

Memory truncation and crosstalk cancellation for efficient Viterbi detection in FDMA systems

Alfred Mertins

Abstract — In this paper, the design of optimal receive filter banks for frequency division multiple access (FDMA) over frequency selective channels is investigated. A new design strategy based on the principle of memory truncation, rather than equalization, is presented. Through the receive filters, each subchannel is truncated to a pre-defined length, and the final data recovery is carried out via low complexity Viterbi detectors. Both closed form designs and adaptive techniques are discussed. Design examples are presented for high speed transmission over copper wires. The examples show that memory truncation allows significant performance improvements over the often used minimum mean squared error (MMSE) equalization.

Keywords — *transmultiplexer, memory truncation, Viterbi detector, dispersive channel, filter banks.*

1. Introduction

The performance of transmission systems based on discrete multitone (DMT) modulation [1–3] or orthogonal frequency division multiplex (OFDM) [4] degrades rapidly when the length of the channel impulse response exceeds the length of the guard interval, which is introduced to cope with non-ideal channels. As a result of an insufficient guard interval, intersymbol interference (ISI) and inter-channel interference (ICI, crosstalk) will occur. To cope with longer channel impulse responses one can increase the length of the guard interval, but this will decrease the efficiency, as less data symbols can be transmitted. Increasing both the length of the guard interval and the number of subchannels allows one to maintain a desired bandwidth efficiency, but this strategy also has its limits. For example, the delay between transmitter and receiver may become unacceptably high. Also, the hardware requirements increase with an increasing number of subchannels. Finally, channels which can be regarded as slowly time-varying when the number of subchannels is low may turn into fast time-varying ones if the number of subchannels and thus the lengths of the transmit and receive filters are significantly increased.

In this paper, new methods for the design of optimal receive filter banks in multichannel transmission systems are proposed. The techniques are presented for a multirate filter bank framework, which gives a common description of

a variety of transmission techniques [5]. The solutions apply to DMT [3, 6], OFDM [4], coded-OFDM [7], transmultiplexers [8, 9], and other transmission techniques where the transmit signal is created as a weighted linear combination of basis sequences with the data symbols being the weights. Even code division multiple access (CDMA) [10, 11] can be seen as a multirate filter bank. Figure 1 shows the general structure of the transmit/channel/receive model used in this paper. Depending on the actual modulation technique (DMT, OFDM, CDMA, etc.), the upsampling factor N , the number of subchannels, M , and the impulse responses $g_k(n)$ and $h_k(n)$ are chosen.

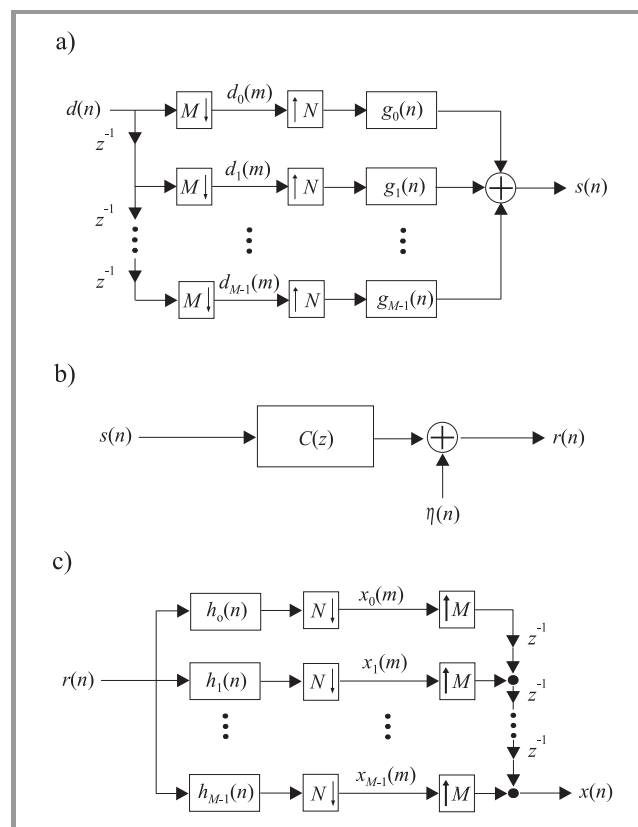


Fig. 1. Multirate discrete-time transmitter/channel/receiver model: (a) transmitter; (b) channel; (c) receiver.

