DOUBLE TALK CANCELLATION IN ECHO CANCELLED DMT-SYSTEMS

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ABSTRACT

The existing ADSL standard allows for two possible transmission systems: frequency division duplexing (FDD) and echo cancelling (EC). Echo cancelling is particularly attractive for its ability to achieve higher data rates compared with FDD schemes. A disadvantage of EC is that adaptive schemes exhibit slow tracking properties in the presence of a far end signal or double talker. This paper presents an efficient method for updating echo canceller taps in ADSL transceivers in the presence of a far end signal, by effectively cancelling most of the received far end energy. The presented scheme extends existing multicarrier echo cancelling schemes with a feedback of decisions on the far end signal. As a result the convergence is much faster and hence fast tracking properties are possible at a low computational complexity.

1 INTRODUCTION

In a DSL system, two directions of communication are possible: downstream and upstream communication. If transmission in both directions takes place over the same loop, the transmitter and receiver at one end are coupled to the line by a *hybrid*. A perfectly balanced hybrid prevents leakage of transmitted signals into the receiver. However, due to large variations in the subscriber loops, a fixed hybrid can not be exactly balanced for all loops and hence leakage occurs. This leakage is called *echo*.

The ADSL standard [1] allows two different options to reduce the echo present at the receiver: frequency division duplexing (FDD) and echo cancelling (EC). In an FDD system up- and downstream transmission are separated in frequency by steep filters, hence echo is effectively filtered out by the front-end filters. A disadvantage is that a frequency gap (or unused bandwidth) is necessary for non-ideal filters. EC systems allow a smaller frequency gap or even overlapping frequency bands with relaxed filter specifications. The transmitted signal will then cause echo in the received signal and an echo canceller is needed [2].

Several echo cancellation structures for discrete multitone transmission (DMT) transceivers have been studied in literature [3][4][5]. All these structures exploit a common principle: the echo channel is estimated through an adaptive updating process and an emulated version of the echo is subtracted from the received signal. In [3], the *emulation* is performed in time domain, while the *updating* process is mainly performed in the frequency domain. Ho *et al.* applied modifications to this structure in [4] by exploiting the 'circular' aspects of the DMT line code (circular echo synthesis). The circular part of the echo emulation is moved to the frequency domain which reduces the total computational complexity.

Due to temperature variations of the line and the modem front end the echo channel will change over time. Hence, adaptive algorithms like the least mean square (LMS) algorithm are used to track these channel variations [6]. However, when the far end signal is not silenced, which is e.g. the case during duplex transmission, the adaptive scheme exhibits a large excess mean square error (MSE). Although the far end signal is uncorrelated with the transmitted echo reference signal, it will cause a large excess MSE in the adaptation process. A large excess MSE of the echo canceller will degrade the signal to noise ratio (SNR) of the far end signal at the receiver. Hence, the achievable bitrate will be reduced. This is the so called *double talk* problem. A well known solution for this problem is to increase the noise averaging by lowering the stepsize in the adaptation process. However, in most cases this stepsize reduction leads to insufficient tracking and/or slow convergence.

The aim of this paper is to modify the updating part of existing echo canceller structures in order to have faster convergence and tracking in the presence of a far end signal. The energy of the double talker will be lowered to the level of the far end inter-symbol-interference (ISI) plus inter-carrier-interference (ICI) noise. Therefore, a larger stepsize can be used for faster convergence.

The paper is organized as follows. In section 2, the data model and notation is introduced. In sections 3 and 4 the

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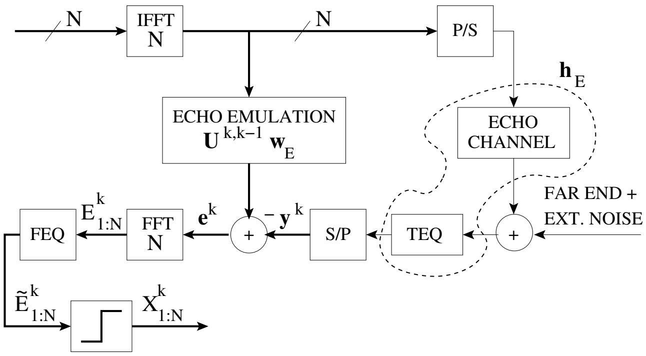

Figure 1: Echo canceling scheme with data model.

double talk canceller is developed for two different types of echo cancellers. Complexity calculations are given in 5. The last sections provide the simulation results and conclusions.

