AN IMPROVED OPTIMIZATION ALGORITHM FOR TIME DOMAIN EQUALIZER DESIGN IN ADSL MODEMS

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Abstract

In this paper, an improved optimization algorithm for the time domain equalizer is developed in case of Frequency Division Multiplexing - based ADSL. In contrast with existing algorithms, our algorithm explicitly disregards the unused tones in the optimization function as well as in the non-triviality constraint. Some simulation results are presented for the upstream channel. It is shown that the achievable bitrate is higher than with a classical algorithm which imposes a unit norm constraint on the full target impulse response.

1 Introduction

Asymmetric Digital Subscriber Line (ADSL) provides a high bitrate downstream channel and a lower bitrate upstream channel over twisted pair copper wire. The transmission method is Discrete Multitone (DMT). It divides the used bandwidth in parallel subchannels or tones. The incoming bitstream is split between all available tones in order to QAM-modulate them. At the emission, a cyclic prefix extension is added to each time symbol, computed by an inverse Fourier Transform. This technique enables the reception to be performed by an FFT, followed by a single complex tap equalizer per tone to compensate for the channel amplitude and phase effects, provided that the channel impulse response is shorter than the cyclic extension. In order to avoid too long cyclic extension, one can advantageously make use of a time domain equalizer (TEQ), which role is to shorten the channel impulse response. Imperfectly shortened channel impulse responses give rise to interblock interference (IBI) between two successive symbols, and intercarrier interference (ICI) between different carriers.

Many algorithms have been developed to initialize this TEQ. Falconer *et al.* [4] showed how a target shortened impulse response can be computed as the eigenvector corresponding to the smallest eigenvalue of a channel dependent matrix. This is used in the context of maximum likelihood sequence estimation. A unit norm constraint is put on this target impulse response to avoid trivial solutions. In [3], Chow *et al.* describe algorithms suited for multicarrier modulation which use adaptation in the frequency domain and windowing in the time domain. Van Bladel *et al.* [7] consider TEQ determination by means of an MMSE criterion, again resulting in an eigenvalue problem. A fast computation algorithm for the TEQ is discussed in [5] by Lee *et al.* where a unit tap constraint on the target impulse response is used. Al-Dhahir *et al.* argue in [1] that the MSE itself is not the optimum equalization criterion in conjunction with DMT. An new criterion is derived, called geometric SNR. Melsa *et al.* [6] derive an optimal shortening algorithm with respect to the shortening SNR. Also in [8], Van Kerckhove *et*

$$\min_{\mathbf{W},\mathbf{B}} \mathcal{E}\left\{\left\| \underbrace{\begin{bmatrix} y_\Delta & y_{\Delta-1} & \cdots & y_{\Delta-(T-1)} \\ y_{\Delta+1} & y_\Delta & & \vdots \\ \vdots & \ddots & & \\ y_{\Delta+N-1} & \cdots & & y_{\Delta+N-T} \end{bmatrix}}_{\mathbf{Y}_T} \cdot \underbrace{\begin{bmatrix} w_0 \\ w_1 \\ \vdots \\ w_{T-1} \end{bmatrix}}_{\mathbf{W}} - \underbrace{\begin{bmatrix} x_1 & x_N & \cdots & x_{N-(\nu-1)} \\ x_2 & x_1 & & \vdots \\ \vdots & \ddots & & \vdots \\ x_N & \cdots & & x_{N-\nu} \end{bmatrix}}_{\mathbf{X}_T} \cdot \underbrace{\begin{bmatrix} b_0 \\ b_1 \\ \vdots \\ b_\nu \end{bmatrix}}_{\mathbf{B}} \right\|_2^2 \right\} \quad (1)$$

al. demonstrate unsatisfactory performance of the MSE algorithm. A unit norm constraint is used for the TEQ to avoid the trivial all-zero solution but then it is shown that an other unwanted TEQ can appear in case Frequency Division Multiplexing is used for up- and downstream: the TEQ-energy moves to the carriers that are not used. This is solved by injecting virtual noise in the stopband.

