

ADAPTIVE FEEDBACK CANCELLING IN SUBBANDS FOR HEARING AIDS

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ABSTRACT

In this paper a hearing aid concept with recruitment of loudness compensation and acoustic feedback cancellation is presented. Special consideration is given to the acoustic feedback canceler which uses only the available (e.g. speech) input signal for adaptation. In principle, the feedback canceler is adapted to the feedback path in the transform domain using a power-normalized least mean square (LMS) algorithm. The transformation into uniform subbands is based on an *augmentation* of the modulated lapped transform (MLT). Together with the hearing-loss compensating forward filter the proposed feedback canceler is computationally very efficient.

1. INTRODUCTION

The main purpose of a hearing aid device (Fig. 1) is to amplify the acoustical signal $v_c(t)$ in order to compensate for the hearing loss of the impaired listener. Due to acoustic and mechanical feedback, the microphone not only picks up the input signal $v_c(t)$, but also the feedback signal $y_c(t)$. In most practical implementations [10], the hearing-loss compensator introduces a time delay T_c due to signal processing.

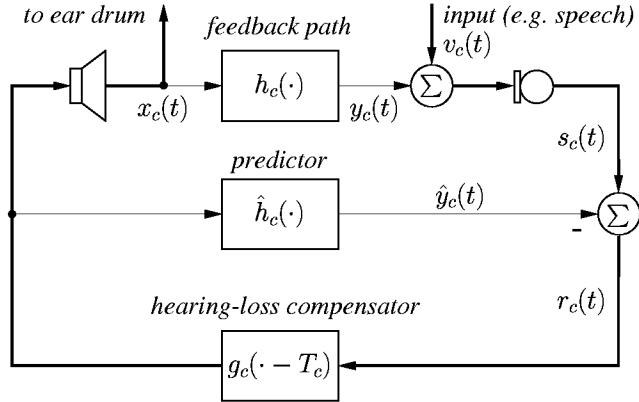


Figure 1: Hearing-aid system with a hearing loss compensator, $g_c(\cdot - T_c)$, and a feedback canceler (predictor), $\hat{h}_c(\cdot)$.

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1.1. Proposed System

The proposed system (Fig. 2) has a clipping device after the compensator, with a clipping level below the saturation level of the loudspeaker, in order to prevent the *linear* predictor from having to estimate a *nonlinear* feedback path [4]. Including D/A- and A/D-converters, microphone and loudspeaker, the feedback path is assumed to be modeled by a FIR filter, $h(\cdot)$, of length N . The block index is denoted by m .

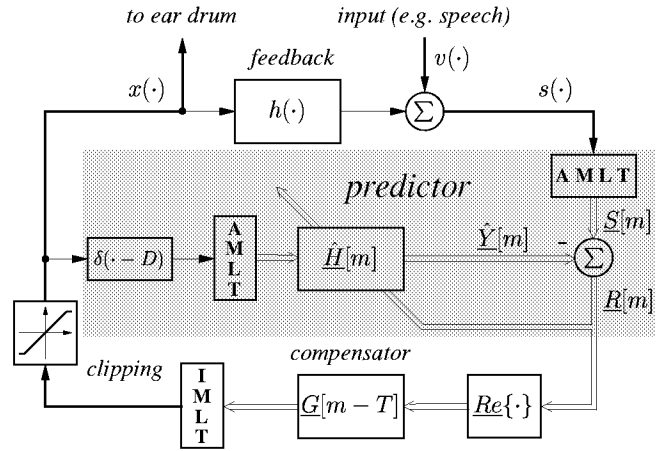


Figure 2: Proposed hearing-aid system

Transform-domain schemes allow a computationally efficient implementation of a digital hearing aid [5, 10]. The proposed system uses three modulated lapped transforms (MLT) and incorporates the essential clipping device (which is missing in [5]).

The advantage of signal processing in the MLT-domain is the computational efficiency and the ability to build any desired band spacing (e.g. auditory filter bank for a psychoacoustic model [6]) by means of hierarchical lapped transforms (HLT).