2 DATA MODEL

The echo signal at the receiver can be modelled as the linear convolution of the echo impulse response, $\mathbf{h}_E$, and the modem's own transmitted sequence. This convolution can be written in matrix-vector notation, i.e.

$$\mathbf{y}^k \quad = \quad \mathbf{U}^{k,k-1}\mathbf{h}_E + \mathbf{n}^k \tag{1}$$

with $\mathbf{y}^k = [y_1^k \dots y_N^k]^{\mathrm{T}}$ a received echo symbol with receive FFT size N at symbol period k, $\mathbf{U}^{k,k-1}$ a $N \times N$ Toeplitz matrix of the transmitted echo reference samples and $\mathbf{n}^k$ the sum of the additive noise and the received far end signal. We assume that the time domain equalizer (TEQ), which is commonly used in DSL systems to reduce ISI and ICI [7], is placed in front of the echo canceller. The echo impulse response is assumed to fit into one DMT symbol of N samples[1] and contains the influence of the front end filters, hybrid circuitry and the TEQ. The external noise and double talk signal, $\mathbf{n}^k$, are filtered by the TEQ before echo cancellation is performed, see **Fig. 1**. Without loss of generality, we assume a symmetric rate setup with aligned far end and echo symbol streams. Extensions to asymmetric and/or misaligned setups[2] can easily be made. The echo cancelled received symbol is described by

$$\mathbf{e}^k \quad = \quad \mathbf{y}^k - \mathbf{U}^{k,k-1} \cdot \mathbf{w}_E \tag{2}$$

where $\mathbf{e}^k = [e_1^k \dots e_N^k]^{\mathrm{T}}$ is of length N and $\mathbf{w}_E$ is the M-taps echo channel estimate, zero padded to length N.

3 SIGNAL-DRIVEN ECHO CANCELLER

The optimal MMSE echo canceller minimizes the following cost function:

$$\min_{\mathbf{w}_E} \mathcal{E}\{(e^k)^2\} \approx \min_{\mathbf{w}_E} \|\mathbf{e}^k\|_2^2, \tag{3}$$

[1] This assumption holds in almost all cases

[2] In a misaligned scheme, $\mathbf{U}^{k,k-1}$ depends on the misalignment. To extend the proposed method to the misaligned case it is assumed that the EC has *knowledge* about the misalignment of the transmitted symbols with respect to the received symbols.

with $\mathcal{E}\{\cdot\}$ the expectation operator. Applying Parceval's theorem, the equivalent cost function in the frequency domain can be stated as:

$$\min_{\mathbf{w}_E} \|E^k\|_2^2 = \min_{\mathbf{w}_E} \sum_{i=1}^{N} |E_i^k|^2, \tag{4}$$

with $E^k = [E_1^k \dots E_N^k] = \mathcal{F}_N \cdot \mathbf{e}^k$ of length N, $\mathcal{F}_N$ is a DFT matrix of size $N \times N$ and i is the tone or frequency index. In DMT systems a one tap complex frequency domain equalizer (FEQ) for each tone i is used to compensate for magnitude and phase distortions introduced by the overall far end channel [7]. When considering the frequency domain error of (4) after the one tap FEQ, denoted by B_i for tone i, the cost function becomes

$$\min_{\mathbf{w}_E} \sum_{i=1}^{N} \frac{|B_i \cdot E_i^k|^2}{|B_i|^2} = \min_{\mathbf{w}_E} \sum_{i=1}^{N} \frac{|\tilde{E}_i^k|^2}{|B_i|^2} \tag{5}$$

$$= \quad \min_{\mathbf{w}_E} \sum_{i=1}^{N} \frac{\left|B_i \cdot \mathcal{F}_N(i,:)(\mathbf{y}^k - \mathbf{U}^{k,k-1}\mathbf{w}_E)\right|^2}{|B_i|^2}, \tag{6}$$

where $\mathcal{F}_N(i,:)$ indicates the ith row of the demodulating DFT matrix. In the case of an ideal echo canceller the numerator contains only far end information and external noise. Moreover, the FEQ and TEQ are designed in such a way that this numerator is as close as possible to the transmitted far end symbol on tone i, X_i^k. These transmitted symbols are available after the decision device following the FEQ outputs. Hence, the energy of the far end signal can be reduced by subtracting X_i, i.e.