Here, we focus on the Frequency Division Multiplexing case, and aim at improving upon the results in [8]. Unlike the approach in [8], our algorithm explicitly disregards the unused tones. The non-triviality constraint is imposed only on the used tones such that trivial solutions in the passband are avoided. This is achieved by introducing weight matrices and is shown to improve the resulting capacity of the system.

This paper is organized as follows. The second section explains some aspects of capacity optimization, the classical approach and our proposed improvements. The third section describes the algorithm. Section 4 defines the constraint and gives an interpretation to it. Simulation results are presented in section 5.

2 Channel shortening versus capacity optimization

The optimal TEQ should maximize the transmission capacity. Since this optimization problem is highly non-linear, we focus ourselves to an approach based on the DMT signal properties.

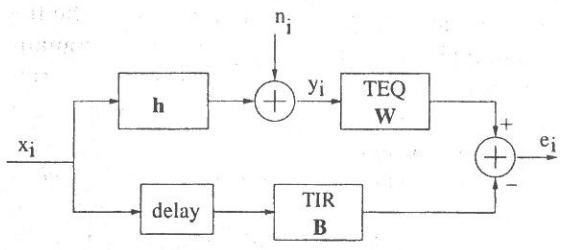


Figure 1: *Model for channel shortening*

An algorithmic approach proven to be efficient in literature [1] - [8] is depicted in figure 1. The mean square error of e_i (in the time domain or E_i in the frequency domain) is minimized over a range of possible delays subject to the constraint that $\|\mathbf{W}\|^2 = 1$ or $\|\mathbf{B}\|^2 = 1$ to avoid trivial solutions, where $\mathbf{W}$ and $\mathbf{B}$ are the coefficient vectors of the TEQ and target impulse response (TIR) respectively. There is however a fundamental problem with this mean square error criterion: there is no direct relation between minimum mean square error and maximum capacity in function of the delay. A smooth capacity function of the delay is desirable to limit the number of tested delays.

Here we try to achieve a higher capacity by inserting weighting matrices in the optimization function in order to minimize the mean square error only over the used tones and modify the constraint to a more meaningful one.

3 Model and algorithm

Our starting point is a formula which explicitly takes into account the nature of the input and output samples. It contains a system of N equations where N is the FFT size, i.e. the length of a symbol without prefix. In our model, ν is the length of the cyclic prefix. The error of time domain shortening of the channel is minimized over the complete symbol length, as shown in equation (1). Here x_i and y_i are respectively input and output data samples of the channel, $\mathbf{W} = w_{0:T-1}$ is the time domain equalizer and $\mathbf{B} = b_{0:\nu}$ the target impulse response. Δ is the delay introduced by the channel. It is noted that the data matrices have a Toeplitz structure. Now we modify this cost function by minimizing only over the used tones. Therefore we transform our system to the frequency domain, leading to

$$\min_{\mathbf{W},\mathbf{B}} \mathcal{E}\left\{\|\mathbf{S} \cdot \mathrm{FFT} \cdot (\mathbf{Y}_T \cdot \mathbf{W} - \mathbf{X}_T \cdot \mathbf{B})\|_2^2\right\} \quad (2)$$

Here, FFT is the $N \times N$ FFT-matrix, $\mathbf{S}_{N\times N}$ is a selector matrix, i.e. a diagonal matrix with ones on the positions corresponding to the used tones and zeros otherwise. By writing the output samples y_i as a function of x_i and the additive white channel