The so-called “overlap-save” technique used in [10] allows a perfect modeling of impulse responses up to half the DFT-length (50% overlap). The MLT subbands have a larger rejection of the sidelobes (Fig. 3), which allows to subtract the feedback prediction $\hat{Y}[m]$ in the transform domain with sufficient accuracy, although the predictor does not convolve $x(\cdot)$ with $\hat{h}(\cdot)$ perfectly. The delay D incorporates a shift to the “center of gravity” of the impulse response.

1.2. Hearing impairment

People with loudness recruitment have a compressed dynamic range between the sound-pressure levels corresponding to threshold and discomfort. This highly frequency-dependent phenomena requires a filter for compensation which depends on the input signal [6]. To achieve the high levels of gain required for profound hearing impairment, the subbands of the compensator must have a large and fast increasing sidelobe attenuation.

2. AUGMENTED MLT (AMLT)

2.1. The MLT

The basis functions of the MLT have length $L=2M$ and are defined with $k \in \{0, \dots, M-1\}$ representing the subband index:

$$p_k(\cdot) = h(\cdot) \cos \left[\left(k + \frac{1}{2} \right) \left(\cdot + \frac{M+1}{2} \right) \frac{\pi}{M} \right]. \quad (1)$$

$h(\cdot) := -\sin \left[\left(\cdot + \frac{1}{2} \right) \frac{\pi}{L} \right]$ is the only possible low-pass prototype satisfying perfect reconstruction (PR) and polyphase normalization [7]. The MLT basis functions have a sidelobe attenuation of more than 23dB compared to 13dB of the DFT. The sidelobe attenuation of the MLT increases by 40dB/decade and has more than 60dB attenuation for normalized frequencies Ω which are off the center frequency Ω_C by more than $\pi/4$ ($M=64$) (Fig. 3).

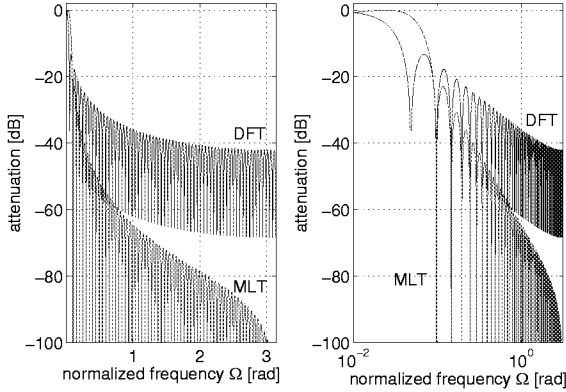


Figure 3: Comparison of DFT and MLT subbands (first band, $M = 64$)

The value of the k th subband signal in the m th block of a MLT is given by

$$X_k[m] = \sqrt{\frac{2}{M}} \sum_{r=0}^{L-1} x_m(r) p_k(r) \quad (2)$$

where $x_m(r) = x(mM - L + 1 + r)$.

2.2. Definition of the AMLT

We propose to augment the MLT basis function $p_k(\cdot)$ by a complex part to have phase information for the predictor. We therefore define the basis function $p_k^C(\cdot)$ of the AMLT as

$$p_k^C(\cdot) := p_k(\cdot) + j \cdot h(\cdot) \sin \left[\left(k + \frac{1}{2} \right) \left(\cdot + \frac{M+1}{2} \right) \frac{\pi}{M} \right]. \quad (3)$$

With the definition of the AMLT, the k th subband signal $X_k^{C,0}[m]$ is given with $e^{j\cdot} = \cos(\cdot) + j \sin(\cdot)$ as

$$X_k^{C,0}[m] = \sqrt{\frac{2}{M}} \sum_{r=0}^{L-1} x_m(r) h(r) e^{j \frac{\pi}{M} (k + \frac{1}{2}) (r + \frac{M+1}{2})} \quad (4)$$

where $\text{Re}\{X_k^{C,0}[m]\} = X_k[m]$. Thus, we have phase information from $x(\cdot)$ for the predictor and we can use the MLT in the forward filter, where no phase information is needed [1]. We see from the definition of the AMLT that its low-pass prototype $h(\cdot)$ is identical to that of the MLT.

