An Adaptive Two-Dimensional Channel Estimator for Wireless OFDM with Application to Mobile DVB-T

Frieder Sanzi and Joachim Speidel

***Abstract*—In this paper an adaptive channel estimator is proposed and investigated to improve the performance of the receiver for pilot aided wireless and mobile OFDM systems. The estimator consists of a two-dimensional Wiener filter which is implemented as a cascade of two one-dimensional filters. We propose an efficient algorithm for adaptation to time varying channels of the second filter. The method is applied to the Terrestrial Digital Video Broadcasting System (DVB-T), which was originally specified for fixed receivers and which is in the process of being extended for mobile reception. The results are shown after inner decoding. Depending on the channel conditions, the signal-to-noise ratio gain can be up to 1.6 dB. The method provides a compatible improvement to DVB-T receivers.**

***Index Terms*—DVB-T, channel estimation, OFDM.**

I. Introduction

ORTHOGONAL Frequency Division Multiplexing (OFDM) has become an important modulation scheme for several applications, like Digital Subscriber Line (DSL), Digital Audio Broadcasting (DAB) and Terrestrial Digital Video Broadcasting (DVB-T). In the discussion on the specification of the next generation wireless mobile communication system (International Mobil Telecommunication IMT 2000, Universal Mobile Telecommunication System UMTS), OFDM was also a strong candidate.

The transfer function $H(f,t)$ of a wireless mobile, multipath channel varies not only with frequency f but also with time t. At the receiver which is using coherent demodulation the channel characteristic has to be equalized. To do so, a channel estimator is required. In OFDM systems, often pilots are inserted into the transmission signal to provide the estimator with knowledge about the channel. This is done e.g. with the DVB-T system.

Normally the estimation is carried out by two cascaded orthogonal one-dimensional filters which are fixed. According to the sampling theorem the first filtering will be executed in time direction and the second in frequency direction in order to estimate channels like the DAB-Hilly Terrain II profile with long delay spreads. In this paper we propose a channel estimation in which the second filter is adaptive. We describe an algorithm which calculates the actual delay spread of the channel and with this result we improve the second filtering in frequency direction.

We apply this method to a DVB-T receiver to improve its performance for mobile reception. As is well known, the DVB-T standard was designed primarily for fixed reception [1]. However, mobile reception of TV signals has become an important aspect for the future to compete against digital satellite and cable as well as the DAB system. The DAB system with some enhancements has already proven its ability to also transmit compressed video with about 1.5 Mbit/s to the receiver, which can move at very high speeds of more than 200 km/h [2], [3]. So, it is a great challenge to improve the DVB-T receiver for fast mobile reception.

In the DVB-T system the lowest possible bitrate is 5 Mbit/s. In this case QPSK is used in addition with strong channel coding. For 16-QAM with maximum coding redundancy the bitrate will be doubled (10 Mbit/s). In contrast to the DAB system nondifferential modulation is used with DVB-T. In this case channel estimation is an important task for a mobile receiver.

This paper is organized as follows: Section II provides an overview of the DVB-T system. In Section III the system model is explained in detail. The two-dimensional sampling of the channel frequency response is presented in Section IV. Section V describes the channel estimation and the adaptive filtering. The computer simulation results are shown in Section VI. Section VII concludes the paper.

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II. DVB-T System

The DVB-T system is designed for transmitting an MPEG-2 transport stream with fixed packet length [1]. This stream is fed into the DVB-T transmitter shown in Fig. 1. First, the energy dispersal scrambles long series of zeros or ones. The energy dispersal is followed by a concatenated channel coding. It consists of an outer Reed-Solomon code, an outer convolutional interleaver, an inner convolutional code and an inner interleaver. The output of the inner coder can be punctered to adjust different code rates $R = 1/2, 2/3, 3/4, 5/6$ or $7/8$.

The system uses OFDM. The number of sub-carriers can either be $2k$ or $8k$ $(k = 1024)$. The sub-carriers are modulated either with QPSK, 16-QAM or 64-QAM. Not all sub-carriers are used to transmit information. Some sub-carriers operate as pilots, which broadcast information known to the receiver. These pilots are used for channel estimation. Some other sub-carriers contain information about the transmission scheme, i.e. channel coding and modulation (TPS transmission parameter signaling).