$$\min_{\mathbf{W},\mathbf{B}} \mathcal{E}\left\{\left\| \mathbf{S}\cdot\mathrm{FFT}\cdot\left[\,\mathbf{X}_T^{ext}\;\middle|\;\mathbf{N}_T\,\right]\cdot\left[\begin{array}{c|c} \mathbf{H} & \begin{array}{c} O_{(\Delta-1)\times(\nu+1)} \\ \hline I_{(\nu+1)\times(\nu+1)} \\ \hline O \end{array} \\ \hline I_{T\times T} & O \end{array}\right]\cdot\begin{bmatrix}-\mathbf{W}\\ \mathbf{B}\end{bmatrix}\right\|_2^2\right\} \tag{3}$$

noise n_i, one obtains equation (3), with

$$\mathbf{H}_{(N-1+T)\times T} = \begin{bmatrix} \boxed{\mathbf{h}} & 0 & \cdots & 0 \\ 0 & \boxed{\mathbf{h}} & 0 & \cdots \\ \vdots & \ddots & \ddots & \ddots \\ 0 & \cdots & 0 & \boxed{\mathbf{h}} \end{bmatrix}^{\mathrm{T}} \tag{4}$$

$$\mathbf{N}_T = \begin{bmatrix} n_\Delta & n_{\Delta-1} & \cdots & n_{\Delta-(T-1)} \\ n_{\Delta+1} & n_\Delta & & \vdots \\ \vdots & \ddots & & \\ n_{\Delta+N-1} & \cdots & & n_{\Delta+N-T} \end{bmatrix} \tag{5}$$

$$\mathbf{X}_T^{ext}{}_{N\times(N-1+T)} = \left[\begin{array}{ccc} x_\Delta & \cdots & x_{\Delta+N-1} \\ \vdots & \ddots & \vdots \\ x_2 & \ddots & \\ \hline x_1 & \ddots & x_N \\ x_N & \ddots & \vdots \\ \vdots & \ddots & \\ \hline \vdots & \ddots & \vdots \\ x & & \end{array}\right]^{\mathrm{T}} \tag{6}$$

$\mathbf{X}_T^{ext}$ is a Toeplitz matrix, and an 'extension' of $\mathbf{X}_T$. $\mathbf{N}_T$ is a Toeplitz matrix with noise samples. $\mathbf{h}_{1\times N}$ contains the channel impulse response. To solve this system in 'square root' form, we need the (square roots of) correlation matrices of noise and input, namely

$$\begin{aligned}\tilde{\mathbf{R}}_{nn} &= \mathcal{E}\left\{\tilde{\mathbf{N}}_T^{\mathrm{H}}\cdot\tilde{\mathbf{N}}_T\right\} \\ &= \mathcal{E}\left\{\left(\mathbf{N}_T^{\mathrm{H}}\cdot\mathrm{IFFT}\cdot\mathbf{S}\right)\cdot\left(\mathbf{S}\cdot\mathrm{FFT}\cdot\mathbf{N}_T\right)\right\}\end{aligned} \tag{7}$$

The $\mathbf{S}$ matrices select the appropriate rows and columns in the FFT respectively IFFT matrix. The product IFFT·$\mathbf{S}$·$\mathbf{S}$·FFT gives a real, symmetric circulant weight matrix, see (8). $\tilde{\mathbf{R}}_{nn}$ then becomes:

$$\tilde{\mathbf{R}}_{nn} = \mathcal{E}\left\{\mathbf{N}_T^{\mathrm{H}}\cdot\begin{bmatrix} a & b & c & \cdots & c & b \\ b & a & b & \ddots & & \vdots \\ c & b & \ddots & & & \\ \vdots & \ddots & & & & \\ b & c & \cdots & & & a \end{bmatrix}_{N\times N}\cdot\mathbf{N}_T\right\} \tag{8}$$

$$= \begin{bmatrix} N\cdot a & (N-1)\cdot b & \cdots & (N-T+1)\cdot f \\ (N-1)\cdot b & N\cdot a & \ddots & \vdots \\ \vdots & \ddots & \ddots & \\ (N-T+1)\cdot f & \cdots & & N\cdot a \end{bmatrix}_{T\times T} \cdot\,\mathcal{E}\left\{n_k^2\right\} \tag{9}$$