$$\min_{\mathbf{w}_E} \sum_{i=1}^{N} \frac{\left|B_i \cdot \mathcal{F}_N(i,:)(\mathbf{y}^k - \mathbf{U}^{k,k-1}\mathbf{w}_E) - X_i^k\right|^2}{|B_i|^2}, \tag{7}$$

$$\begin{aligned} = \quad & \min_{\mathbf{w}_E} \|\mathrm{diag}\{B\}^{-1} \cdot \\ & (\mathrm{diag}\{B\}\mathcal{F}_N(\mathbf{y}^k - \mathbf{U}^{k,k-1}\mathbf{w}_E) - X_{1:N}^k)\|_2^2 \end{aligned} \tag{8}$$

with $B = [B_1 \dots B_N]$ a vector of length N containing the FEQs for all tones. Formula (7) effectively corresponds to (6) supplemented with **double talk cancellation**. In this way, the level of the far end signal (double talker) in the squared error is reduced to the level of the external noise. Of course, due to imperfect equalization by TEQ and FEQs some residual far end ISI/ICI will also be present.

The LMS updating formulas for **time domain echo cancelling** can be obtained by calculating the gradient of (8) with respect to $\mathbf{w}_E$ and are given by:

$$\begin{aligned} \mathbf{w}_E^{k+1} \leftarrow & \mathbf{w}_E^k + \mu(\mathrm{diag}\{B\}^{-1}\mathrm{diag}\{B\}\mathcal{F}_N\mathbf{U}^{k,k-1})^{\mathrm{H}} \cdot \\ & \mathrm{diag}\{B\}^{-1}(\mathrm{diag}\{B\}\mathcal{F}_N(\mathbf{y}^k - \mathbf{U}^{k,k-1}\mathbf{w}_E^k) - X_{1:N}^k) \end{aligned} \tag{9}$$

$$\mathbf{w}_E^{k+1} \leftarrow \mathbf{w}_E^k + \mu \mathbf{U}^{k,k-1^{\mathrm{T}}} \mathcal{I}_N \mathrm{diag}\{B\}^{-1}(\tilde{E}_{1:N}^k - X_{1:N}^k) \tag{10}$$

where $\{\cdot\}^{\mathrm{H}}$ and $\{\cdot\}^{\mathrm{T}}$ denote complex conjugate transpose and transpose resp., $\mathcal{I}_N$ is an inverse DFT matrix of size N and μ is the stepsize.This equation indicates that instead of using a time domain error in the update,the error is transformed to the frequency domain. After rotating and scaling the frequency domain error by the FEQs, the far end signal is removed. Finally, the inverse FEQ operation is applied and the error is transformed back to time domain. In case time domain echo cancelling is performed, it is clear that some extra computational complexity is added.

4 DATA-DRIVEN ECHO CANCELLER

In [3], a data-driven multitone echo canceller was developed, where the *filter updating part is moved to the frequency domain, while the filtering is performed in the time domain*. Ho *et al.* use the same principle in [4], but construct a circular echo synthesis by which the filtering part is shared over time and frequency domain. Since the method used for filtering is not relevant in this paper, we continue by using filtering completely in time domain to keep notation simple.

Here also the double talk cancelling can be applied, but in this case with almost no extra cost. The updating formulas without double talk canceller are

$$W_{E,1:N}^{k+1} \leftarrow W_{E,1:N}^{k} + \mu \mathrm{diag}\{U_{1:N}^{k^*}\} \cdot \mathcal{F}_N(\mathbf{y}^k - \mathbf{U}^{k,k-1}\mathbf{w}_E^k) \quad (11)$$

where $\{\cdot\}^*$ denotes complex conjugation, $W_{E,1:N} = \mathcal{F}_N \mathbf{w}_E$ and $U_{1:N}^k$ are the transmitted echo reference symbols in the frequency domain [3]. Adding double talk cancelling analogous to (9), results in

$$W_{E,1:N}^{k+1} \leftarrow W_{E,1:N}^{k} + \mu \mathrm{diag}\{U_{1:N}^{k^*}\}\mathrm{diag}\{B\}^{-1} \cdot (\mathrm{diag}\{B\}\mathcal{F}_N(\mathbf{y}^k - \mathbf{U}^{k,k-1}\mathbf{w}_E^k) - X_{1:N}^k) \quad (12)$$

$$W_{E,1:N}^{k+1} \leftarrow W_{E,1:N}^{k} + \mathrm{diag}\{\mu_{1:N}\}\mathrm{diag}\{U_{1:N}^{k^*}\} \cdot (\tilde{E}_{1:N}^k - X_{1:N}^k) \quad (13)$$

