LOW-DELAY HEARING AID BASED ON COCHLEAR MODEL WITH NONUNIFORM SUBBAND ACOUSTIC FEEDBACK CANCELLATION

Maxim Vashkevich[1], *Elias Azarov*[1] *and Alexander Petrovsky*[2]

[1]Department of Computer Engineering, Belarusian State University
of Informatics and Radioelectronics 6, P. Brovky str., 220013, Minsk, Belarus
[2]Department of Digital Media and Computer Graphics, Bialystok University of Technology,
ul. Wiejska 45A, 15-351 Bialystok, Poland,
email: a.petrovsky@pb.edu.pl

ABSTRACT

The paper presents a hearing aid (HA) system based on a low-delay cochlear filter bank derived from the discrete cochlear model. The spectral gain shaping method (SGSM) that includes dynamic range compression, hearing loss compensation and noise reduction is applied in perceptually matched frequency bands. The acoustic feedback cancellation is implemented as an off-the-forward path scheme and does not introduce an additional delay to the forward path. The cancellation itself is made by adaptive filtering of non-uniform subband signals obtained from an oversampled warped cosine-modulated filter bank.

Index Terms— Hearing aid, acoustic feedback cancellation

1. INTRODUCTION

Hearing aids (HA) perform frequency dependent amplification of the incoming signal to improve speech intelligibility for hearing impaired people. The main difficulty in HA design is the low-delay requirement. In [1] has been found that long forward path delays are not desirable due to the comb filter effect. It occurs when the processed sound is combined at the eardrum with the unprocessed sound, which directly travels through the HA's vent. It is stated that even delays as short as 4–8 ms are detectable and reduce perceived sound quality [2]. In modern HAs signal processing is usually performed in frequency subbands introducing analysis-synthesis delays in the forward path. Many research efforts have been focused on this problem [1, 3], however the delays of these solutions are still high (6-8 ms). Good low-delay filtering schemes based on peaking filters [4] and side-branch processing [2] has been recently proposed.

Work was supported by Bialystok University of Technology under the grant S/WI/408.

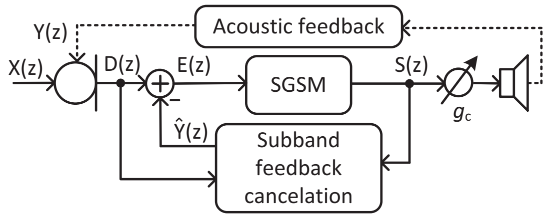

Fig. 1. Low-delay hearing aid system

In the present paper we use a low-delay (less than 4 ms) filtering based on the cochlear filter bank (FB) [5] that is implemented as a set of parallel second-order tunable band-pass filters [6]. The main component of the presented HA system is a low-delay spectral gain shaping method (SGSM). The SGSM combines the following procedures: dynamic range compression (DRC), noise reduction and hearing loss compensation. Another important component of the HA is a subband acoustic feedback cancellation (AFC) system that does not increase the forward path delay. The AFC system employs oversampled warped cosine-modulated filter banks (WCMFB) to decompose the signals into nonuniform subband components [7].

The paper is organized as follows. Section 2 gives a description of all components of the proposed HA, including the cochlear filter bank, subband AFC system and SGSM block. In section 3 computational complexity of the proposed low-delay HA is discussed. Some additional considerations and concluding remarks are given in section 4.

2. STRUCTURE OF HEARING AID SYSTEM

2.1. System overview

The block diagram of the proposed low-delay hearing aid (figure 1) includes the SGSM block and off-the-forward path subband AFC system. The broadband gain function g_c is used to adjust the output sound level being available to the user as a volume control. Dynamic range compression, hearing loss compensation and noise reduction constitute spectral gain shaping block that is implemented as a single computational framework which will be discussed in section 3. Signal processing in SGSM block is performed in frequency subbands using the cochlear FB.

2.2. Cochlear filter bank

Cochlear FB is used in the HA system for the following reasons: i) it decomposes the incoming signal into perceptually matched subband components; ii) cochlear FB has a low group delay – an especially important property for HA; iii) it is computationally efficient since it can be implemented using fourth-order IIR filters.