Classically, the system in Fig. 1 is used to convert time-division multiplexed (TDM) data signals into frequency

division multiplexed (FDM) signals and vice versa. The synthesis filter bank in Fig. 1a then provides the TDM-to-FDM conversion. The FDM signal is transmitted through the channel, and the receive signal is finally fed into the analysis filter bank in Fig. 1c which converts the FDM signal back into a TDM one [8, 9]. Examples are DMT and OFDM, which both use block wise DFT's to do the TDM-to-FDM and FDM-to-TDM conversions. For DMT and OFDM the impulse responses of the filters in Fig. 1 are the complex exponentials occurring in the DFT, with maximum filter length N . A better frequency selectivity can be obtained if one uses longer filters, designed according to multirate filter bank theory. In [12] a comparison between cosine-modulated filter banks and DMT was given which shows that filter banks offer greater potential than block transforms.

Various solutions to the problem of reducing ISI through channel equalization have been proposed [5, 10, 12 – 19]. Most of them are based on minimizing the mean squared error (MSE) between the sent data and the equalizer output, either using a general MSE or a zero forcing (ZF) concept. Decision feedback equalizers (DFE) have been considered in [17]. Minimum mean squared error and ZF solutions with a joint design of receiver and transmitter have been proposed in [5, 16, 18]. Such a joint design can be useful in cases where communication takes place in both directions. In this paper, we concentrate on the optimal receiver design, thus addressing cases where the transmitter is fixed. The methods proposed in this paper are extensions of the technique in [19] to the design of entire receive filter banks for the critically sampled and oversampled cases (i.e. $N \geq M$). Furthermore, methods for adaptive receiver design are presented. The design criterion is based on the idea of memory truncation [20, 21], where the receiver does not try to fully equalize the channel and leaves a residual system in the data path. In the optimum, the MSE between the equalizer output and a filtered version of the input data sequence is minimized. The final data detection then takes place via a Viterbi detector which needs to consider only the residual impulse responses. The lengths of the residual filters can be chosen arbitrarily and will typically be a few taps, thus allowing the use of low-complexity Viterbi detectors. The advantage of memory truncation over equalization is that critical channel zeros (e.g. zeros close to or even on the unit circle) need not be equalized, so that the problem of noise amplification through the equalizer can be avoided.

Note that for DMT transmission, memory truncation has also been proposed in a different form where the channel memory is shortened to the length of the guard interval prior to the DFT analysis in the receiver [22, 23]. In the present paper, however, memory truncation is incorporated as a property of the receive filters, and we can even treat cases where no guard interval is introduced at all.

The paper is organized as follows. In Section 2 the input-output relations for the multirate system in Fig. 1 are discussed. Section 3 addresses the design of optimal receive filter banks. Methods for adaptive receiver design are pre-

sented in Section 4. Results are discussed in Section 5, and finally, conclusions are given in Section 6.

Notation: The superscript T denotes transposition of a vector or matrix. The superscripts $*$ and H denote complex conjugation and conjugate transposition ($\mathbf{r}^H = [\mathbf{r}^*]^T$), respectively. \mathbf{I} is an identity matrix of appropriate size. $E\{\}$ denotes the expectation operation, and $\delta_{i,k}$ is the Kronecker symbol.

2. Input-output relations

We consider the system in Fig. 1. The sequences $d_k(m)$, $k = 0, 1, \dots, M-1$ are created through a series-to-parallel conversion of a single data sequence $d(m)$ in the form $d_k(m) = d(mM - k)$, $k = 0, 1, \dots, M-1$. In other words, they are polyphase components of the sequence $d(n)$. In the next step, the data sequences $d_k(m)$ are upsampled by a factor of N and then fed into the M respective synthesis filters with impulse responses $g_k(n)$, $k = 0, 1, \dots, M-1$. The sum of the filtered signals finally forms the transmit signal

$$s(n) = \sum_{k=0}^{M-1} \sum_{m=-\infty}^{\infty} d_k(m) g_k(n - mN). \quad (1)$$

Typically, the filters $g_k(n)$ are chosen to be frequency selective, so that each data sequence $d_k(m)$ is transmitted in a distinct frequency band.