In expression (9), the elements of the real, symmetric Toeplitz matrix are formed by taking the sum of the diagonals of the weight matrix of equation (8). This can be seen as follows, take one element of $\tilde{\mathbf{R}}_{nn}$:

$$\left\{\tilde{\mathbf{R}}_{nn}\right\}_{ij} = \mathcal{E}\left\{\mathbf{N}_T^{\mathrm{H}}(i,:)\cdot\mathrm{IFFT}\cdot\mathbf{S}\cdot\mathbf{S}\cdot\mathrm{FFT}\cdot\mathbf{N}_T(:,j)\right\} \tag{10}$$

Suppose $i < j$ (the same can be shown for $i \geq j$). Then the $N-(j-i)$ first elements of the ith column of $\mathbf{N}_T$ are the same as the $N-(j-i)$ last elements of the jth column because $\mathbf{N}_T$ is Toeplitz. When developing (10), only terms with a product of white noise samples with the same index have to be considered. These are the only terms left after taking the expected value. Now suppose the $j-i+1$ column of the weight matrix IFFT·$\mathbf{S}$·$\mathbf{S}$·FFT starts with the element g then

$$\begin{aligned}\left\{\tilde{\mathbf{R}}_{nn}\right\}_{ij} = \mathcal{E}\Bigg\{\mathbf{N}_T^{\mathrm{H}}(i,:)\cdot&\begin{bmatrix} a & b & \cdots & g & \cdots & b \\ b & \ddots & \ddots & & g & \vdots \\ \vdots & \ddots & & & & \ddots \\ g & & & & & \\ \vdots & g & & & & \\ b & \cdots & \ddots & & & a \end{bmatrix}_{N\times N} \\ &\cdot\,\mathbf{N}_T(:,j)\Bigg\} \\ = (N-(j-i))\cdot g\cdot&\,\mathcal{E}\left\{n_k^2\right\}\end{aligned} \tag{11}$$

In calculating $\tilde{\mathbf{R}}_{xx}^{ext}$, we make some assumptions such that we can form it similar to $\tilde{\mathbf{R}}_{nn}$. We neglect the prefix structure in $\mathbf{X}_T^{ext}$ and assume the

input samples x_i are white. This is an approximation in that it is based on the assumption that all tones are sent. As already mentioned, the optimization itself is made over the tones that are effectively used. This leads to

$$\begin{aligned}\tilde{\mathbf{R}}_{xx}^{ext} &\approx \mathcal{E}\left\{\tilde{\mathbf{X}}_T^{ext\mathrm{H}} \cdot \tilde{\mathbf{X}}_T^{ext}\right\} \\ &\approx \mathcal{E}\left\{\mathbf{X}_T^{ext\mathrm{H}} \cdot \mathrm{IFFT} \cdot \mathbf{S} \cdot \mathbf{S} \cdot \mathrm{FFT} \cdot \mathbf{X}_T^{ext}\right\} \\ &\approx \mathbf{M} \cdot \mathcal{E}\left\{x_k^2\right\}\end{aligned} \tag{12}$$

$\mathbf{M}$ is a Toeplitz matrix. It is formed in the same way as the matrix in equation (9). However its dimension is larger, namely $(N-1+T) \times (N-1+T)$. Therefore the $T-1$ outer off-diagonals consist of zeros:

$$\mathbf{M} = \begin{bmatrix} N{\cdot}a & (N{-}1){\cdot}b & \dots & b & 0 & \dots & 0 \\ (N{-}1){\cdot}b & \ddots & \ddots & & b & 0 & \vdots \\ \vdots & \ddots & & & & \ddots & \ddots \\ b & & & & & & \\ 0 & b & & & & & \\ \vdots & 0 & \ddots & & & \ddots & \\ 0 & \dots & \ddots & & & & N{\cdot}a \end{bmatrix} \tag{13}$$

