COMBINED EQUALIZATION AND ALIEN CROSSTALK CANCELLATION IN ADSL RECEIVERS

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ABSTRACT

We address the problem of mitigating a dominant alien near-end crosstalker such as HDSL, SDSL or HPNA in DMT-based ADSL and VDSL receivers. Due to the different symbol rates of crosstalker and signal of interest, linear time-invariant filtering is not appropriate. Recently, Zeng et al. presented a method to tackle this problem [4]. In this paper, we present an alternative procedure, which, unlike the solution of [4], does not require any prior knowledge (such as transmission and crosstalk channel, noise characteristics, etc.) other than the crosstalker symbol rate. Moreover, it is fully adaptive, integrates equalization and crosstalk cancellation and optimizes the signal-to-noise ratio for each tone separately. Both algorithms exploit spectral correlation in the received crosstalking signal, thanks to oversampling and excess bandwidth. The method present here is based on the application of so called FREquency-SHift (FRESH) filtering.

1 INTRODUCTION

Copper-wire telephone lines have become a high-speed data transmission medium through the application of various digital subscriber line (DSL) technologies. These xDSL technologies vary in data rate, reach, sampling rate, bandwidth and modulation scheme. Examples include HDSL (high-bitrate DSL), SDSL (symmetric single-pair high-bitrate DSL), ADSL (asymmetric DSL) and VDSL (very high-bitrate DSL) ([1],[2]).

One of their major impairments is the severe crosstalk (XT) between copper pairs in the same or neighboring bundles. Crosstalk is classified as near-end crosstalk (NEXT), if it originates from a transmitter at the same bundle end. In contrast, far-end crosstalk (FEXT) is caused by a transmitter at the opposite bundle end. Another classification distinguishes between self-crosstalk and alien crosstalk. Crosstalking by systems using the same xDSL technology is referred to as self-crosstalk: e.g. in ADSL, self-NEXT can be avoided by frequency division duplexing upstream and downstream. Alien crosstalk considers xDSL technologies with different transmission schemes that overlap in frequency: e.g. HDSL and SDSL are both single carrier schemes that cause alien NEXT in a discrete multitone (DMT)-based system, such as ADSL. Non-xDSL applications, such as Home LAN (HPNA) using indoor copper wiring [3], can cause alien NEXT in VDSL systems in the absence of appropriate rejection filters.

The different symbol rates of crosstalker and signal of interest make linear time-invariant crosstalk suppression inappropriate: the crosstalking signal is not stationary with respect to the impaired receiver. In [4], a non-adaptive frequency-domain method is presented to cancel one dominant alien NEXT signal (e.g. SDSL, HDSL) in an ADSL downstream receiver. The method exploits the fact that the crosstalk signal, after receiver sampling, is cyclostationary - the interferer symbol rate is lower than the ADSL sampling rate - and has a large excess bandwidth (up to 100 %), hence considerable spectral correlation. It is assumed that both the ADSL transmission channel and crosstalk coupling channel, are identified by a third party and hence that a channel model is available [5]. Moreover, the noise characteristics (of sources other than the dominant crosstalker) and the sampling clock offset between interferer and ADSL should be known and a channel shortening time-domain equalizer is used (although not explicitly mentioned).

The method, discussed in this paper, addresses the same problem. Here also, the crosstalk suppression takes place in the frequency-domain by exploiting the spectral redundancy of the crosstalker. However, it differs from the method proposed in [4] by the fact that it does not require any prior knowledge other than the crosstalker symbol rate. It is based on the application of so called FREquency-SHift (FRESH) filtering to exploit spectral redundancy in interfering signals [6]. This leads to an adaptive structure, integrating DMT equalization and crosstalk suppression and optimizing SNR for each tone separately: it is trained during start-up and allows to track changes in channels and/or interference scenario afterwards.

2 TOWARDS A COMBINED PER-TONE CROSSTALK CANCELLER AND EQUALIZER

We consider an alien crosstalker with a symbol rate lower than the DMT receiver sampling rate. As a consequence, the received crosstalk signal is oversampled, be it at an non-integer multiple of the symbol rate, and cyclostationary. Moreover, signals such as HDSL, (certain versions of) SDSL and HPNA have considerable excess bandwidth (EBW), up to 100 %. We aim at exploiting the resulting spectral correlation between the interferer main lobe and excess band.

