = K + L_{\text{cp}}$. The overall transmit signal is $s_i[m] = \sum_{n=-\infty}^{\infty} s_n^{(i)}[m - nN_{\text{s}}]$.

Receiver. For a receiver with an M-element antenna array, the received signal vector is given by

$$\mathbf{y}[m] = \sum_{i=0}^{I-1} \sum_{l=0}^{L_i} \mathbf{h}_i[m,l]\, s_i[m-l-\eta_i] + \mathbf{w}[m]. \tag{1}$$

Here, $\mathbf{h}_i[m,l] = \big[h_i^{(1)}[m,l], \ldots, h_i^{(M)}[m,l]\big]^T$, where $h_i^{(j)}[m,l]$ denotes the impulse response of the random time-varying wireless channel between transmitter i and receive antenna j. Furthermore, $L_i \le L_{\text{cp}}$ is the maximum delay of $\mathbf{h}_i[m,l]$, η_i is a time offset (note that the transmitters are not synchronous), and $\mathbf{w}[m] = \big[w_1[m], \ldots, w_M[m]\big]^T$ is zero-mean, temporally and spatially white, Gaussian noise. We assume $|\eta_i - \eta_{i'}| \ge L_{\text{cp}}$ for $i \ne i'$ since otherwise the interfering OFDM signals cannot be separated.

After synchronizing to the ith transmitter (i.e., determination of η_i), the receiver discards the cyclic prefix and demodulates the received signal $\mathbf{y}[m]$ by means of a DFT,

$$\mathbf{x}_i[n,k] = \frac{1}{\sqrt{K}} \sum_{m=0}^{K-1} \mathbf{y}[nN_{\text{s}} + \eta_i + m]\, e^{-j2\pi km/K}.$$

Assuming that the channel impulse response $\mathbf{h}_i[m,l]$ varies negligibly within a symbol period, one obtains

$$\mathbf{x}_i[n,k] = \mathbf{H}_i[n,k]\, a_i[n,k] + \mathbf{z}_{\text{I}}^{(i)}[n,k] + \mathbf{z}_{\text{N}}^{(i)}[n,k], \tag{2}$$

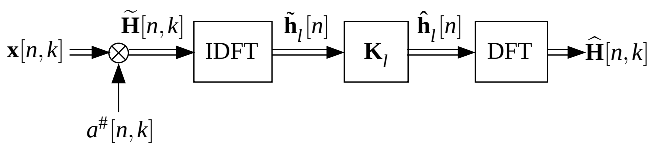


Figure 1: *Block diagram of the proposed channel estimator.*

with the channel coefficients

$$\mathbf{H}_i[n,k] = \sum_{l=0}^{L_{\text{cp}}} \mathbf{h}_i[nN_\text{s}+\eta_i, l]\, e^{-j2\pi kl/K}, \tag{3}$$

the co-channel interference

$$\mathbf{z}_\text{I}^{(i)}[n,k] = \frac{1}{\sqrt{K}} \sum_{m=0}^{K-1} \sum_{\substack{i'=0 \\ i'\neq i}}^{I-1} \sum_{l=0}^{L_{\text{cp}}} \mathbf{h}_{i'}[nN_\text{s}+\eta_i+m, l] \cdot s_{i'}[nN_\text{s}+m-l+\eta_i-\eta_{i'}]\, e^{-j2\pi km/K}, \tag{4}$$

and the noise

$$\mathbf{z}_\text{N}^{(i)}[n,k] = \frac{1}{\sqrt{K}} \sum_{m=0}^{K-1} \mathbf{w}[nN_\text{s}+\eta_i+m]\, e^{-j2\pi km/K}.$$

Channel statistics. In what follows, we assume that the time-varying channels associated to different transmitters and receive antennas are mutually uncorrelated and satisfy the wide-sense stationary uncorrelated scattering (WSSUS) assumption [12, 13], i.e.,

$$\mathrm{E}\{\mathbf{h}_i[m'+m,l]\,\mathbf{h}_{i'}^H[m',l']\} = R_i[m,l]\,\delta[l-l']\delta[i-i']\,\mathbf{I}.$$

Note that the time-delay correlation function $R_i[m,l]$ is assumed to be equal for all antennas. We model $R_i[m,l]$ as being separable,

$$R_i[m,l] = r_i[m]\,P_i[l], \tag{5}$$

with time correlation function $r_i[m]$ and delay profile $P_i[l]$. The channel's Doppler profile is given by [12, 13]

$$S_i(\nu) = \sum_{m=-\infty}^{\infty} r_i[m]\,e^{-j2\pi\nu m}, \tag{6}$$

where ν denotes the normalized Doppler frequency.