To make certain that the input data can be recovered at least theoretically from the transmit signal $s(n)$, the upsampling factor N must be chosen such that $N \geq M$ [9]. In many practical systems $N > M$ is used, which means that the transmitter introduces redundancy. This redundancy can be utilized in the receiver for enhancing the performance in the presence of frequency selective channels.

For the discussion in this paper, the transmission channel is assumed to be time-invariant. However, since adaptive methods for the receiver design are proposed the channel may, in practice, even be slowly time varying with respect to the filter lengths involved. Considering a time invariant channel, the receive signal is given by

$$r(n) = \left[\sum_{m=-\infty}^{\infty} c(m) s(n - m) \right] + \eta(n), \quad (2)$$

where $\eta(n)$ is an additive, data independent noise process and $c(n)$ is the channel impulse response. The noise is assumed to be zero mean and wide-sense stationary.

On the receiver side, the signal $r(n)$ is fed into the analysis filter bank, as shown in Fig. 1, and the filter output signals are subsampled by a factor of N to form the final output signals

$$x_k(m) = \sum_{n=0}^{L_h-1} h_k(n) r(mN - n), \quad (3)$$

$$k = 0, 1, \dots, M-1.$$

In Eq. (3), L_h is the length of the receive filters. Combining Eqs. (1), (2) and (3) we get the input-output relation

$$x_k(m) = \sum_{n=0}^{L_h-1} \sum_{\mu=-\infty}^{\infty} \sum_{i=0}^{M-1} \sum_{\ell=-\infty}^{\infty} h_k(n) c(\mu) d_i(\ell) \times \\ \times g_i(mN-n-\mu-\ell N) + \sum_{n=0}^{L_h-1} h_k(n) \eta(mN-n), \quad (4) \\ k = 0, 1, \dots, M-1.$$

Under ideal conditions where the analysis and synthesis filters of the transmission system form a perfect reconstruction (PR) filter bank and where the channel is noise free and ideal (i.e. $\eta(n) = 0 \forall n$ and $c(n) = \delta_{n,0}$) the transmit/receive system allows us to recover the data $d_k(m)$ without error:

$$x_k(m) = d_k(m - m_0). \quad (5)$$

The term m_0 is the overall delay of the system. The PR conditions for the filter bank itself are

$$\sum_{n=0}^{L_h-1} h_k(n) g_i(mN-n) = \delta_{i,k} \delta_{m,m_0} \quad (6)$$

with $i, k = 0, \dots, M-1$. A practical problem is that even transmitter/receiver systems satisfying Eq. (6) will be unable to perfectly recover the data if a non-ideal channel is introduced. Thus, the channel should be taken into account when designing the receive filter bank. Methods for this will be discussed in the next section. Since the channel is usually not known a priori in practice, adaptation rules will be presented in Section 4.

3. Design of optimal receive filter banks

In this section, we derive methods for the design of optimal receive filter banks. For this we define an error signal as the difference between the receiver output signals $x_k(m)$ and filtered versions of the data sequences:

$$e_k(m) = \sum_{j=0}^{L_h-1} h_k(j) r(mN-j) + \\ - \sum_{i=0}^{L_p-1} p_k(i) d_k(m-m_0-i), \quad (7) \\ k = 0, \dots, M-1.$$

The optimality criteria for the design of the M receive filters are the MSE's given by

$$Q_k = E\{|e_k(m)|^2\}, \quad k = 0, \dots, M-1, \quad (8)$$

which are to be minimized under the energy constraints

$$\sum_{n=0}^{L_p-1} |p_k(n)|^2 = 1, \quad k = 0, 1, \dots, M-1. \quad (9)$$

The constraints (9) are needed to avoid the trivial solution $h_k(n) = 0, p_k(n) = 0$.