The OFDM is followed by the guard interval insertion. The guard interval is a cylic prefix which contains a copy of the last part of the OFDM symbol. The relative length Δ of the guard interval, which is the ratio between guard interval length T_g and OFDM symbol duration T_u, can take on several values. These are $\Delta = 1/4, 1/8, 1/16$ or $1/32$. The guard interval

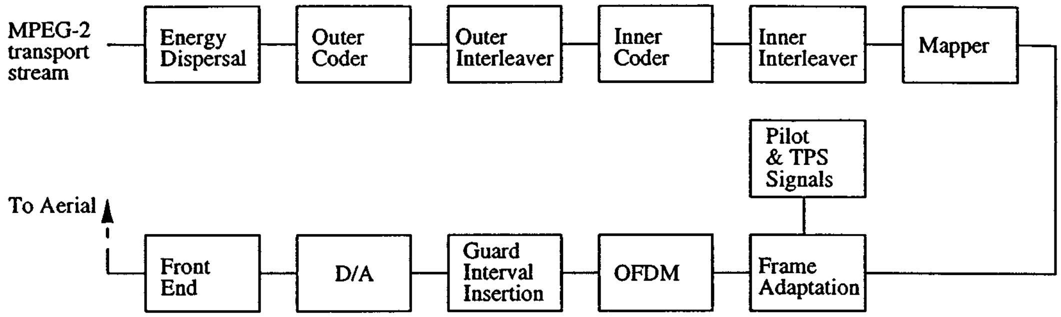

Fig. 1. Block diagram of DVB-T transmitter (nonhierarchical mode).

eliminates intersymbol interference (ISI) between consecutive OFDM symbols if the guard interval is longer than the delay spread of the channel. We assume that the channel characteristic does not change significantly during one OFDM symbol. In this case the guard interval preserves the orthogonality of the sub-carriers at the receiver.

The bitrate R_b which can be achieved with this system in case of nonhierarchical transmission mode is given by

$$R_b = \frac{N_u v R_{\mathrm{RSC}} R}{T_u(1+\Delta)} \tag{1}$$

whereby N_u is the number of useful sub-carriers with $N_u = 1512$ in $2k$-mode or $N_u = 6048$ in $8k$-mode, v is the number of bits per sub-carrier, R_{RSC} is the code rate of the Reed-Solomon code with $R_{\mathrm{RSC}} = 188/204$, R is the code rate of the inner convolutional code, T_u is the OFDM symbol duration with $T_u = 224\ \mu$s in $2k$-mode or $T_u = 896\ \mu$s in $8k$-mode and Δ is the relative guard interval length. v depends on the selected modulation. After inserting these values into Eq. (1) we obtain

$$R_b = \frac{423vR}{68(1+\Delta)} \text{ Mbit/s.} \tag{2}$$

In case of hierarchical transmission mode the bitrate will be a sum of two terms similar to Eq. (2).

III. System Modelling

For the mobile channel we use the wide-sense stationary uncorrelated scattering (WSSUS) channel model introduced in [4]. The frequency response of the channel can be expressed as

$$H(f,t) = \lim_{N\to\infty} \frac{1}{\sqrt{N}} \sum_{n=1}^{N} e^{j(\phi_n + 2\pi f_{D_n} t - 2\pi f \tau_n)}, \tag{3}$$

where ϕ_n is the phase, f_{D_n} the Doppler frequency and τ_n the delay of the nth path. The ϕ_n, f_{D_n} and τ_n are randomly choosen depending on the corresponding joint probaility density function $p_{\phi,f_D,\tau}(\phi, f_D, \tau)$ of the considered channel model [5], [6].

In our computer simulation environment we can model five different channel profiles which are listed in Table I. These are the four GSM profiles specified in COST 207 [7], and one channel model specified in EUREKA 147, see [8] and [9].

For our further considerations we assume the channel characteristic to be approximately unchanged during the duration of one OFDM symbol plus its guard interval. Under this assumption and provided that the guard interval is longer than the delay spread of the channel the cyclic prefix avoids inter-carrier interference (ICI) and also ISI. In this case we can compute the received constellation points as