With $W_L = e^{j \frac{2\pi}{L}}$ and $u_m(r) = x_m(r) h(r)$ we get:

$$X_k^{C,0}[m] = \underbrace{\sqrt{\frac{2}{M}} \left(\sum_{r=0}^{L-1} u_m(r) W_L^{r(k + \frac{1}{2})} \right)}_{X_k^{C,1}[m]} \underbrace{W_L^{(k + \frac{1}{2})(\frac{M+1}{2})}}_{\phi(k)}. \quad (5)$$

The interpretation of $X_k^{C,1}[m]$ leads to the following figure:

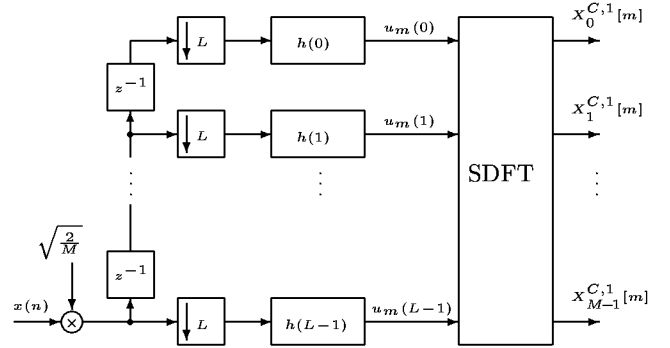


Figure 4: Analysis filter bank (without the rotations $\phi(k)$): Note, that there are only M bands in the AMLT.

In Fig. 4, we defined the *symmetric* DFT (SDFT) as:

$$\text{SDFT: } F[i] = \sum_{n=0}^{N-1} f[n] W_N^{n(i + \frac{1}{2})}, \quad (6)$$

with $W_N = e^{j \frac{2\pi}{N}}$. The inverse AMLT (IAMLT) can be found similarly. Fig. 4 shows the close connection of the AMLT and a DFT filter bank.

It can be shown by FFT pruning, proper placing of the rotations $\phi(k)$ and manipulation of the twiddle factors [8] that the computational complexity of a AMLT is 20% larger compared to a FFT (both $M=64$ bands). Thus, comparing the system in [10] with four FFTs and the proposed system with three AMLTs we achieve a reduction of the computational complexity of more than 10%.¹

3. FEEDBACK CANCELLING

3.1. Hearing-Aid impulse responses

In Fig. 5 impulse responses $h_i(\cdot)$ of two behind-the-ear hearing aids (BTE) and of two ITEs (in-the-ear hearing aid) out

¹For simplicity we did not encounter the projections [2] necessary to eliminate the temporal aliasing (FFT case).

of 29 impulse responses are plotted. All impulse responses have been normalized to $\|h_i(\cdot)\|^2 = 1$.

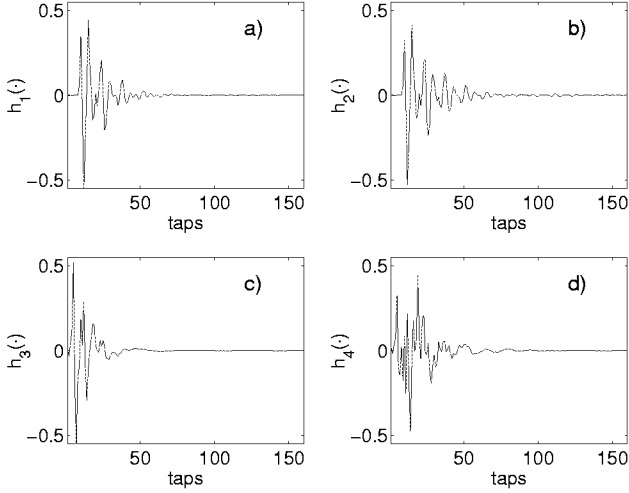


Figure 5: a) BTE with no obstacle; b) BTE with wooden board 10 cm parallel to ear; c) ITE with no obstacle; d) ITE with hand covering ear

If we define the effective length $\tilde{N}_i$ of each impulse response by $\tilde{N}_i = \arg \min_N \left\{ \sum_{r=0}^N h_i^2(r) > 0.9 \right\}$, we get the following lengths $\tilde{N}_i$ for the four hearing aids: 25, 27, 13, 25. We see that the BTE impulse responses have a length of roughly 2 times the length of a normal ITE, whereas the pathological case d) has a length approximately equal to a BTE.