Note that the error measure (7) is different from the MSE as defined for conventional MSE equalizers [10, 12 – 15]. The idea behind the proposed approach is to truncate the channel memory and not to delete it completely. The impulse responses $p_k(m)$ are to be understood as residual impulse responses of arbitrarily chosen length L_p . Both the optimal residual systems $p_k(m)$ and receive filters $h_k(n)$ need to be found through minimization of Eq. (8).

Because of the existence of residual systems $p_k(m)$, minimizing Eq. (8) does, in general, not result in an equalization of the channel. Even if $Q_k = 0$ there will be a remaining ISI between L_p consecutive data samples in each of the subchannels. The crosstalk between different channels $i \neq k$ will be reduced as much as possible with FIR filters of the given length L_h .

With analysis filters designed through the minimization of Eq. (8) the overall system can be modeled with little error as a set of M independent channels with

$$x_k(m) = \left[\sum_{i=0}^{L_p-1} p_k(i) d_k(m-m_0-i) \right] + \eta'_k(m), \quad (10) \\ k = 0, \dots, M-1.$$

The modified noise processes $\eta'_k(m)$ contains the filtered and subsampled original noise and all modeling errors made by simplifying the real system to the form (10).

To recover the data, the signals $x_k(m)$, $k = 0, \dots, M-1$ are fed into M independently operating Viterbi detectors which have to consider the respective channels $p_k(m)$, $k = 0, \dots, M-1$. Since the lengths of these channels are chosen arbitrarily, one can choose lengths which result in a manageable computational cost for the Viterbi detectors while maintaining a low noise variance at the detector inputs. Clearly, the longer the systems $p_k(m)$ are, the smaller the modeling errors in (10) and thus the smaller the variances $E\{|\eta'_k(m)|^2\}$ are. For $L_p = 1$ the Viterbi detectors degenerate to simple threshold detectors, at the expense of an increased noise variance compared to cases where $L_p > 1$.

Note that in the special case of $L_p = 1$, Eq. (8) states a standard MSE criterion, and the optimized analysis filters $h_k(n)$ can be regarded as MMSE equalizers. Then the proposed solution becomes equivalent to other known MMSE solutions [10, 12 – 15].

To obtain a compact formulation of the objective function, we now introduce the following vectors:

$$\mathbf{h}_k = [h_k(0), \dots, h_k(L_h-1)]^T, \quad (11)$$

$$\tilde{\mathbf{r}}(m) = [r(mN), \dots, r(mN-L_h+1)]^T, \quad (12)$$

$$\mathbf{p}_k = [p_k(0), \dots, p_k(L_p-1)]^T, \quad (13)$$

$$\mathbf{d}_k(m) = [d_k(m), \dots, d_k(m-L_p+1)]^T. \quad (14)$$

We get

$$e_k(m) = \tilde{\mathbf{r}}^T(m) \mathbf{h}_k - \mathbf{d}_k^T(m) \mathbf{p}_k. \quad (15)$$

Using this notation the cost functions (8) can finally be written as

$$Q_k = \mathbf{h}_k^H \mathbf{R}_{rr} \mathbf{h}_k - \mathbf{h}_k^H \mathbf{R}_{rd}^{(k)} \mathbf{p}_k - \mathbf{p}_k^H \mathbf{R}_{dr}^{(k)} \mathbf{h}_k + \mathbf{p}_k^H \mathbf{R}_{dd}^{(k)} \mathbf{p}_k \quad (16)$$

with

$$\mathbf{R}_{rr} = E \{ \mathbf{r}^*(m) \mathbf{r}^T(m) \},$$

$$\mathbf{R}_{rd}^{(k)} = [\mathbf{R}_{dr}^{(k)}]^H = E \{ \mathbf{r}^*(m) \mathbf{d}_k^T(m - m_0) \},$$

$$\mathbf{R}_{dd}^{(k)} = E \{ \mathbf{d}_k^*(m - m_0) \mathbf{d}_k^T(m - m_0) \}.$$