TABLE I
Channel Models

channel/model		type	τ_{max}	reference
P_1	GSM	rural area no LOS	$0.7\mu s$	[7]
P_2	GSM	urban area	$7.0\mu s$	[7]
P_3	GSM	hilly urban area	$10.0\mu s$	[7]
P_4	GSM	hilly terrain	$20.0\mu s$	[7]
P_6	DAB	hilly terrain II	$45.0\mu s$	

$$Y_{k,l} = H_{k,l} X_{k,l} + N_{k,l} \tag{4}$$

whereby l is the OFDM symbol index, k is the subcarrier index, $X_{k,l}$ are the transmitted signal constellation points and $N_{k,l}$ are independent and identically distributed complex Gaussian noise variables. The $H_{k,l}$ are sample values of the channel frequency response.

$$H_{k,l} = H(k\Delta f, lT_s) \tag{5}$$

whereby $T_s = T_u + T_g$ is the duration of an OFDM symbol plus the guard interval and Δf is the sub-carrier distance with $\Delta f = 1/T_u$. In [10] an analysis of ICI is given.

The receiver has to minimize the Bit Error Ratio (BER) after inner decoding. For decoding we use the MAP algorithm. Beause the MAP decoder expects Log-likelihood-values (L-values, [11]), the demapper has to produce them with respect to the channel frequency response and noise. The channel estimator at the receiver has to find a good estimate of the $H_{k,l}$. Note that the channel frequency response is only known at pilot positions with

$$\hat{H}_{k_P,l_P} = \frac{Y_{k_P,l_P}}{X_{k_P,l_P}} \tag{6}$$

whereby k_P and l_P describe a pilot position and X_{k_P,l_P} is the transmitted pilot signal which is known at the receiver. The remaining $H_{k,l}$ have to be estimated by means of interpolation based on the known $\hat{H}_{k_P,l_P}$.

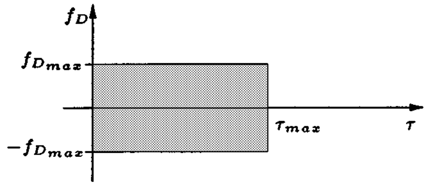

Fig. 2. Top view of the delay doppler spread function $|S(\tau, f_D)|$.

IV. 2-Dimensional Sampling of the Channel Frequency Response

At the receiver the channel frequency response is known at the pilot positions as outlined in the last section. So the channel frequency response is sampled at the receiver. In this section we first consider the 2-dimensional sampling theorem before we proceed to the channel estimation in the next section.

The delay doppler spread function $S(\tau, f_D)$ of the channel can be obtained from the time variant frequency response $H(f,t)$ by applying the Fourier transforms.

$$S(\tau, f_D) = \int_{-\infty}^{+\infty} \int_{-\infty}^{+\infty} H(f,t)\, e^{j2\pi\tau f}\, e^{-j2\pi f_D t}\, df\, dt \quad (7)$$

$S(\tau, f_D)$ can be considered to be bandlimited with respect to f_D by the maximal Doppler frequency $f_{D\ \max}$ and with respect to τ by the maximal delay $\tau_{\max}$. In principle, $|S(\tau, f_D)|$ is depicted in Fig. 2. Outside the dotted area $|S(\tau, f_D)|$ is very small. So, $|S(\tau, f_D)|$ can be approximated as a lowpass filter with cut-off frequencies $\tau_{\max}$ and $f_{D\ \max}$.

For a periodical sampling grid the sampled time variant frequency response $H_A(f,t)$ is given by

$$H_A(f,t) = H(f,t) \sum_{n=-\infty}^{+\infty} \sum_{k=-\infty}^{+\infty} \delta(f - ns_{11}\Delta f - ks_{12}\Delta f, t - ns_{21}T_s - ks_{22}T_s). \quad (8)$$

The sampling grid is defined by s_{11}, s_{12}, s_{21} and s_{22} as well as Δf and T_s. A consise treatment of multidimensional sampling can be found e.g. in [12].

By applying the 2-dimensional Fourier-Transform to $H_A(f,t)$ in Eq. (8), we obtain the corresponding delay doppler spread function [13]

$$S_A(\tau, f_D) = \frac{1}{|s|\Delta f T_s} \sum_{n=-\infty}^{+\infty} \sum_{k=-\infty}^{+\infty} S\left(\tau - \frac{ns_{22}}{s\Delta f} - \frac{ks_{21}}{s\Delta f}, f_D - \frac{ns_{12}}{sT_s} - \frac{ks_{11}}{sT_s}\right), \quad (9)$$

with

$$s = s_{11}s_{22} - s_{12}s_{21}. \quad (10)$$

In the DVB-T system the scattered pilots are used for channel estimation. In our notation the sampling grid defined by the positions of the scattered pilots is achieved for $s_{11} = 3, s_{12} = 6$, $s_{21} = 1$and $s_{22} = -2$. In this case we obtain the following delay doppler spread function from Eq. (9).