The algorithm is applied to each received DMT block symbol k. As in [4], the contaminated tones are divided into two sets. A note on the choice of these sets is added in Section 3. In the set $\mathbf{S}_1$, e.g. corresponding to the interferer excess band, the crosstalk is

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Step 0	Selection of sets of tones $\mathbf{S}_1$ and $\mathbf{S}_2$
Step 1	Per-tone equalization on $\mathbf{S}_1$; no crosstalk cancellation
Step 2	ADSL cancellation on tones $\mathbf{S}_1$
Step 3	Combined FRESH per-tone equalization and crosstalk cancellation on $\mathbf{S}_2$ by exploiting spectral correlation with crosstalk in $\mathbf{S}_1$

Table 1: FRESH per-tone crosstalk cancellation.

treated as additive Gaussian noise, ADSL is detected and cancelled. For the detection, we use the adaptive per-tone equalizer for DMT-based systems (Section 2.1) [7]. As the crosstalk is treated as Gaussian noise, there is no performance gain in set $\mathbf{S}_1$. The crosstalk information on the tones of $\mathbf{S}_1$, after ADSL cancellation (Section 2.2), is then used to remove crosstalk in the second set of tones, $\mathbf{S}_2$, e.g. the interferer main lobe. The crosstalk cancellation method is based on the application of so called FREquency-SHift (FRESH) filtering to exploit the spectral redundancy of the interfering signal (Section 2.3) [6]. Section 2.4 explains how to transform the time-domain FRESH filtering into an adaptively implemented "per-tone" frequency-domain version. The algorithm is summarized in Table 1.

2.1 Per-tone equalization revisited

In [7], an alternative DMT equalizer structure is derived, starting from the traditional time-domain equalizer (TEQ) based structure. Consider tone n, then the usual TEQ operation (channel shortening), together with 1-tap frequency-domain equalization (FEQ), is written as

$$\hat{U}_k[n] = D_n \mathcal{F}[n,:](\mathbf{Y}_k \mathbf{w}) \tag{1}$$

with $\hat{U}_k[n]$ the symbol estimate, D_n the 1-tap FEQ, $\mathcal{F}[n,:]$ the n-th row of the $N \times N$ DFT matrix, $\mathbf{w}$ the T-tap TEQ in a column vector and $\mathbf{Y}_k$ a $N \times T$ Toeplitz matrix constructed from the received signal vector for block symbol k

$$\mathbf{y}_k = \begin{bmatrix} y_k[-T+1] & \cdots & y_k[N-1] \end{bmatrix}^T \tag{2}$$

with $\begin{bmatrix} y_k[0] & \cdots & y_k[-T+1] \end{bmatrix}$ and $\begin{bmatrix} y_k[0] & \cdots & y_k[N-1] \end{bmatrix}^T$ the first row and first column of $\mathbf{Y}_k$. Rewriting (1) leads to:

$$\hat{U}_k[n] = \underbrace{D_n \bar{\mathbf{w}}^T \mathbf{T}_n}_{\mathbf{v}_n} \underbrace{\begin{bmatrix} \Delta \mathbf{y}_k \\ Y_k[n] \end{bmatrix}}_{\mathbf{z}_k[n]} \begin{matrix} \updownarrow T-1 \\ \updownarrow 1 \end{matrix} \quad . \tag{3}$$

Here $\mathbf{T}_n$ is a tone-dependent upper triangular matrix ($\alpha_n = e^{-j2\pi n/N}$)

$$\mathbf{T}_n = \begin{bmatrix} 1 & \alpha_n & \cdots & \alpha_n^{T-1} \\ 0 & 1 & \ddots & \alpha_n^{T-2} \\ \vdots & & \ddots & \vdots \\ 0 & 0 & \cdots & 1 \end{bmatrix}, \tag{4}$$

$\bar{\mathbf{w}}$ denotes the vector $\mathbf{w}$ in reverse order; the $(T-1)$ difference terms are defined as

$$\Delta \mathbf{y}_k = \begin{bmatrix} y_k[-T+1] - y_k[N-T+1] \\ \vdots \\ y_k[-1] - y_k[N-1] \end{bmatrix} \tag{5}$$

and finally $Y_k[n]$ is the FFT output for tone n:

$$Y_k[n] = \mathcal{F}[n,:] \begin{bmatrix} y_k[0] & \cdots & y_k[N-1] \end{bmatrix}^T . \tag{6}$$

Formula (3) indicates that the TEQ-filter is replaced by a tone-dependent $\mathbf{v}_n$-filter acting upon the FFT output $Y_k[n]$ and a set of $T-1$ difference terms $\mathbf{\Delta y}_k$. The vector $\mathbf{v}_n$ in (3) can then be optimized by solving a least squares problem for each tone separately, hence the term "per-tone equalization". The scheme always results in better performance compared to a TEQ-based scheme for constant T, while keeping complexity during data transmission at the same level.