Here, the scaling with the inverse of the FEQ coefficients can be absorbed by a tone dependent stepsize, $\mu_{1:N}$. **Fig. 2** depicts the double talk principle for a data-driven echo canceller. The update error is constructed by taking the difference between input and output of the decision device. The updating of the echo transfer function estimate is done in the frequency domain, while the filtering part is in the time domain. Hence, an extra IFFT operation is needed to transform the echo coefficients back to time domain. In cases where $M < N$ an extra FFT operation can be performed after zeroing $N - M$ time domain taps [3]. This operation is not depicted in the figure.

The goal of the receiver is to detect the transmitted far end symbols as accurately as possible. This detection process will be distorted by external noise, far end ISI/ICI and residual echo. A good design criterion for the echo canceller is to make sure that the residual echo is much lower than the external noise plus ISI/ICI. As a result, residual echo will not be the dominant noise source. The residual echo seen by the receiver can be expressed as the excess MSE of the echo cancelling scheme like in (11).

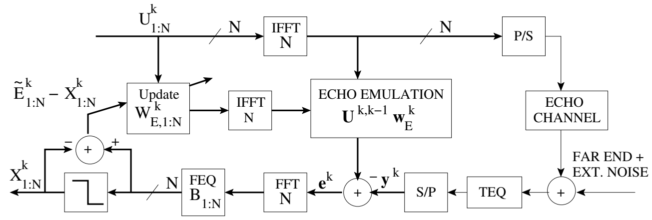

Figure 2: Echo cancelling scheme with double talk cancellation.

Cioffi *et al.* have shown in [3] that in case all tones are excited, this excess MSE per tone i is given by

$$\xi_{\mathrm{excess},i} = \frac{\mu\sqrt{\frac{M}{N}}}{2}\frac{1}{1-\gamma_i}\sigma_{u,i}^2\sigma_{n,i}^2 \quad (14)$$

with γ_i a tone dependent constant, $\sigma_{u,i}^2$ and $\sigma_{n,i}^2$ the echo transmit power and noise power per subchannel. Hence, when the noise power is increased due to double talk, the stepsize should be lowered to reach the same excess MSE as in the case where the far end signal is silenced. With double talk cancellation, the stepsize can be increased, resulting in better tracking performance of the echo channel.

5 COMPLEXITY CALCULATIONS

In this section the complexity of the double talk canceller is calculated. Only the complexity added by the double talk cancellation algorithm is given. Complexity figures for the echo canceller itself can be found in [3][4][5]. It should be noted that in DSL DMT-systems tones appear in complex conjugate pairs. Therefore only half of the tones are considered in the complexity calculations. The complexity is expressed in the number of real additions and real multiplications per symbol period; e.g. a complex multiplication is counted as four real multiplications.

1. signal-driven echo canceller
   - difference complex input and output of decision device: $2\frac{N}{2}$ add./symbol
   - complex multiplication with inverse FEQ: $4\frac{N}{2} + 2\frac{N}{2}$ mult./symbol, $2\frac{N}{2} + \frac{N}{2}$ add./symbol, $2\frac{N}{2}$ div./symbol
   - transformation of frequency error to time domain with complex to real ifft: $N \log \frac{N}{2}$ mult./symbol and $N \log \frac{N}{2}$ add./symbol
2. data-driven echo canceller
   - difference complex input and output of decision device: $2\frac{N}{2}$ add./symbol
   - computation[3] $\mu_{1:N}$: $4\frac{N}{2}+2\frac{N}{2}$ mult./symbol, $2\frac{N}{2}+\frac{N}{2}$ add./symbol, $2\frac{N}{2}$ div./symbol

[3]It is assumed that the FEQ is also adaptive, such that $\mu_{1:N}$ is recomputed in every iteration.