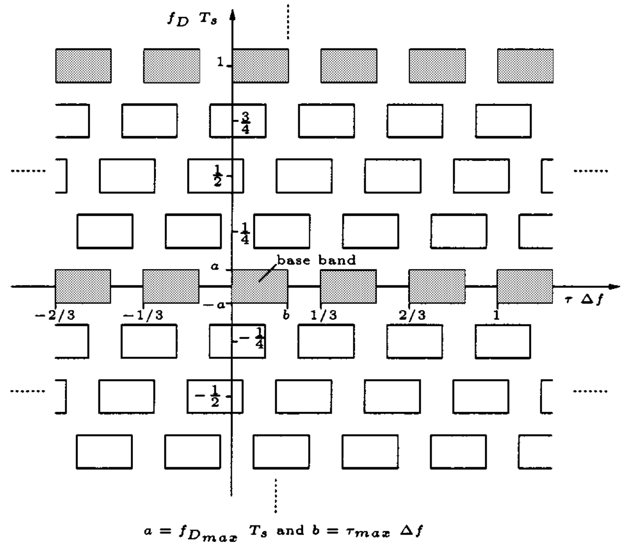

Fig. 3. Top view of the delay doppler spread function $|S'_A(\tau, f_D)|$.

$$S'_A(\tau, f_D) = \frac{1}{12\Delta f T_s} \sum_{n=-\infty}^{+\infty} \sum_{k=-\infty}^{+\infty} S\left(\tau - \frac{n}{6\Delta f} + \frac{k}{12\Delta f}, f_D + \frac{n}{2T_s} + \frac{k}{4T_s}\right) \quad (11)$$

If we insert $|S(\tau, f_D)|$ given in Fig. 2 into Eq. (11), we obtain the delay doppler spread function $|S'_A(\tau, f_D)|$ as shown in Fig. 3. Obviously, the base band spectrum $|S(\tau, f_D)|$ of Fig. 2 is repeated with a two-dimensional structure as given by Eq. (11). In Fig. 3 we have assumed a channel with a large delay spread $\tau_{\max}$. The spectra in this figure do not overlap (no aliasing) in f_D-direction if $a < 1/8$, which means that the maximal Doppler frequency $f_{D\ \max}$ has to be

$$f_{D\ \max} < \frac{1}{8T_s}. \quad (12)$$

There is no spectral overlap in τ-direction, if $b < 1/3$ which results in

$$\tau_{\max} < \frac{1}{3\,\Delta f} = \frac{T_u}{3}. \quad (13)$$

The two-dimensional sampling theorem is also considered in [14] and [15].

It is obvious that in the case of a long delay spread according to Fig. 3 Wiener filtering (see next section) has to be done in time direction first. After that filtering only the gray spectra in Fig. 3 will remain. The second filtering will be done in frequency direction. According to Fig. 4 only the gray spectra have to be kept.

V. Channel Estimation

For channel estimation the Wiener filtering is introduced in [5], [6] and [15]. Because the covariance of the channel is not known at the receiver rectangular shaped probability density

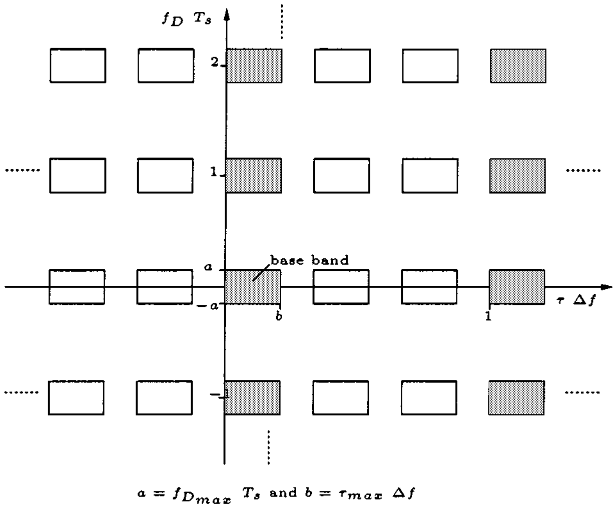

Fig. 4. Top view of $|S'_A(\tau, f_D)|$ after filtering in time-direction.