3.2. Predictor $\hat{H}[m]$

For simplicity and in contrast to [3] we have chosen $\hat{Y}[m] = \mathbf{X}^H[m] \hat{H}[m]$ where $\mathbf{X}[m]$, the predictor input matrix, is given by $\mathbf{X}[m] = \text{diag}(\mathbf{X}^{C,0}[m])$. The weights are updated independently by the normalized LMS algorithm [9]:

$$\hat{H}[m+1] = \hat{H}[m] + \boldsymbol{\mu}[m] \cdot \mathbf{X}^H[m] \mathbf{R}[m], \quad (7)$$

with the diagonal step-size matrix $\boldsymbol{\mu}[m]$ given by $\mu_{i,i}[m] = \mu_0 / \mathcal{E}\{|\mathbf{X}_i^{C,0}[m]|^2\}$ [2] and $\mathbf{R}[m] = \mathbf{S}[m] - \hat{Y}[m]$. In practice $\mathcal{E}\{\cdot\}$ is replaced by short-time averaging with the forgetting factor γ (integrator gain [7]).

3.3. Delay D

We calculated the “optimal” D ’s (in the sense of best feedback cancelling performance) by transforming the signal $y(\cdot) = x(\cdot) * h(\cdot)$ and $x(\cdot - D)$ into the AMLT-domain and solving $\mathbf{Y}[m] = \mathbf{X}^H[m] \hat{H}$ for $\hat{H}$ in the least square sense over several blocks m . The error level is defined as $\frac{\|r(\cdot)\|}{\|x(\cdot)\|}$ with the error signal $r(\cdot)$ being the IAMLTL of $\hat{H}[\cdot]$.

As M increases (Fig. 6) the error level gets smaller and flattens, thus the influence of D on the performance of the predictor is neglectable for $\tilde{N} \ll M$. The input signal $x(\cdot)$ was a 10th-order AR process with a spectral power density

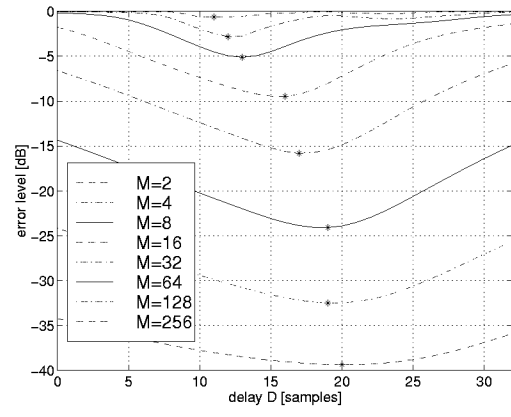


Figure 6: Error level in function of delay D and number of subbands M for $h_1(\cdot)$; * depicts the optimal D

equal to that of a long term average of speech.

The optimal delay D for $\log\left(\frac{\tilde{N}_i}{M}\right) \approx 0$ ($\tilde{N}_i$ and M have the same dimension) can be approximated by the “center of gravity” (COG) of the impulse response:

$$\text{COG}_i = \frac{\sum_r r \cdot h_i^2(r)}{\|h_i(\cdot)\|^2}. \quad (8)$$

i	COG _{i}	Optimal delay $D = f(M)$, $x(\cdot) = \text{white noise}$					
		$M=8$	$M=16$	$M=32$	$M=64$	$M=128$	$M=256$
1	15.65	13	15	17	18	19	
2	17.91	13	16	18	20	21	
3	8.74	7	8	10	10	11	
4	15.84	14	14	15	17	18	