3 MMSE Channel Estimation

We next consider estimation of the channel coefficients $\mathbf{H}_i[n,k]$ for a given i corresponding to one of the I transmitters (the respective time offset η_i will be set equal to 0 for simplicity). For convenience of notation, we will suppress the index i in what follows. A block diagram of the proposed channel estimator is shown in Fig. 1. Due to the off-line, blockwise processing performed by the ANTIUM receiver, the demodulated sequence $\mathbf{x}[n,k]$ in (2) is available for $(n,k) \in \mathscr{Q} \triangleq [0,N-1]\times[K_\text{min},K_\text{max}]$, where N is some block length. According to Fig. 1, our method first compensates for the scattered pilot symbols and then performs MMSE estimation of $\mathbf{H}[n,k]$ using a DFT-based implementation.

Division by pilot symbols. The scattered pilot symbols $a[n,k]$ on the right-hand side of (2) are known at the receiver for $(n,k)\in\mathscr{P}$. Their effect can be removed by calculating

$$\widetilde{\mathbf{H}}[n,k] = \mathbf{x}[n,k]\,a^{\#}[n,k], \tag{7}$$

with

$$a^{\#}[n,k] \triangleq \begin{cases} 1/a[n,k], & (n,k)\in\mathscr{P}, \\ 0, & \text{else}. \end{cases}$$

Inserting (2) into (7), we obtain

$$\widetilde{\mathbf{H}}[n,k] = \begin{cases} \mathbf{H}[n,k] + \mathbf{z}[n,k], & (n,k)\in\mathscr{P}, \\ 0, & \text{else}, \end{cases} \tag{8}$$

where $\mathbf{z}[n,k] \triangleq \mathbf{z}_\text{I}[n,k]/a[n,k] + \mathbf{z}_\text{N}[n,k]/a[n,k]$ corresponds to the interfering transmitters and the noise. We will model $\mathbf{z}[n,k]$ as being white, which can be justified as follows. First, $\mathbf{z}_\text{N}[n,k]$ is white, so $\mathbf{z}_\text{N}[n,k]/a[n,k]$ is white, too. Second, $\mathbf{z}_\text{I}[n,k]/a[n,k]$ can also be modeled as white since the scattered pilot symbols are a pseudo-random binary sequence (note, however, that the $\mathbf{z}_\text{I}[n,k]$ in (4) might be highly correlated).

MMSE channel estimator. According to (8), MMSE channel estimation based on $\widetilde{\mathbf{H}}[n,k]$ corresponds to suppression of the interference/noise $\mathbf{z}[n,k]$ for $(n,k)\in\mathscr{P}$ and interpolation of the missing channel coefficients $\mathbf{H}[n,k]$, $(n,k)\notin\mathscr{P}$.

To obtain an efficient implementation of the MMSE estimator as well as reliable estimation of the channel statistics (see Section 4), we first apply a (scaled) IDFT to (8). This yields

$$\tilde{\mathbf{h}}_l[n] = \frac{1}{\alpha}\sum_{k=0}^{K-1} \widetilde{\mathbf{H}}[n,k]\,e^{j2\pi lk/K}, \qquad (n,l)\in\mathscr{U}, \tag{9}$$

where $\alpha = \sqrt{K(K_\text{max}-K_\text{min})}/12$ is chosen to preserve energy and $\mathscr{U} = [0,N-1]\times[0,L_\text{cp}]$. Due to the subsampling in (8) (every 12th subcarrier contains a scattered pilot symbol), the $\tilde{\mathbf{h}}_l[n]$ in (9) are periodic in l with period $K/12$. However, aliasing is avoided since the channel's maximum delay was assumed to satisfy $L \le L_\text{cp} < K/12$. Therefore, (9) can be written as

$$\tilde{\mathbf{h}}_l[n] = \mathbf{h}_l[n] + \tilde{\mathbf{z}}_l[n], \tag{10}$$

where

$$\mathbf{h}_l[n] \triangleq \mathbf{h}[nN_\text{s}, l] \tag{11}$$

is the subsampled impulse response of the channel (cf. (3)) and $\tilde{\mathbf{z}}_l[n] \triangleq \frac{1}{\alpha}\sum_{k=0}^{K-1}\mathbf{z}[n,k]\,e^{j2\pi lk/K}$ is white noise.