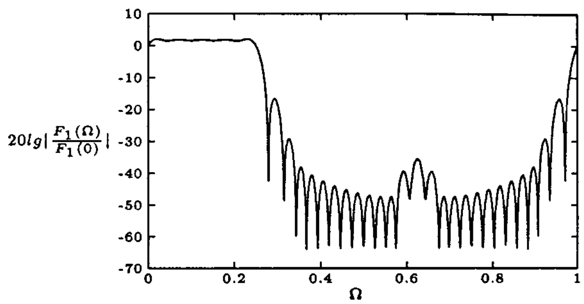

Fig. 5. Delay respones $|F_1(\Omega)|$ of the Wiener filter.

functions are assumed. Mismatch analysis between interpolation filter and actual channel statistics is given in [16].

For filtering in time direction we use a Wiener filter with two taps and in frequency direction one with 12 taps. For the filtering in time direction we design the filter with the assumption of $f_{D\ \max} = 200$ Hz (f_D uniformly distributed between $-f_{D_{\max}}$ and $f_{D_{\max}}$) and a maximal guard interval of $T_u/4 = 56\ \mu$s ($2k$ mode). In frequency direction we have assumed a maximal delay equal to the guard interval duration T_g (τ uniformly distributed between 0 and T_g). This approach is taken as a reference for our further considerations. Whereas in [8] and [9] both filter characteristics are fixed for a transmission with a certain guard interval T_g, we propose an adaption of the bandwith of the second filter to the channel. As will be shown in the following, this will significantly improve receiver performance.

Let us assume that the mobile receiver is in an area, where the channel can be modeled by P_6 in Table I. Obviously, this channel exhibits maximal delay spread of $\tau_{\max} = 45\ \mu$s. In this area, the transmitter operates with a guard interval $T_u/4 = 56\ \mu$s in $2k$ mode to prevent ISI and ICI. The delay response $F_1(\Omega)$ of the Wiener filter which estimates in frequency direction is shown in Fig. 5 with $\Omega = \tau\Delta f$ and $E_s/N_0 = 10$ dB where E_s is the average energy per data symbol and N_0 is the one-sided noise spectral density. The bandwidth Ω_g of this filter is

$$\Omega_g = T_g\Delta f = 56\ \mu\text{s}\frac{1}{224\ \mu\text{s}} = 0.25. \tag{14}$$

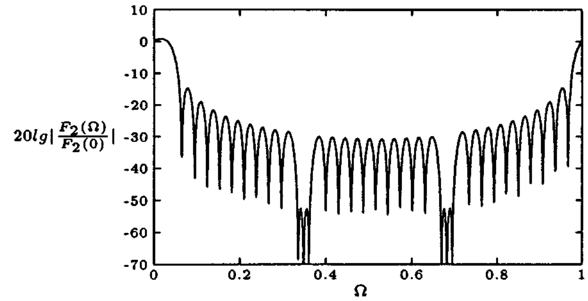

Fig. 6. Delay respones $|F_2(\Omega)|$ of the Wiener filter.

When the mobile receiver moves into an area with smaller delay spread, where e.g. the channel models P_1 and P_2 in Table I are applicable, the bandwith of the filter $F_1(\Omega)$ is by about factor 8 to large. As a consequence, the noise impact on the estimation result is high. To prevent from this effect, the filter bandwidth has to be switched. For channel models P_1 and P_2 a guard interval at the receiver of $\tau_g = 7\mu$s is sufficiant. With this, the Wiener filter with delay response $|F_2(\Omega)|$, depicted in Fig. 6, results ($E_s/N_0 = 10$ dB). The bandwidth of $F_2(\Omega)$ is smaller than that of $F_1(\Omega)$. As a consequence the noise of the estimated channel samples is reduced. So, if the filter bandwidth is adapted to the channel, noise can be kept low and thus channel estimation will be improved.