4 Non-triviality constraint

As a non-triviality constraint, we take

$$\mathcal{E}\left\{\left\|\tilde{\mathbf{X}}_T \cdot \mathbf{B}\right\|_2^2\right\} = 1 \tag{14}$$

This can be written as

$$\begin{aligned}&\mathcal{E}\left\{\mathbf{B}^{\mathrm{H}} \cdot \mathbf{X}_T^{\mathrm{H}} \cdot \mathrm{IFFT} \cdot \mathbf{S}^2 \cdot \mathrm{FFT} \cdot \mathbf{X}_T \cdot \mathbf{B}\right\} = 1 \\ &\mathcal{E}\Big\{\left[\mathbf{B}^{\mathrm{H}} \quad O_{1\times(N-\nu)}\right] \cdot \mathbf{X}_T^{circ\mathrm{H}} \cdot \mathrm{IFFT} \cdot \mathbf{S}^2 \cdot \\ &\qquad \mathrm{FFT} \cdot \mathbf{X}_T^{circ} \cdot \begin{bmatrix}\mathbf{B} \\ O_{(N-\nu)\times 1}\end{bmatrix}\Big\} = 1\end{aligned} \tag{15}$$

with ${\mathbf{X}_T^{circ}}_{N\times N}$ the circular extension of $\mathbf{X}_T$ with prefix taken in account

$$\mathbf{X}_T^{circ} = \begin{bmatrix} x_1 & x_N & \dots & x_2 \\ x_2 & x_1 & \ddots & \vdots \\ \vdots & \ddots & & \vdots \\ x_N & \dots & & x_1 \end{bmatrix} \tag{16}$$

After insertion of an IFFT·FFT pair between the circulant matrix and the target impulse response vector, one obtains

$$\begin{aligned}&\mathcal{E}\Big\{\left[B_0^* \quad \dots \quad B_{N-1}^*\right] \cdot \begin{bmatrix} X_0^* & & \\ & \ddots & \\ & & X_{N-1}^* \end{bmatrix} \cdot \mathbf{S}^2 \cdot \\ &\qquad \begin{bmatrix} X_0 & & \\ & \ddots & \\ & & X_{N-1} \end{bmatrix} \cdot \begin{bmatrix} B_0 \\ \vdots \\ B_{N-1} \end{bmatrix}\Big\} = 1\end{aligned} \tag{17}$$

This means

$$\boxed{\textstyle\sum_{i\in\{used\ tones\}} |B_i|^2 = 1} \tag{18}$$

under the assumption that $\mathcal{E}\left\{|X_i|^2\right\} = 1$. So, as far as the constraint is concerned, the prefix is taken in account. Practically, the constraint is calculated as follows

$$\begin{aligned}&\left[B_0^* \quad \dots \quad B_{N-1}^*\right] \cdot \mathbf{S}^2 \cdot \begin{bmatrix} B_0 \\ \vdots \\ B_{N-1} \end{bmatrix} = 1 \\ &\left[\mathbf{B}^{\mathrm{H}} \quad O_{1\times(N-\nu)}\right] \cdot \mathrm{IFFT} \cdot \mathbf{S}^2 \cdot \mathrm{FFT} \cdot \begin{bmatrix}\mathbf{B} \\ O_{(N-\nu)\times 1}\end{bmatrix} = 1 \\ &\mathbf{B}^{\mathrm{H}} \cdot \mathbf{R}^{\mathrm{H}} \cdot \mathbf{Q}^{\mathrm{H}} \cdot \mathbf{Q} \cdot \mathbf{R} \cdot \mathbf{B} = 1\end{aligned} \tag{19}$$