Since both $\mathbf{h}_l[n]$ and $\tilde{\mathbf{z}}_l[n]$ are uncorrelated for different delays l, the $\mathbf{h}_l[n]$ can be *separately* estimated from the $\tilde{\mathbf{h}}_l[n]$ according to

$$\left[\hat{\mathbf{h}}_l[0],\ldots,\hat{\mathbf{h}}_l[N-1]\right] = \left[\tilde{\mathbf{h}}_l[0],\ldots,\tilde{\mathbf{h}}_l[N-1]\right]\mathbf{K}_l, \tag{12}$$

with $l = 0,\ldots,L_\text{cp}$. Here, the $\mathbf{K}_l$ are estimator coefficient matrices of size $N\times N$ that will be discussed presently. From the estimates $\hat{\mathbf{h}}_l[n]$, the desired estimates of the channel coefficients $\mathbf{H}[n,k]$ are finally obtained according to (3):

$$\widehat{\mathbf{H}}[n,k] = \sum_{l=0}^{L_\text{cp}} \hat{\mathbf{h}}_l[n]\,e^{-j2\pi kl/K}, \qquad (n,k)\in\mathscr{Q}.$$

We finally determine the matrices $\mathbf{K}_l$ minimizing the MSE

$$\varepsilon \triangleq \frac{1}{M}\mathrm{E}\left\{\left\|\mathbf{H}[n,k]-\widehat{\mathbf{H}}[n,k]\right\|^2\right\} = \frac{1}{M}\sum_{l=0}^{L}\mathrm{E}\left\{\left\|\mathbf{h}_l[n]-\hat{\mathbf{h}}_l[n]\right\|^2\right\}. \tag{13}$$

Inserting (12) and (10) into (13) and using the orthogonality principle [14], it can be shown that the coefficient matrices of the MMSE estimator are given by

$$\mathbf{K}_{l,\text{opt}} = P[l]\left(P[l]\,\mathbf{R}+\sigma_z^2\,\mathbf{I}\right)^{-1}\mathbf{R}. \tag{14}$$

Here, the $N\times N$ correlation matrix $\mathbf{R}$ is Hermitian Toeplitz with first row $[r[0], r[N_\text{s}],\ldots,r[(N-1)N_\text{s}]]$ (recall that $r[m]$ and $P[l]$ were defined in (5)) and σ_z^2 denotes the variance of $\tilde{\mathbf{z}}_l[n]$ in (10).

Discussion. From (14), is seen that the estimator coefficient matrices $\mathbf{K}_{l,\text{opt}}$—and, thus, also the estimates $\hat{\mathbf{h}}_l[n]$ in (12)—are nonzero only for those delays l where the delay profile $P[l]$ is nonzero. Hence, for a channel with maximum delay L, the $\hat{\mathbf{h}}_l[n]$ are nonzero only for $l\in[0,L]$. The resulting nulling of the interference associated to the remaining values of l can be shown to produce a reduction of the interference/noise level by about a factor of $K/(12L)$.

For a small channel length L, this interference/noise reduction is substantial (e.g., for $L=20$ we obtain $10\log_{10}K/(12L)\approx 15\,\text{dB}$). Of course, this interference/noise reduction presupposes knowledge of the channel's delay profile $P[l]$ (see next).

4 Estimation of Channel and Noise Statistics

Calculation of the optimal coefficient matrices $\mathbf{K}_{l,\text{opt}}$ according to (14) requires knowledge of the channel's delay profile $P[l]$ and time correlation function $r[m]$ (or, equivalently, Doppler profile $S(\nu)$, cf. (6)), as well as knowledge of the interference/noise variance σ_z^2. The estimation of these quantities is considered next.