Moreover, an efficient recursive least squares (RLS) algorithm with inverse updating is available for initializing $\mathbf{v}_n$ [8]. At first sight, a lower triangular $T \times T$ matrix $\mathbf{L}_k[n]$, corresponding to the inverse transpose of the Cholesky factor of the sample autocovariance matrix of $\mathbf{z}_k[n]$ in (3), must be stored and updated for each tone. However, the top $T-1$ rows of $\mathbf{L}_k[n]$ are common for all tones, only the last row is tone-dependent. This yields a drastic reduction in required memory and computational effort.

This per-tone equalization is applied, as a **first step**, to $\mathbf{S}_1$ where crosstalk is treated as additive Gaussian noise.

2.2 ADSL cancellation

In a **second step**, the received ADSL signal on the tones $\mathbf{S}_1$ is subtracted from the received time-domain signal vector $\mathbf{y}_k$. The crosstalk information in $\mathbf{S}_1$ will then be used to cancel crosstalk in $\mathbf{S}_2$ in the third step.

The ADSL signal on the n-th FFT output (6) in $\mathbf{S}_1$ is a superposition of the desired symbol part and intersymbol/intercarrier interference (ISI/ICI) from neighboring block symbols and tones. The desired symbol part can be written as $H_k[n]U_k[n]$ with $U_k[n]$ the transmitted frequency-domain symbol (detected in step 1 or available during training) and $H_k[n]$ the channel frequency response, both on tone n. As an approximation, we only cancel the (time-domain contribution of the) desired symbol part $H_k[n]U_k[n]$ in $\mathbf{y}_k$. The amount of residual ISI/ICI will influence the quality (crosstalk-to-noise ratio) of the crosstalk information in $\mathbf{S}_1$.

The cancelling operation for all tones n in $\mathbf{S}_1$ can then be written as (assuming correct decisions $U_k[n]$):

$$\tilde{\mathbf{y}}_k = \mathbf{y}_k - \sum_{n \in \mathbf{S}_1} H_k[n] U_k[n] \mathcal{I}_{ext}[:,n] \tag{7}$$

where $\mathcal{I}_{ext}[:,n] = \begin{bmatrix} \mathcal{I}[N-T+1:N-1,n]^T & \mathcal{I}[:,n]^T \end{bmatrix}^T$ is an extended version of the n-th column of the $N \times N$ IDFT matrix $\mathcal{I}$. Provided that $T-1$ is smaller than or equal to the cyclic prefix length, the extension corresponds to the last $T-1$ samples of the cyclic prefix.

The ADSL cancellation step (7) requires an estimate of the channel frequency response $H_k[n]$ for each tone n of $\mathbf{S}_1$. An unbiased estimate for $H_k[n]$ can be obtained by recursively solving the least squares problem (8). From the derivation in [9] where the per-tone equalizer has been derived assuming the transmission channel is represented with an IIR model, one can show that the following approximation holds:

$$\frac{\hat{U}_j[n]}{\mathbf{v}_n[T]} \approx H_j[n]U_j[n] + N_j[n],\ j = 1, \cdots, k \quad (8)$$

where $\hat{U}_j[n]$, as in (1) and (3), is the equalizer output before slicing, $\mathbf{v}_n[T]$ is the last (i.e. T-th) equalizer coefficient in (3) on tone n and $N_j[n]$ is a noise term.

2.3 Frequency-shift (FRESH) filtering [6]

The **third step**, the crosstalk cancellation in $\mathbf{S}_2$, is based on FRESH filtering to exploit spectral correlation in the interfering signal.