For adaptive filter selection, we propose an algorithm which determines the shortest possible guard interval length for the actual channel. In one OFDM symbol duration the channel is estimated every third carrier after filtering in time direction. With these samples of the channel frequency response we calculate the channel impulse response with an IFFT of order k ($2k$ mode). We obtain the channel impulse response sampled with $T_A = 1/(3\ 1024\Delta f)$. In a next step we compute the energy E_i in every segment of length 7 μs (96 samples). Note that the duration of the shortest guard interval is 7 μs. After this we consider the first n segments and search for the segment with the maximal energy $E_{\max}$. The number of segments n is calculated according to

$$n = \frac{T_g}{7\ \mu\text{s}}. \tag{15}$$

Additionally, we compute the relative energy ΔE_i of each segment

$$\Delta E_i = \frac{E_i}{E_{\max}} \tag{16}$$

where $i \in \{1, 2, \ldots, n\}$. Note that $n = 2^x$ with $x \in \{0, 1, 2, 3\}$. All relative energies which are lower than a certain bound are set to zero. With the remaining segments we calculate the shortest possible guard interval length $\tilde{T}_g$ according to the actual channel. For this we search for the relative energy $\Delta E_{\tilde{i}} \neq 0$ with the highest $\tilde{i}$. After this we calculate $\tilde{x}$ according to

$$\tilde{x} = \lceil \log_2(\tilde{i}) \rceil, \tag{17}$$

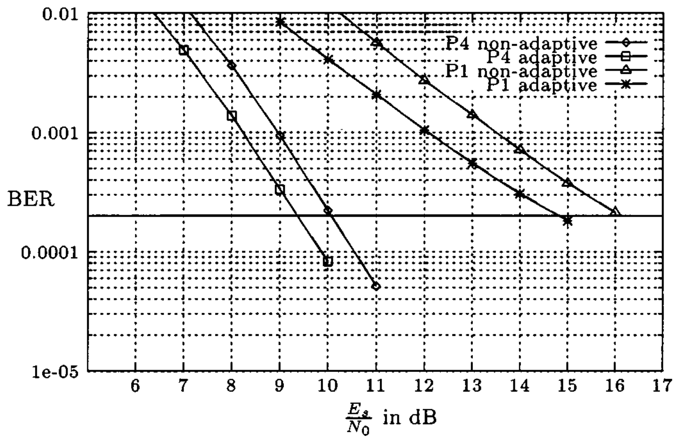

Fig. 7. BER comparision for nonadaptive and adaptive filtering for channel P_1 and P_4, as well as $f_{D_{\max}} = 53$ Hz.

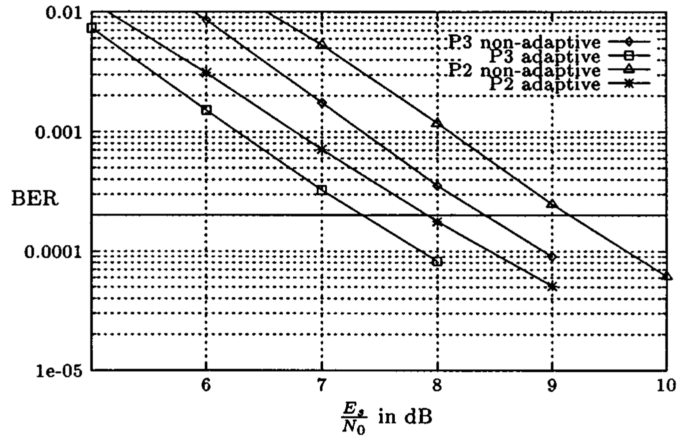

Fig. 8. BER comparision for nonadaptive and adaptive filtering for channel P_2 and P_3, as well as $f_{D_{\max}} = 53$ Hz.

whereby $\lceil y \rceil$ denotes the smallest integer number larger or equal to y. With the obtained $\tilde{x}$ the shortest possible guard interval length $\tilde{T}_g$ is calculated according to

$$\tilde{T}_g = \frac{2^{\tilde{x}}}{n} T_g. \tag{18}$$

With the obtained guard interval length $\tilde{T}_g$ the filtering in frequency direction is done.

Normally, a DVB-T receiver is equipped with a set of filter coefficients for different guard intervals, namely $T_g = 2^i 7\ \mu$s ($i \in \{0, 1, 2, 3\}$). A simple implementation of the filter adaption can be done by selecting an appropriate filter out of the given set. The performance of the receiver is further increased, if the filter bandwidth is adapted closer to the delay spread of the actual channel.

In the next section according to the bandwidth reduction for the second filter the improvement for BER after inner decoding is shown.

VI. Simulation Results

We have simulated a complete DVB-T chain on a computer. For simulation the BER after inner decoding is measured. We chose QPSK with coderate $R = 1/2$ for transmission and we take the $2k$ mode with a maximal guard interval duration of $T_g = 56\ \mu$s into account. The maximal Doppler frequency is $f_{D_{\max}} = 53$ Hz (later $f_{D_{\max}} = 193$ Hz).