Delay profile. Estimating the delay profile $P[l]$ amounts to estimating the powers of the stationary processes $\mathbf{h}_l[n]$, $l=0,\ldots,L_{\text{cp}}$ by properly averaging $\tilde{\mathbf{h}}_l[n]$ in (10). In what follows, let $h_l^{(j)}[n]=h^{(j)}[nN_s,l]$ $(j=1,2,\ldots,M)$ denote the elements of the vector $\mathbf{h}_l[n]$ in (11). Let us choose B such that BN_s is less than the channel's coherence time (in our simulations, we obtained good results with $B=10$). Then $h_l^{(j)}[n]$ varies negligibly within intervals $n\in[qB,(q+1)B-1]$, and we can thus perform the preliminary "coherent" averaging

$$\bar{h}_l^{(j)}[q] = \frac{1}{B}\sum_{n=0}^{B-1}\tilde{h}_l^{(j)}[qB+n]\,, \qquad q=0,\ldots,\left\lfloor\frac{N}{B}\right\rfloor-1\,.$$

Here, $\tilde{h}_l^{(j)}[n]$ is the jth element of the vector $\tilde{\mathbf{h}}_l[n]$. We have $\bar{h}_l^{(j)}[q]\approx h_l^{(j)}[qB]+\bar{z}_l^{(j)}[q]$, where $\bar{z}_l^{(j)}[q]$ is the block-average of the jth element of $\tilde{\mathbf{z}}_l[n]$. This coherent averaging results in the suppression of a large part of the interference/noise.

Next, an initial estimate of the delay profile is obtained by averaging the power of $\bar{h}_l^{(j)}[q]$ over all q and all antennas:

$$\tilde{P}[l] = \frac{1}{M\left\lfloor\frac{N}{B}\right\rfloor}\sum_{j=1}^{M}\sum_{q=0}^{\lfloor\frac{N}{B}\rfloor-1}\big|\bar{h}_l^{(j)}[q]\big|^2.$$

Due to interference and noise, $\tilde{P}[l]$ will be nonzero even for delays l where $P[l]=0$. We thus calculate the final estimate of the delay profile by thresholding the initial estimate $\tilde{P}[l]$,

$$\hat{P}[l] = \begin{cases}\tilde{P}[l]\,, & \text{if } \tilde{P}[l]\geq\gamma\bar{P},\\ 0\,, & \text{elsewhere.}\end{cases} \tag{15}$$

Here, the parameter $\gamma>0$ adjusts the threshold about the mean $\bar{P}=\frac{1}{L_{\text{cp}}+1}\sum_{l=0}^{L_{\text{cp}}}\tilde{P}[l]$ of $\tilde{P}[l]$. (We obtained good results for $\gamma=1.1$.)

Doppler profile. We estimate the Doppler profile $S(\nu)$ in (6) by calculating the periodogram of $\tilde{h}_l^{(j)}[n]$ with respect to n and averaging/summing over all delays and antennas. Using a Doppler resolution of $\nu_0=1/(\kappa N)$, $\kappa\in\mathbb{N}$, this yields the initial estimate

$$\tilde{S}(u\nu_0) = \frac{1}{M\kappa}\sum_{j=1}^{M}\sum_{l=0}^{L_{\text{cp}}}\left|\frac{1}{N}\sum_{n=0}^{N-1}\tilde{h}_l^{(j)}[n]\,e^{-j2\pi u\nu_0 n}\right|^2. \tag{16}$$

The function $\tilde{S}(u\nu_0)$ is calculated for $u\in[-V,V]$, with $V\leq\kappa N$, which corresponds to the normalized Doppler frequencies $\nu=0,\pm\nu_0,\ldots,\pm V\nu_0$. (The factor $1/\kappa$ in (16) guarantees that $\sum_{u=-V}^{V}\tilde{S}(u\nu_0)$ is the same for all κ.) As an example, consider a DVB-T system in 8K mode with symbol duration $T_s=924\,\mu\text{s}$ [3]. With $N=204$, $\kappa=10$, and $V=200$, it can be shown that (16) estimates the Doppler profile in the range $\pm 106\,\text{Hz}$ with a resolution of $\nu_0/T_s=0.53\,\text{Hz}$.

Again, the interference/noise will cause $\tilde{S}(u\nu_0)$ to be nonzero even for those u where $S(u\nu_0)=0$. Thus, the final estimate of the Doppler profile is obtained by thresholding,

$$\hat{S}(u\nu_0) = \begin{cases}\tilde{S}(u\nu_0)\,, & \text{if } \tilde{S}(u\nu_0)\geq\zeta\,\bar{S},\\ 0\,, & \text{elsewhere,}\end{cases} \tag{17}$$

where the parameter $\zeta>0$ adjusts the threshold about the mean $\bar{S}=\frac{1}{2V+1}\sum_{u=-V}^{V}\tilde{S}(u\nu_0)$ of $\tilde{S}(u\nu_0)$. (We obtained good results for $\zeta=1.5$.) Finally, an estimate of the time correlation function $r[m]$ involved in (14) is computed from $\hat{S}(u\nu_0)$ in (17) by means of an IDFT (cf. (6)). The parameter κ in $\nu_0=1/(\kappa N)$ must be chosen large enough so that aliasing errors due to the discretization $\nu=u\nu_0$ are sufficiently small.