Table 1: Comparison of COG and the optimal D

3.4. Simulation results

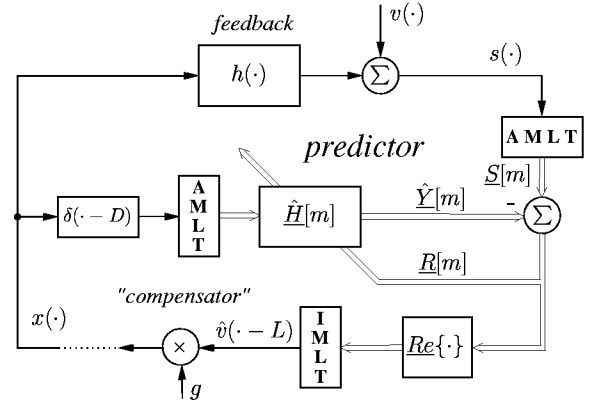


Figure 7: Simulation set-up

We made two different simulations both with feedback path $h_1(\cdot)$. In the first simulation (Fig. 8) we measured the feedback cancellation error level $\mathcal{E}\left\{\left|\frac{\hat{v}(\cdot-L)}{x(\cdot)}\right|\right\}$ with $v(\cdot) \equiv 0$, $x(\cdot)$ being a 10th-order AR process, $\mu_0 = 0.02$ and $\gamma = 0.9$. With this setup the error level is at its steady-state within 1sec (at 16kHz sampling rate). In contrast to Fig. 7, $\hat{v}(\cdot-L)$ was calculated by the IAMLTL of $\hat{H}[\cdot]$.

In the second setup we closed the loop, i.e. $x(\cdot) = g \cdot \hat{v}(\cdot - 3M)$ ($3M \equiv 12\text{ms}$ = transformation delays and one block for processing). We set g 10dB above the critical gain g_{crit} (where the open loop gain of the system without feedback canceler is unity at its critical frequency). μ_0 was 0.009, $\gamma = 0.98$, the input $v(\cdot)$ was a 10th-order AR process and the error level was $\mathcal{E}\left\{\left|\frac{\hat{v}(\cdot - 3M) - v(\cdot)}{v(\cdot)}\right|\right\}$. All curves are the ensemble average $\mathcal{E}\{\cdot\}$ of 2000 runs. M was chosen to be 64 which gives enough flexibility to model an auditory filter bank and having error levels below -20dB.

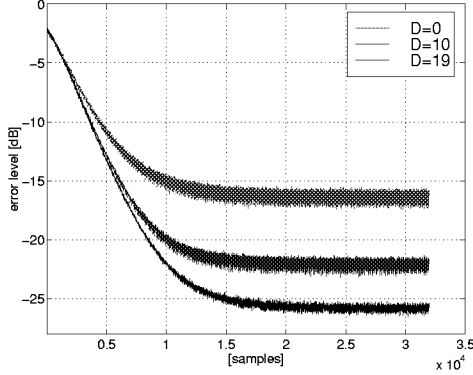


Figure 8: Simulations results of the open loop system

The difference of the error levels (Fig. 8) for $D=0$ and $D=19$ is $\approx 10\text{dB}$, which corresponds exactly to the difference in Fig. 6 ($M=64$). Thus, D is an important design factor for the predictor performance as long as $\log\left(\frac{\tilde{N}_i}{M}\right) \approx 0$.

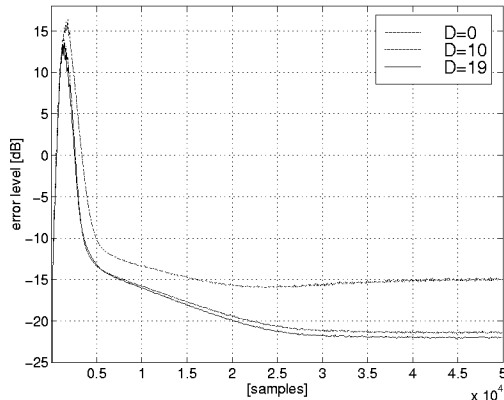


Figure 9: Simulations results of the closed loop system

The steady-state error level in Fig. 9 depends less on D because of the “peakiness” of the transfer function of the impulse response as long the gain g is smaller than the error level of the predictor (Fig. 8). In the first 1500 samples we see that the closed system is unstable but adapts to the stable region. An other possibility to achieve a similar performance with MLT is to use filter partitioning and large oversampling ratio which would be extremely costly. The start-up performance and the tracking behavior can be improved by introducing a stability loop-gain measure and a step-size control [4].

4. SUMMARY

We have presented a hearing aid concept with recruitment of loudness compensation and acoustic feedback cancellation. We used an augmented MLT with a delay to achieve the best feedback cancellation performance. We showed that the delay D can improve the system performance by up to 10dB. The implementation is efficient because of the connection of the AMLT and the FFT. The computational complexity is less than the system in [10] and with a greater flexibility in designing the forward filter (compensator), e.g. modeling of the auditory filters.

5. ACKNOWLEDGMENT

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6. REFERENCES

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