For the sake of simplicity, let us assume that all data sequences $d_k(m)$ are spectrally white and have the same variance σ_d^2 . Then the autocorrelation matrices $\mathbf{R}_{dd}^{(k)}$, $k = 0, \dots, M-1$ are diagonal with diagonal entries σ_d^2 ,

$$\mathbf{R}_{dd}^{(k)} = \sigma_d^2 \mathbf{I}, \quad (17)$$

and Eq. (16) simplifies to

$$Q_k = \mathbf{h}_k^H \mathbf{R}_{rr} \mathbf{h}_k - \mathbf{h}_k^H \mathbf{R}_{rd}^{(k)} \mathbf{p}_k - \mathbf{p}_k^H \mathbf{R}_{dr}^{(k)} \mathbf{h}_k + \sigma_d^2 \mathbf{p}_k^H \mathbf{p}_k. \quad (18)$$

We now consider the minimization of Eq. (18) with respect to \mathbf{p}_k and \mathbf{h}_k under the energy constraints (9). To derive the optimal filters we first derive the optimal vector \mathbf{h}_k given a fixed residual system \mathbf{p}_k . From $\partial Q_k / \partial \mathbf{h}_k = 0$ with Q_k as in Eq. (18), we get

$$\mathbf{h}_k^{(\text{opt})} = \mathbf{R}_{rr}^{-1} \mathbf{R}_{rd}^{(k)} \mathbf{p}_k. \quad (19)$$

Substituting $\mathbf{h}_k^{(\text{opt})}$ into Eq. (18) results in

$$Q_k = -\mathbf{p}_k^H [\mathbf{R}_{rd}^{(k)}]^H \mathbf{R}_{rr}^{-1} \mathbf{R}_{rd}^{(k)} \mathbf{p}_k + \sigma_d^2 \mathbf{p}_k^H \mathbf{p}_k, \quad (20)$$

which now is to be minimized with respect to \mathbf{p}_k under the constraint (9). This yields the eigenvalue problems

$$\left[\sigma_d^2 \mathbf{I} - [\mathbf{R}_{rd}^{(k)}]^H \mathbf{R}_{rr}^{-1} \mathbf{R}_{rd}^{(k)} \right] \mathbf{p}_k = \lambda_k \mathbf{p}_k, \quad (21)$$

$$k = 0, 1, \dots, M-1$$

which are essentially similar to the one in [20] for the single-channel case. The optimal vectors \mathbf{p}_k are the eigenvectors that belong to the respective smallest eigenvalues λ_k , $k = 0, \dots, M-1$.

The receive filters designed according to the method described above minimize the error measures Q_k under the energy constraint and thus maximize the signal-to-noise ratios (SNR's) at the filter outputs. Since the filter output signals, together with the residual systems, are fed into the Viterbi detectors, the algorithm maximizes the SNR's as seen by the Viterbi detectors.

4. Adaptive receiver design

The receiver design method presented in the previous section may be difficult to implement under real-world conditions where the required computational power and accuracy are not available. Also, a real-world channel may be slowly time varying, which causes problems for the receiver design above. To avoid such problems we now derive adaptation rules for the receiver design. For this we follow the strategy for the single-channel case in [20]. During adaptation, we assume that the data sequences $d_k(m)$ are known or have been correctly estimated by the receiver, and we use the received samples as noisy estimates of the required correlation terms. We first look at the design of filter $h_k(n)$ (vector \mathbf{h}_k) according to the rule

$$\mathbf{h}_k^{(\mu+1)} = \mathbf{h}_k^{(\mu)} - \gamma_h e_k(\mu) \tilde{\mathbf{r}}_k^*(\mu), \quad (22)$$

where μ denotes the iteration step. The value $e_k(\mu)$ is the value of the error defined in Eq. (7) in the μ th iteration step, and $\tilde{\mathbf{r}}_k(\mu)$ is the receive vector in the μ th step. Finally, γ_h is a factor that controls the step size and convergence speed. Thus, Eq. (22) is similar to the well-known LMS adaptation rule for equalizer design.