Interference/noise variance. An estimate of the interference/noise variance σ_z^2 is obtained as a by-product of the thresholding operation (15). Since the values of $\tilde{P}[l]$ that are less than $\gamma\bar{P}$ are attributed to the interference/noise, a simple estimator of σ_z^2 is given by

$$\widehat{\sigma_z^2} = \frac{1}{|\mathscr{L}|}\sum_{l\in\mathscr{L}}\tilde{P}[l]\,,$$

where $\mathscr{L}$ is the set of indices $l\in[0,L_{\text{cp}}]$ for which $\tilde{P}[l]<\gamma\bar{P}$ (or, equivalently, $\hat{P}[l]=0$) and $|\mathscr{L}|$ is the number of such indices.

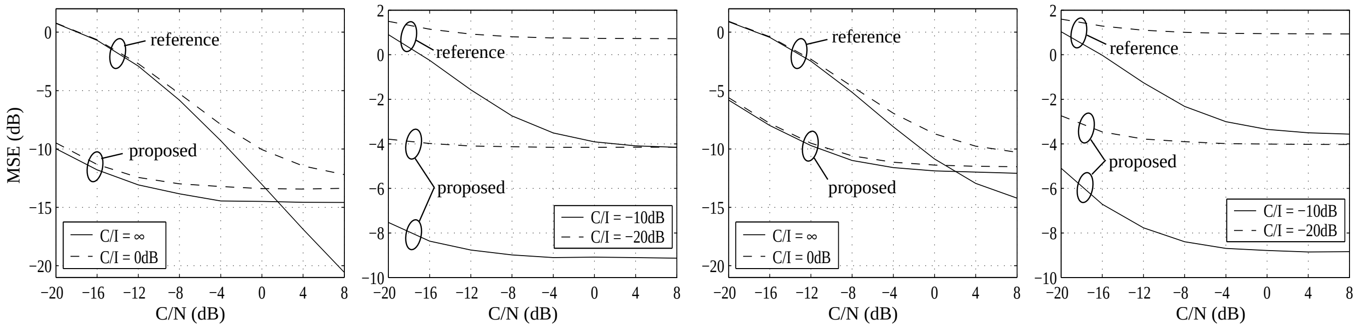
5 Simulation Results

We simulated a scenario with $I=3$ DVB-T transmitters, each with carrier frequency 625 MHz. The transmit signals were consistent with the 8K mode of [3]. Here, 6817 out of $K=8192$ subcarriers are used ($K_{\min}=0$, $K_{\max}=6816$), corresponding to transmit bandwidths of 8 MHz. The cyclic prefix length was $1/32$ of the "useful" symbol duration (i.e., the symbol duration minus the cyclic prefix length). The transmission was nonhierarchical with a uniform 64-QAM signal constellation. The receiver used a uniform circular antenna array with $M=5$ antennas. It was synchronized to the first transmitter (corresponding to $i=0$) and the time offsets in (1) were chosen as $0\,\mu\text{s}$, $109\,\mu\text{s}$, and $328\,\mu\text{s}$ for $i=0$, 1, and 2, respectively. A signal block containing $N=204$ OFDM symbols was recorded.

The channels corresponding to the three transmitters were synthesized using a sum-of-sinusoids fading channel simulator [12] with identical parameters for all channels. We randomly picked 20 propagation paths with 10 subpaths each, resulting in an exponential delay profile with a maximum delay of $5\,\mu\text{s}$. The angles of incidence at the receiver and the subpath phases were uniformly distributed in $[0,2\pi]$. We simulated both time-invariant channels (no Doppler) and channels with a maximum Doppler frequency of 42 Hz (corresponding to a vehicular velocity of 20 m/s), which is quite large for a DVB-T system. The received signal was the sum of the three channel outputs and temporally and spatially white noise.