The multiplication with the one tap FEQs is not taken into account, since it belongs to the equalization operations of the modem. The total complexity for the double talk canceller for the signal-driven echo canceller becomes $(3+\log\frac{N}{2})N\cdot\frac{F_s}{N+\nu}$ real mult., $(2.5+\log\frac{N}{2})N\cdot\frac{F_s}{N+\nu}$ real add. and $N\cdot\frac{F_s}{N+\nu}$ div. per second, with F_s the sampling rate and ν the size of the cyclic prefix. Similarly, for the data-driven echo canceller, the total number of operations equals $3N\cdot\frac{F_s}{N+\nu}$ real mult., $2.5N\cdot\frac{F_s}{N+\nu}$ real add. and $N\cdot\frac{F_s}{N+\nu}$ per second. Since these complexity figures are small compared to the complexity of the echo canceller itself (typically $2-4\%$), the double talk cancelling comes almost at no extra cost.

6 SIMULATION RESULTS

The results above are verified by simulations at central office (CO) for an ADSL 26awg line of 3000 m. The far end modem transmits the upstream tones $8-32$, while the echo reference signal is a downstream signal with tones $34-256$. In these simulations, each tone transmits a 4-QAM signal constellation. The downstream and upstream signal transmit with -40 dBm/Hz and -38 dBm/Hz respectively. To ensure convergence the echo reference signal contains 20 extra tones with 20 dB lower power. The external additive noise is white and Gaussian at -140 dBm/Hz. The transmit block length of the far end and echo IFFT are resp. 128 and 512 (see [4] for details on algorithm modification for such cases). The receive FFT is of size 128. The true echo channel is of size 512 samples at 2.2 MHz, while the number of used echo canceller taps is $M=300$.

Fig. 3 illustrates the excess MSE after an initialization phase in the presence of a double talker. The excess MSE of the data-driven echo canceller summed over all tones as a function of the stepsize is depicted after convergence, with and without double talk canceller, i.e.

$$\hat{\xi}_{\text{excess}}=\sum_{i=1}^{N/2+1}|\mathcal{F}_N(i,:)(\mathbf{U}^{k,k-1}\mathbf{h}_E-\mathbf{U}^{k,k-1}\mathbf{w}_E)|^2 \quad (15)$$

The optimal excess MSE is obtained, using (14). The external noise (after TEQ) together with the remaining ISI/ICI summed over all tones equals -118 dBm, while the received far end energy (after TEQ) is -69 dBm for this simulation. Hence, if the residual echo has to be 10 dB below the external noise floor, the stepsize can be 2^{10} with double talk cancellation and much smaller than 2^1 (optimal stepsize is not depicted on the figure) without double talk canceller. Since a significantly larger stepsize can be used in the case double talk cancellation is incorporated, also tracking performance will be greatly improved.

7 CONCLUSIONS

In this paper we presented a low complexity solution for double talk cancellation in ADSL echo cancelled systems. The proposed method allows to use a larger stepsize in the adaptation process resulting in fast tracking and/or convergence in the presence of a far end signal, and hence, increased robustness against, e.g., temperature changes in the echo-path.

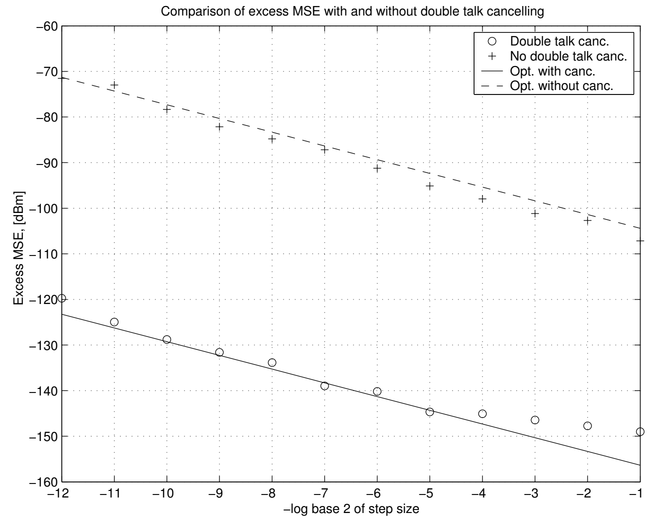

Figure 3: Performance of double talk cancellation compared to no double talk cancellation.

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