Now call $\tilde{\mathbf{B}} = \mathbf{R} \cdot \mathbf{B}$. The resulting optimization function then becomes equation (20), subject to $\tilde{\mathbf{B}}^{\mathrm{H}} \cdot \tilde{\mathbf{B}} = 1$. This constraint is meaningful because of the following reason: the solution of the constrained MSE problem equals the solution of the unconstrained MSE problem

$$\min_{\mathbf{W},\mathbf{B}} \frac{\left[-\mathbf{B}^{\mathrm{H}}\ \mathbf{W}^{\mathrm{H}}\right] \cdot \begin{bmatrix}\tilde{\mathbf{R}}_{xx} & \tilde{\mathbf{R}}_{xy} \\ \tilde{\mathbf{R}}_{yx} & \tilde{\mathbf{R}}_{yy}\end{bmatrix} \cdot \begin{bmatrix}-\mathbf{B} \\ \mathbf{W}\end{bmatrix}}{\mathbf{B}^{\mathrm{H}} \cdot \tilde{\mathbf{R}}_{xx} \cdot \mathbf{B}} \tag{21}$$

where the $\tilde{\mathbf{R}}$-matrices are correlations of $\tilde{\mathbf{Y}}_T = \mathbf{S} \cdot \mathrm{FFT} \cdot \mathbf{Y}_T$ and $\tilde{\mathbf{X}}_T = \mathbf{S} \cdot \mathrm{FFT} \cdot \mathbf{X}_T$. This formula can be seen as 1/SNR for the subband of the used tones.

5 Simulation results

In this section we present some simulation results for the proposed optimization algorithm. It is shown that there is an improvement for the upstream case where an 128-point FFT is used. Of the 65 tones, tone 8 up to tone 32 are used. In the downstream case, based on a 512-point FFT, only tones 38 up to 256 of the 257 carriers are used. So relatively speaking more tones are used in the downstream case. Probably for this reason, the weight factors of our algorithm have not much influence on the resulting capacity of the downstream system.

$$\min_{\mathbf{W},\tilde{\mathbf{B}}} \left\| \begin{bmatrix} \text{chol}(\tilde{\mathbf{R}}_{xx}^{ext}) & O \\ O & \text{chol}(\tilde{\mathbf{R}}_{nn}) \end{bmatrix} \cdot \left[\begin{array}{c|c} \mathbf{H} & \begin{array}{c} O_{(\Delta-1)\times(\nu+1)} \\ \hline I_{(\nu+1)\times(\nu+1)} \\ \hline O \end{array} \\ \hline I_{T\times T} & O \end{array}\right] \cdot \begin{bmatrix} I & O \\ O & \mathbf{R}^{-1} \end{bmatrix} \cdot \begin{bmatrix} -\mathbf{W} \\ \tilde{\mathbf{B}} \end{bmatrix} \right\|_2^2 \qquad (20)$$

In the simulations, we assume additive white channel noise in the TEQ as well as in the capacity calculations. For upstream we have: $N = 128$, $\nu = 8$, sample frequency $F_s = 0.552$ MHz and to calculate the capacity: SNR gap $\Gamma = 9.8$ dB, noise margin $\gamma_m = 6$ dB, coding gain $\gamma_c = 3$ dB. The amount of bits b assigned to tone i is calculated as follows:

$$b_i = \text{floor}\left(\log_2\left(1 + 10^{\frac{SNR_i - \Gamma - \gamma_m + \gamma_c}{10}}\right)\right) \qquad (22)$$

The 'floor' operator is inserted to round b_i to the nearest smaller integer. The rate is calculated with the formula

$$rate = \left(\sum_{i=used\ tone} b_i\right) \cdot \frac{F_s}{N+\nu} \qquad (23)$$

The transfer characteristics of the channels are plotted in figures 2 en 3. Figures 4 to 9 show the simulation results. The full and dashed lines in