Given a received signal $y[k]$ that is the sum of a desired signal $d[k]$ with symbol rate f_d and an interfering signal $i[k]$ with symbol rate f_i. Assume both $d[k]$ and $i[k]$ are baseband, oversampled, hence cyclostationary, and have excess bandwidth. The FRESH filter to extract $d[k]$ from $y[k]$ is then given by

$$\hat{d}[k] = \sum_f w^f[k] \otimes (e^{j2\pi fk} y[k]) \quad (9)$$

where $\otimes$ denotes the convolution operation. The estimate $\hat{d}[k]$ is a filtered sum of frequency shifted versions of the received signal $y[k]$. The frequency shifts f are linear combinations of f_d and f_i: $f = af_d + bf_i$ with a, b integers. The optimal number n_f and choice of frequency shifts depend on the excess bandwidth of $d[k]$ and $i[k]$ and is not always easy to determine. Although the overall operation to extract $\hat{d}[k]$ from $y[k]$ is time-varying, the filters $w^f[k]$ are linear time-invariant. The optimal $w^f[k]$ can thus be seen as the solution to a multiple-input, single-output filtering problem with the n_f frequency shifted versions of $y[k]$ as inputs.

Oversampling the DMT signal $d[k]$ is beyond the scope of this paper. Only the interferer $i[k]$ is cyclostationary. With up to 100 % EBW (e.g. HDSL, SDSL, HPNA), the optimal time-domain FRESH filtering then becomes:

$$\begin{aligned}\hat{d}[k] &= w^0[k] \otimes y[k] + \\ &\quad w^{-f_i}[k] \otimes (e^{-j2\pi f_i k} y[k]) + w^{f_i}[k] \otimes (e^{j2\pi f_i k} y[k])\end{aligned} \quad (10)$$

The first term has no frequency shift ($f = 0$): $w^0[k]$ corresponds to the TEQ. The second and third term are complex conjugate to make sure that $\hat{d}[k]$ is real. By shifting $y[k]$ over $f = \pm f_i$ Hz, the interference part in $y[k]$ and $e^{\pm j2\pi f_i k} y[k]$ are correlated, while the desired signal part is not.

2.4 FRESH per-tone crosstalk cancellation

The time-invariant filters $w^f[k]$ in (10) can be transferred to the frequency-domain in a way that is very similar to the transfer of the TEQ from time-domain to frequency-domain in Section 2.1. This allows us to use FRESH filtering on the tones of $\mathbf{S}_2$. Tone-specific FRESH filters allow to optimize the SNR for each tone separately.

Extending the TEQ operation in (1) with frequency shifts along the lines of (10) leads to:

$$\begin{aligned}\hat{U}_k[n] &= D_n \mathcal{F}[n,:](\tilde{\mathbf{Y}}_k \mathbf{w}^0 + \\ &\quad \tilde{\mathbf{Y}}_k^{f_i} \mathbf{w}^{f_i} + \tilde{\mathbf{Y}}_k^{-f_i} \mathbf{w}^{-f_i})\end{aligned} \quad (11)$$

where $\tilde{\mathbf{Y}}_k$ is an $N \times T$ Toeplitz matrix constructed (in the same way as $\mathbf{Y}_k$ in (1)) from $\tilde{\mathbf{y}}_k$ in (7), i.e. the received block symbol k with the ADSL symbols of $\mathbf{S}_1$ cancelled; $\tilde{\mathbf{Y}}_k^{f_i}$ and $\tilde{\mathbf{Y}}_k^{-f_i 1}$ are obtained by multplying each element $\tilde{y}_k[l]$ of $\tilde{\mathbf{Y}}_k$ with $e^{\pm j2\pi f_i(ks+l)}$ (k is the block symbol index, $s = N + \nu$ is the block symbol length, ν is the cyclic prefix length); $\mathbf{w}^0$ corresponds to the TEQ; $\mathbf{w}^{f_i}$ and $\mathbf{w}^{-f_i}$ are column vectors with the coefficients of the filters $w^{-f_i}[k]$ and $w^{f_i}[k]$ in (10).

In the case of an interferer with at most 100% EBW (i.e. bandlimited between $-f_i$ and f_i), the interfering signal, frequency shifted over $-f_i$ Hz, has no spectral content in the band of interest (i.e. the frequency bins considered after FFT modulation) between 0 and $f_d/2$ Hz (with f_d the DMT sampling rate). As a consequence, the last term in (11) can be omitted. Equation (11) is then equivalent to:

$$\begin{aligned}\hat{U}_k[n] &= \underbrace{D_n \bar{\mathbf{w}}^{0^T} \mathbf{T}_n}_{\mathbf{v}_n^0} \underbrace{\begin{bmatrix} \Delta\tilde{\mathbf{y}}_k \\ \tilde{Y}_k[n] \end{bmatrix}}_{\tilde{\mathbf{z}}_k[n]} + \underbrace{D_n \bar{\mathbf{w}}^{f_i^T} \mathbf{T}_n}_{\mathbf{v}_n^{f_i}} \underbrace{\begin{bmatrix} \Delta\tilde{\mathbf{y}}_k^{f_i} \\ \tilde{Y}_k^{f_i}[n] \end{bmatrix}}_{\tilde{\mathbf{z}}_k^{f_i}[n]} \\ &= \underbrace{\begin{bmatrix} \mathbf{v}_n^0 & \mathbf{v}_n^{f_i} \end{bmatrix}}_{\mathbf{v}_n^{0,f_i}} \underbrace{\begin{bmatrix} \tilde{\mathbf{z}}_k[n] \\ \tilde{\mathbf{z}}_k^{f_i}[n] \end{bmatrix}}_{\tilde{\mathbf{z}}_k^{0,f_i}[n]}\end{aligned} \quad (12)$$

The tone-dependent transformation matrix $\mathbf{T}_n$ is defined in (4). The difference term vectors $\Delta\tilde{\mathbf{y}}_k$ and $\Delta\tilde{\mathbf{y}}_k^{f_i}$ are similarly defined as $\Delta\mathbf{y}_k$ in (5). As the ADSL cancellation step in (7) makes use of periodically extended columns of the IDFT matrix $\mathcal{I}$, it is seen that $\Delta\mathbf{y}_k = \Delta\tilde{\mathbf{y}}_k$. Similarly, $\tilde{Y}_k[n] = Y_k[n]$ for tones n of $\mathbf{S}_2$, meaning that no extra FFT is needed to calculate $\tilde{Y}_k[n]$. A (partial) extra FFT is applied to the frequency shifted sequence $\begin{bmatrix} \tilde{y}_k^{f_i}[0] & \cdots & \tilde{y}_k^{f_i}[N-1] \end{bmatrix}$ to obtain $\tilde{Y}_k^{f_i}[n]$ on $\mathbf{S}_2$.

As the optimal time-domain filters $\mathbf{w}_f$ are time-invariant, so are the tone-dependent vectors $\mathbf{v}_n^{0,f_i}$. They can be optimized by solving a least squares problem for each tone separately. The coefficients $\mathbf{v}_n^0$ correspond to the per-tone equalizer; adding the FRESH part $\mathbf{v}_n^{f_i}$ results in crosstalk cancellation, hence the term "combined FRESH per-tone equalization and crosstalk cancellation". The recursive least squares initialization algorithm with inverse updating of [8] can be extended. To come to an efficient updating scheme, the data vector $\tilde{\mathbf{z}}_k^{0,f_i}[n]$ in (12) should be rearranged as follows:

$$\tilde{\tilde{\mathbf{z}}}_k^{0,f_i}[n] = \begin{bmatrix} \Delta\mathbf{y}_k^T & \Delta\tilde{\mathbf{y}}_k^{f_i^T} & Y_k[n] & \tilde{Y}_k^{f_i}[n] \end{bmatrix}^T. \quad (13)$$

Again, a tone-dependent lower triangular matrix $\tilde{\mathbf{L}}_k[n]$ is needed in the RLS updating scheme. It satisfies

$$\tilde{\mathbf{L}}_k[n]^H \tilde{\mathbf{L}}_k[n] = \left(\sum_{j=1}^{k} \tilde{\tilde{\mathbf{z}}}_j^{0,f_i}[n]^* \tilde{\tilde{\mathbf{z}}}_j^{0,f_i}[n]^T \right)^{-1} \quad (14)$$

and has the following properties: the first $T-1$ rows are equal to the top rows of $\mathbf{L}_k[n]$ of set $\mathbf{S}_1$. The following $T-1$ rows of $\tilde{\mathbf{L}}_k[n]$ are common for all tones of $\mathbf{S}_2$. The last two rows are tone-dependent. These properties can straightforwardly be exploited to reduce the initialization complexity.

3 SIMULATION RESULTS

Time-domain simulations for downstream ADSL loops (tones 32 to 255) of several lengths are done. The contaminating interferer is SDSL with a symbol rate of 1040 kHz and 100% EBW. An experimentally obtained crosstalk coupling function is used. Self-NEXT is assumed negligible; 24 self-FEXT sources are included. This simulation set-up corresponds to [4].

The proper selection of the sets of tones $\mathbf{S}_1$ and $\mathbf{S}_2$ in an optimal way is beyond the scope of this paper. In [4], it is argued that in the

[1] For simplicity, we use the same number of taps T for the filters $w^0[k]$, $w^{f_i}[k]$ and $w^{-f_i}[k]$. The extension for different numbers of taps is straight forward.