In Fig. 7 the BER for the channels P_1 and P_4 are shown. Adaptive filtering is compared to conventional nonadaptive filtering. The BER after inner decoding should be equal or lower than $2\,10^{-4}$. In this case the signal will be almost error free after outer decoding. The improvement for the channel P_1 is 1.3 dB, which is bigger than for P_4 (0.7 dB), because the delay spread is shorter (second filtering with lower bandwidth).

The BER for the channels P_2 and P_3 are given in Fig. 8. The gain of the adaptive filtering for channel P_2 is 1.3 dB and for channel P_3 1.1 dB. According to the previous result (Fig. 7), P_2 has a shorter delay spread than channel P_3.

For the channel P_6 no improvement is possible, because the channel exhibits maximal delay $\tau_{\max} = 45\ \mu$s. In this case the transmitted guard interval is $T_g = 56\ \mu$s. A reduction at the receiver is not possible because the next smaller guard interval is $T_g = 28\ \mu$s according to [1]. The filter according to this guard interval has a bandwidth which is too small to estimate the channel correctly.

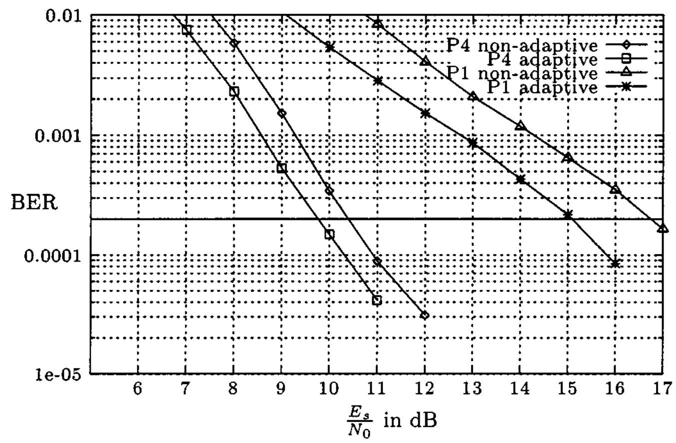

Fig. 9. BER comparision for nonadaptive and adaptive filtering for channel P_1 and P_4, as well as $f_{D_{\max}} = 193$ Hz.

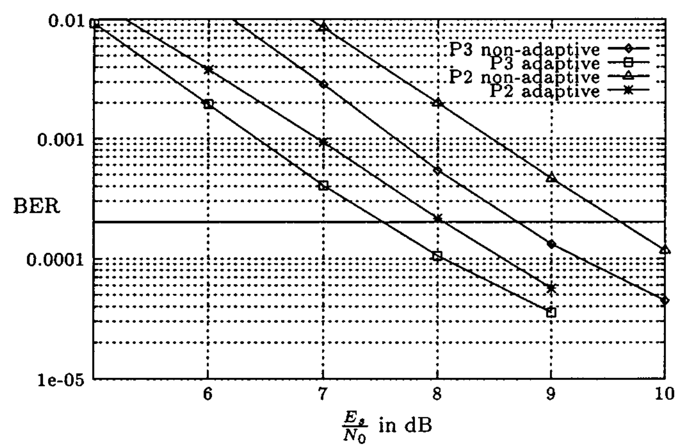

Fig. 10. BER comparision for nonadaptive and adaptive filtering for channel P_2 and P_3, as well as $f_{D_{\max}} = 193$ Hz.

In Figs. 9 and 10 the results are presented for a Doppler frequency of $f_{D_{\max}} = 193$ Hz. The improvement of the adaptive filtering compared to conventional filtering can be up to 1.6 dB (P_1 and P_2). As can be seen from our simulations the higher the Doppler frequency, the larger the E_s/N_0 gain is.

VII. Conclusion

In this paper we have proposed an adaptive algorithm for estimation of multipath, time varying channels for OFDM mobile transmission systems which use pilot frequencies like the DVB-T system. Each carrier is QPSK modulated and the code rate is $1/2$. As a result of computer simulations the BER after inner decoding was presented. The improvement of signal to noise ratio (E_s/N_0) for this adaptive filtering compared to conventional filtering can be up to 1.6 dB depending on the channel characteristic. The smaller the delay spread of the channel the higher the gain is.

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