A rule for adapting \mathbf{p}_k can be stated as

$$\mathbf{q}_k^{(\mu+1)} = \mathbf{p}_k^{(\mu)} + \gamma_p e_k(\mu) [\mathbf{d}_k(\mu)]^*, \quad (23)$$

$$\mathbf{p}_k^{(\mu+1)} = \frac{\mathbf{q}_k^{(\mu+1)}}{\|\mathbf{q}_k^{(\mu+1)}\|}. \quad (24)$$

The entire iteration is given by Eqs. (22), (23) and (24), where the normalization step in Eq. (24) is needed to ensure that the energy condition (9) will be satisfied by the final filter $\mathbf{p}_k^{(\infty)}$. Using the same arguments as in [20] one can show that the iteration indeed converges to the MMSE solution where the final vector $\mathbf{p}_k^{(\infty)}$ is the eigenvector of $\left[\sigma_d^2 \mathbf{I} - [\mathbf{R}_{rd}^{(k)}]^H \mathbf{R}_{rr}^{-1} \mathbf{R}_{rd}^{(k)} \right]$ that corresponds to the smallest eigenvalue.

5. Results

To demonstrate the performance of the proposed algorithms, we consider data transmission over telephone lines in an ADSL/VDSL related setting [24]. Figure 2 shows the channel impulse response considered in this example. It is assumed that the channel noise is comprised of near and far end crosstalk as well as white Gaussian noise, resulting in the total power spectral density depicted in Fig. 3. We consider the use of a cosine-modulated filter bank for creating the transmit signal, which is an interesting alternative to blockwise DFT's as in DMT. In [12, 25] it was shown that such filter bank based systems offer greater potential than blockwise DFT's because of their longer impulse responses and better frequency selectivity. However,

they need equalization on the receiver side. In the present example, the transmit signal is synthesized via a 16-band cosine-modulated filter bank with extended lapped transform (ELT) prototype [26]. As in [12, 25] pulse amplitude modulation is used to create a real-valued transmit signal.

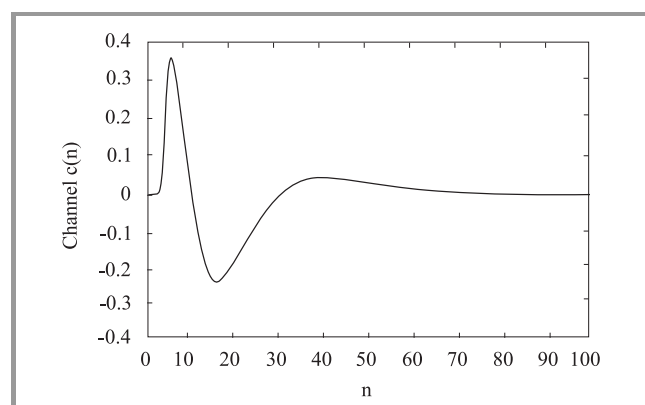


Fig. 2. Transmission channel impulse response.

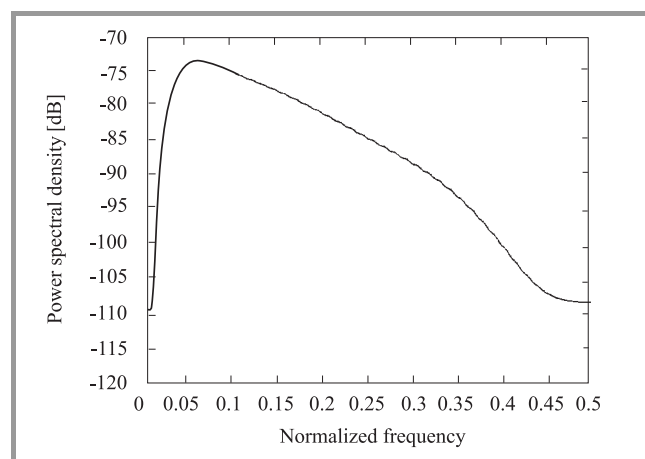


Fig. 3. Noise power spectral density.