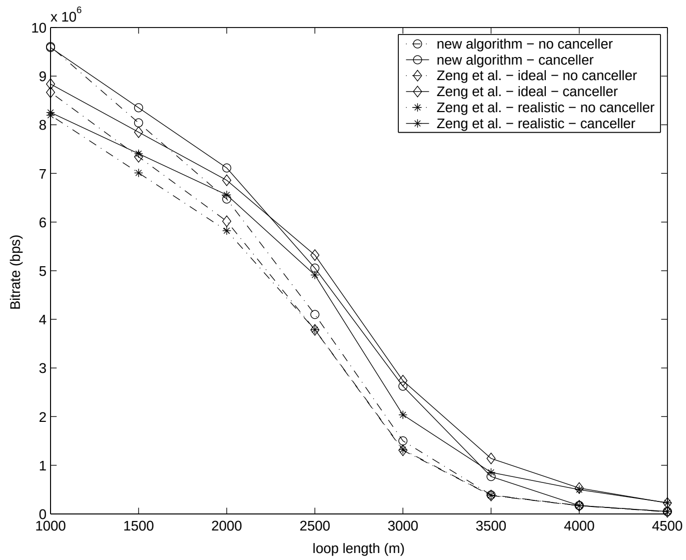

Figure 1: Performance of the crosstalk cancelling algorithms.

ADSL context (under the assumption of perfect channel shortening and white background noise) it is beneficial to choose $\mathbf{S}_1$ in the interferer excess lobe and cancel crosstalk in the main lobe, $\mathbf{S}_2$. Our simulations show that, as a rule of thumb, this also holds for our algorithm in case of longer loops. On shorter loops, there is hardly any performance difference if $\mathbf{S}_1$ and $\mathbf{S}_2$ are chosen in the opposite way. In the simulations, $\mathbf{S}_1$ consists of tones 121 to 209; $\mathbf{S}_2$ contains the tone intervals $[32 : 120]$ and $[210 : 236]$ (using the interferer spectral correlation in the upstream band). The remaining tones 237 to 255 use pure per-tone equalization.

In **Figure 1**, the non-adaptive algorithm of [4][2] is compared with the new adaptive algorithm. The adaptive algorithm uses 1 frequency shift of $f_i = 1040$ kHz and $T = 32$ taps per frequency shift (total of 64 taps for each tone). The non-adaptive algorithm is evaluated with perfect channel shortening (i.e. no residual ISI/ICI) as well as with a more realistic TEQ (32 taps, MMSE-based, unit-norm constraint on the target impulse response) introducing residual ISI/ICI. The achieved bitrate with (solid line) and without (dashed line) crosstalk cancellation is depicted.

The relative performance increase for all algorithms is largest between 2500 and 3500m. For all lengths up to around 3500m, the new adaptive algorithm performs better than the non-adaptive algorithm with realistic TEQ, mainly owing to the use of per-tone equalization. The performance increase, due to crosstalk cancellation, is roughly the same for both algorithms. Note that there is a significant performance loss of the non-adaptive algorithm in the usual case where the channel is not perfectly shortened (diamonds versus stars). Moreover, simulations have shown that including more tones (e.g. from the upstream band) in $\mathbf{S}_1$ of the non-adaptive algorithm with imperfect channel shortening does not necessarily lead to better performance. It should be mentioned that all algorithms regain only a fraction of the performance loss, due to the crosstalker (e.g. at 3000m, almost 4.5Mbps remains lost).

[2] The optimal version with application of the matrix inversion lemma for each block symbol is used.

4 CONCLUSIONS

We have presented an algorithm to mitigate a dominant alien near-end crosstalker such as HDSL, SDSL or HPNA in DMT-based ADSL and VDSL receivers. Recently, Zeng et al. presented a method to tackle the same problem. Both algorithms exploit spectral correlation in the received interfering signal, thanks to oversampling and excess bandwidth. However, our algorithm does not require any prior knowledge (such as transmission and crosstalk channel, noise characteristics, etc.) other than the interferer symbol rate. Moreover, it is fully adaptive, integrates per-tone equalization and crosstalk cancellation and optimizes the signal-to-noise ratio for each tone separately. It is based on the application of so called FREquency-SHift (FRESH) filtering. It is shown that the new adaptive algorithm performs significantly better than the non-adaptive algorithm of [4] over a wide range of loop lengths.

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