We applied the channel estimator described in Sections 3 and 4 to the estimation of the channel corresponding to the first transmitter. This estimation was performed for various values of the transmit and noise powers. The resulting MSE (13) (obtained empirically by averaging over 2040 OFDM symbols and 10 channel realizations) is shown in Fig. 2 versus the carrier-to-noise ratio C/N $\triangleq$ $\mathrm{E}\{\|\mathbf{H}_0[n,k]\,a_0[n,k]\|^2\}/\mathrm{E}\{\|\mathbf{z}_{\mathrm{N}}^{(0)}[n,k]\|^2\}$, for several different carrier-to-interference ratios C/I $\triangleq$ $\mathrm{E}\{\|\mathbf{H}_0[n,k]\,a_0[n,k]\|^2\}/\mathrm{E}\{\|\mathbf{z}_{\mathrm{I}}^{(0)}[n,k]\|^2\}$ (cf. (2)). For comparison, the results obtained with a reference channel estimator [15] are also shown. This reference estimator uses linear interpolation in the time direction and an MMSE estimator/interpolator with 145 filter coefficients in the frequency direction.

Time-invariant channels. Fig. 2(a) shows the estimation MSE obtained for time-invariant channels with no interference (C/I $=\infty$) and weak interference (C/I $=0$ dB). In this case, the performance of our estimator is seen to be mostly noise-limited, i.e., the MSE



(a) (b) (c) (d)

Figure 2: *MSE of channel estimate vs. C/N in the presence of two interfering DVB-T transmitters: (a) Time-invariant channels with high C/I, (b) time-invariant channels with low C/I, (c) time-varying channels with high C/I, and (d) time-varying channels with low C/I.*

is strongly dependent on C/N (it degrades from about -15 dB and -13 dB for C/N $= 8$ dB to about -10 dB for C/N $= -20$ dB) but only weakly dependent on C/I. The reference estimator is seen to be noise-limited, too; however, for C/N below 0 dB it performs much worse than our estimator. It is seen that the reference estimator has a performance advantage only for the case of no interference (C/I $= \infty$) and large C/N. The poorer performance of our estimator in this case is due to a systematic error that is introduced by the DFT-based implementation using (9). Indeed, whereas inversion of (3) requires the channel coefficients $\mathbf{H}_i[n,k]$ to be available for $k = 0,\ldots,K-1$, the DVB-T system uses only the subcarriers $k = K_{\min},\ldots,K_{\max}$. This causes (9) to be only an approximate inversion of (3) even in the noise-free case. However, the resulting performance degradation is noticeable only in the case of high C/I and C/N.

The results obtained for the time-invariant case but strong interference (C/I $= -10$ dB and C/I $= -20$ dB) are presented in Fig. 2(b). Here, the performance of our channel estimator is seen to be mostly interference-limited, i.e., the MSE varies only slightly with C/N (in particular, for C/I $= -20$ dB the MSE is about -4 dB for all C/N levels). Furthermore, the MSE is noticeably larger than for weak interference. However, again our channel estimator considerably outperforms the reference channel estimator whose MSE becomes unacceptable for strong interference and noise.

Time-varying channels. Fig. 2(c) shows the MSE obtained for time-varying channels (maximum Doppler frequency 42 Hz) with no interference (C/I $= \infty$) and weak interference (C/I $= 0$ dB). The general behavior is seen to be similar to the time-invariant case (cf. Fig. 2(a)). Again, the performance of both our estimator and the reference estimator is noise-limited. The channels' time-variation causes the performance to degrade with respect to the time-invariant case. This is due to the shorter coherence time of the time-varying channel, which causes our estimator to average over smaller time durations. Again, our estimator outperforms the reference estimator except for the case where C/I $= \infty$ and C/N > 0 dB.

Finally, the results for the time-varying case and strong interference (C/I $= -10$ dB and C/I $= -20$ dB) are provided in Fig. 2(d). The performance of both estimators is slightly worse than in the time-invariant case although still predominantly interference-limited. In contrast to the reference estimator, our estimator shows acceptable performance even for high C/I and C/N levels.

6 Conclusions

We proposed an MMSE estimator for time-varying channels within a DVB-T system with potentially strong co-channel interference. The estimator operates in a blockwise manner, using an efficient implementation in the time-delay domain. Computer simulations showed that for strong interference and noise levels, our channel estimator features significantly better performance than a reference estimator. The high interference/noise immunity of our channel estimator is a result of the time-delay domain implementation (which allows to eliminate a substantial part of the interference/noise) and the accurate estimation of channel statistics.

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