Figure 4 shows the signal-to-noise ratios within the different bands at the equalizer output for several configurations. In all cases the lengths of the receive filters are chosen as $L_h = 128$. We first look at the results depicted in Fig. 4a. In this case, all bands are loaded with the same input power σ_d^2 . This means that the transmission system is critically sampled and that no redundancy (e.g. in form of a guard interval) is introduced. The comparison of the three curves in Fig. 4a shows that, especially for the low-frequency channels, memory truncation ($L_p > 1$) significantly outperforms MMSE equalization ($L_p = 1$). A significantly better performance of all methods under consideration is obtained if the first frequency band remains unloaded. Results are depicted in Fig. 4b. This strategy has been proposed in [25] as a possibility to introduce redundancy. Leaving out a particular band has two effects. Firstly, the system becomes oversampled, which means that the transmitter introduces redundancy in form

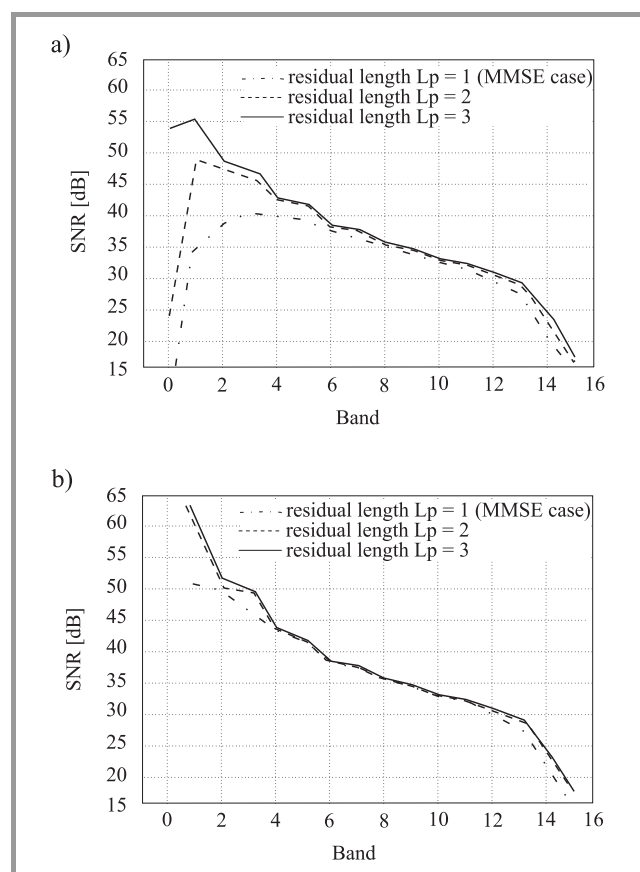


Fig. 4. Signal-to-noise ratios at detector input using a 16-band cosine-modulated filter bank as transmit filters: (a) all bands are loaded; (b) only bands 1–15 are loaded.

of excess bandwidth. Secondly, the receive filters do not need to suppress crosstalk from the dropped channel and have more freedom to equalize their own data paths. As the results in Fig. 4b show, almost all channels gain from the fact that the first band has been left out. When comparing the three curves in Fig. 4b, we see that memory truncation still results in a noticeable improvement over MMSE equalization for a number of bands. The performance difference between $L_h = 2$ and $L_h = 3$, however, is only marginal in Fig. 4b.

From the above examples we see that a receiver based on memory truncation receive filters and low-cost Viterbi detectors can yield a significant improvement over MMSE equalization and threshold detection. In general, the amount of SNR improvement of memory truncation over MMSE equalization depends on the channel in question. Significant improvements can be expected whenever it is difficult to equalize a channel because of extreme frequency selectivity.