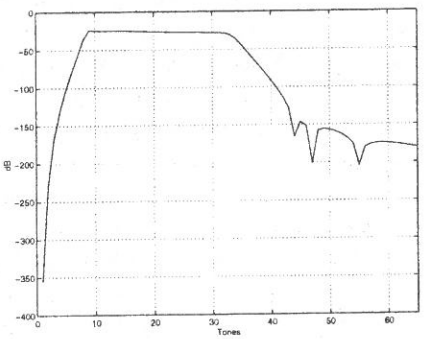


Figure 2: *Transfer characteristic of an 1500m upstream channel*

these figures are respectively the results of the algorithm with weight matrix corresponding to the used tones and of the classical algorithm with $\|\mathbf{B}\|^2 = 1$. The target impulse responses are in each case calculated by the singular value decomposition after qr-factorization of the appropriate system of equations.

Figure 4 shows the capacity in a delay range [-15:50] of an upstream channel of 1500 m. The TEQ length T is 64, the power spectral densities are psd_x = -38 dBm/Hz and psd_n = -118 dBm/Hz. The zero reference delay is taken as the maximum of the absolute value of the channel impulse response. One sees that the rate is less dependent of the delay.

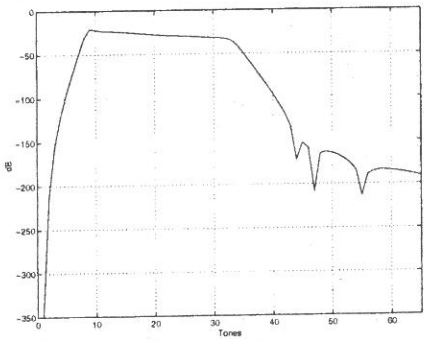


Figure 3: *Transfer characteristic of an 4000m upstream channel*

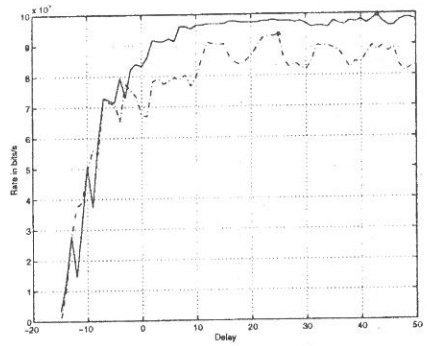


Figure 4: *Capacity in function of the delay for an 1500m upstream channel with additive white channel noise and TEQ length T=64*

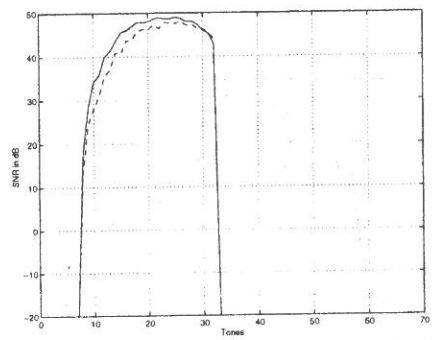


Figure 5: *SNR for the best TEQ for the two methods shown in figure 4*

Figure 5 shows the SNR for the best TEQ's of the two methods. In figures 6 and 7 the same is plotted for T=16. The difference between the reachable

capacities is higher as is also clear in the SNR plot. Figures 8 and 9 are made for an upstream channel of 4000 m with T=32. Again, the capacity obtained with weighting matrix based approach is higher.

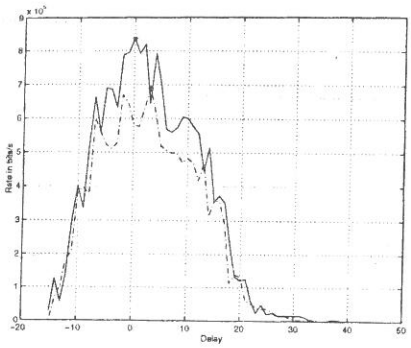


Figure 6: *Capacity in function of the delay for an 1500m upstream channel with additive white channel noise and TEQ length T=16*