To demonstrate the adaptive approach, we consider the receive filter design for the first channel in the above setting. The residual filter length is chosen as $L_p = 3$. At the beginning of the iteration, the receive filter $h_1(n)$ was set to zero, and the residual system $p_1(n)$ was set

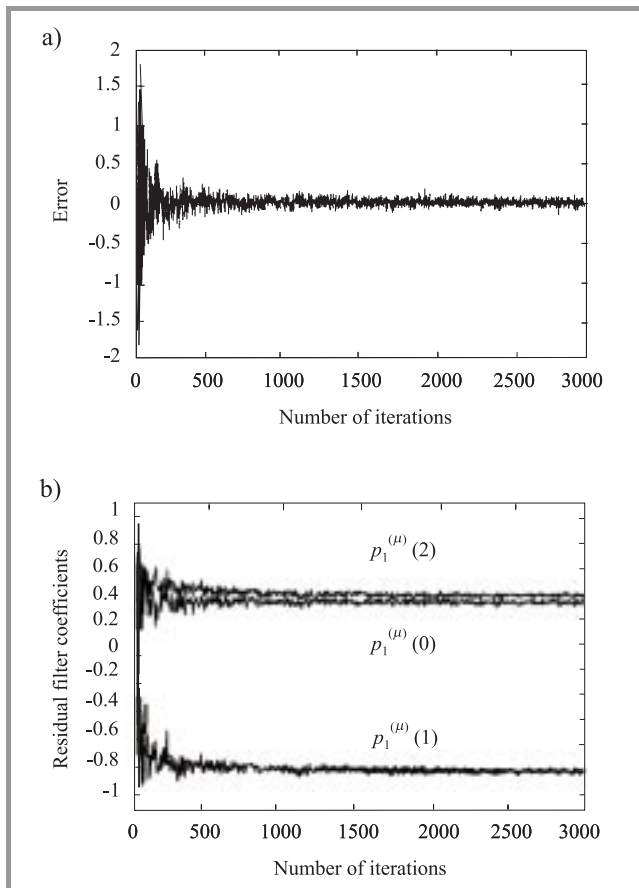


Fig. 5. Receiver adaptation: (a) error signal $e_1(\mu)$; (b) residual filter coefficients $p_1^{(\mu)}(m)$.

to $\{p_1^{(0)}(m)\} = \{1, 0, 0\}$. The step sizes were chosen as $\gamma_h = 0.0015$ and $\gamma_p = 0.33$. Figure 5a depicts the error signal $e_1(\mu)$, and it can be seen that the error rapidly decreases during the first few hundred iterations. Figure 5b shows the adaptation of the three residual coefficients. For comparison, the closed-form solution yields $\{p_1^{(\text{opt})}(m)\} = \{0.382, -0.822, 0.421\}$. These values are approached after some hundred iterations.

6. Conclusions

In this paper, optimal receive filter banks for FDM transmission systems have been presented. The receive filters are designed in such a way that the overall subchannel impulse responses become truncated to predefined lengths. Using an example of high-speed transmission over copper wires it was shown that the SNR can be significantly improved over MMSE equalizer banks. In general, the amount of improvement clearly depends on the channel in question, and there may be cases where MMSE approaches work equally well. The design methods presented are applicable to all transmultiplexing systems where the transmit signal is formed as a linear combination of transmit filter impulse responses with the data symbols being the weights

(e.g. DMT, OFDM, CDMA). Extensions of the proposed methods to a joint transmitter/receiver design are under investigation.

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Alfred Mertins received his Dipl.-Ing. degree from the University of Paderborn, Germany, in 1984, the Dr.-Ing. degree in electrical engineering and the Dr.-Ing. habil. degree in telecommunications from the Hamburg University of Technology, Germany, in 1991 and 1994, respectively. From 1986 to 1991 he was with the Hamburg University of Technology, Germany, from 1991 to 1995 with the Microelectronics Applications Center Hamburg, Germany, from 1996 to 1997 with the University of Kiel, Germany, and from 1997 to 1998 with the University of Western Australia. Since September 1998 he has been with the University of Wollongong as a Senior Lecturer. His research interests include image and video processing, wavelets and filter banks, and digital communications.

e-mail: mertins@uow.edu.au
 University of Wollongong
 School of Electrical, Computer,
 and Telecommunications Engineering
 Northfields Avenue
 Wollongong, NSW 2522, Australia