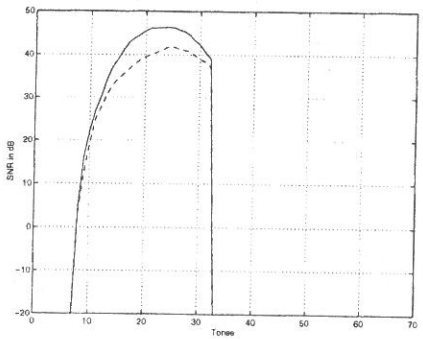


Figure 7: *SNR for the best TEQ for the two methods shown in figure 6*

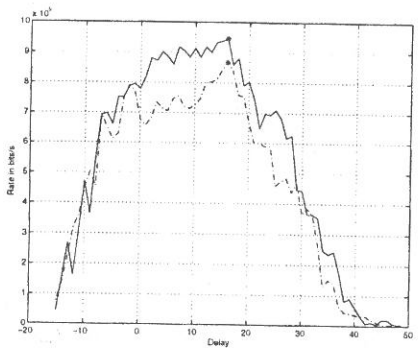


Figure 8: *Capacity in function of the delay for an 4000m upstream channel with additive white channel noise and TEQ length T=32*

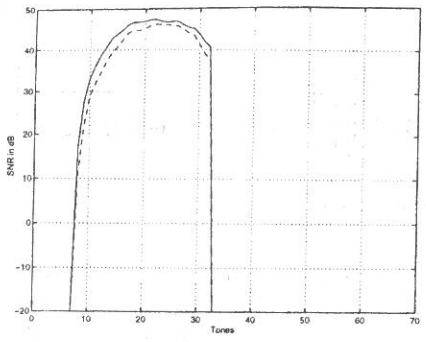


Figure 9: *SNR for the best TEQ for the two methods shown in figure 8*

6 Conclusions

In this paper we have developed an algorithm for the TEQ in FDM-ADSL modems. An improved performance with respect to $\|\mathbf{B}\|^2 = 1$ is shown for the upstream channel. The Δ dependence is lowered which results in less complexity and is also inversely proportional to the number of TEQ-taps T.

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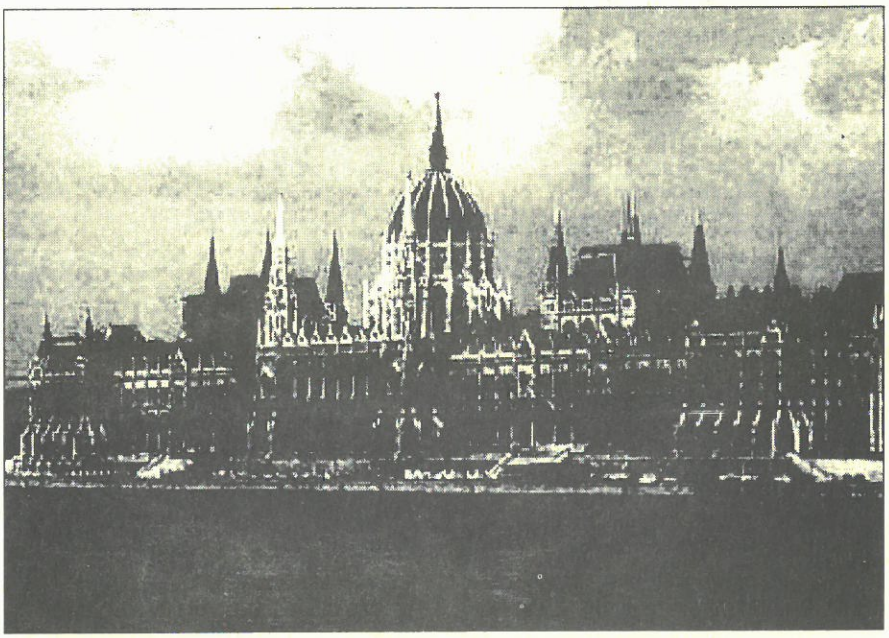
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