In the proposed HA system a 22-channel cochlear filter bank is used. These filters simulate the cochlea behavior and are designed on the basis of the 2D nonlinear cochlear model for discrete space and time:

$$y_k[n] + b_{1,k} y_k[n-1] + b_{2,k} y_k[n-2] = A_k a_{0,k} \left(u_s[n] - u_s[n-2] \right),$$

where $b_{1,k}$, $b_{2,k}$, $a_{0,k}$ and A_k – parameters that are determined on the basis of physical characteristics of the basilar membrane [5]. The model is transformed into a bank of digital bandpass filters with transfer function

$$H_k(z) = A_k \frac{a_{0k}(1 - z^2)}{1 + b_{1k} z^{-1} + b_{2k} z^{-2}}.$$

The center frequency (ω_{0k}) and 3 dB bandwidth ($\Delta\omega_{0k}$) of k-th cochlear filter can be obtained as follows

$$\cos\omega_{0k} = -b_{1k}/(1 + b_{2k}), \quad \Delta\omega_{0k} = 2(1 - b_{2k})/(1 + b_{2k}).$$

Cochlear filters can be efficiently implemented using IIR filter structure with tunable bandwidth and center frequency presented in [6]:

$$H(z) = a_0 \frac{1 - z^{-2}}{1 + (a_0 - 1) g z^{-1} + (1 - 2a_0) z^{-2}},$$

where the coefficients a_0 and g only depend on the bandwidth $\Delta\omega$ and center frequency ω_0 respectively (T is the sampling period):

$$a_0 = (\Delta\omega T)/(2 + \Delta\omega T), \quad g = 2\cos\omega_0 T.$$

In order to improve selectivity of the system, the cochlear filters are represented by a cascade of two identical second order sections, as described above. The magnitude response of the 22-channel cochlear FB used in the HA system is shown in figure 2. The group delay of the filter bank does not exceed 4 ms.

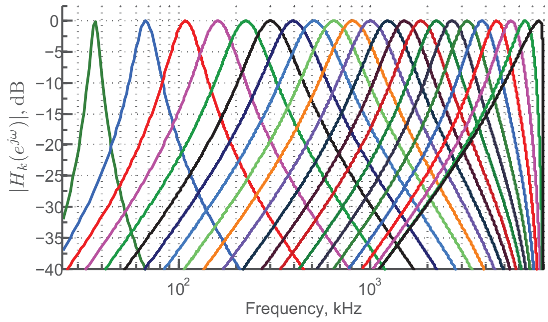

Fig. 2. Magnitude response of cochlear filter bank

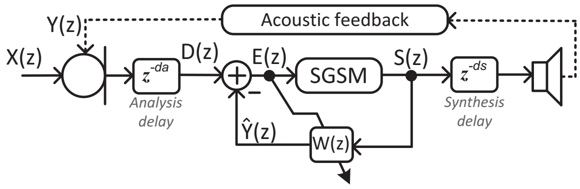

Fig. 3. Straightforward implementation of subband AFC system

2.3. Subband acoustic feedback cancellation system

Acoustic feedback occurs when the aid's speaker produces an acoustic signal that leaks back to the microphone. The normalized adaptive least-mean-square (NLMS) algorithm is a popular technique to address this problem. It has been shown that performance of NLMS can be considerably improved (regarding convergence rate and computational complexity) by subband signal decomposition with a filter bank [8]. However straightforward incorporation of a filter bank in the feedback cancellation branch adds an additional analysis/synthesis delay to the forward path (figure 3). In the figure analysis and synthesis delays are denoted as da and ds respectively.

In order to avoid forward path extension we propose an alternative scheme – figure 4. The subband AFC system is divided into two parts: the adaptive filtering branch and the coefficients estimation branch. The filtered signal $\hat{Y}(z)$ is

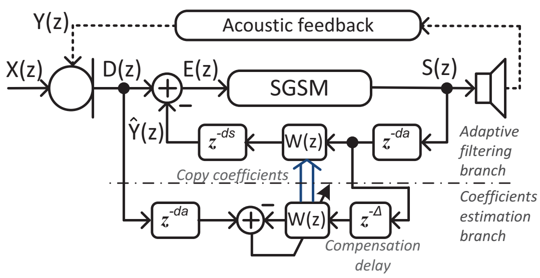

Fig. 4. Off-the-forward path subband AFC system

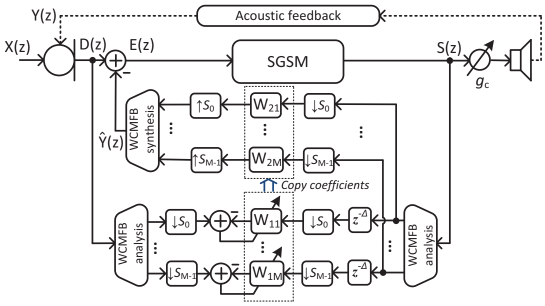

Fig. 5. Subband acoustic feedback cancellation system

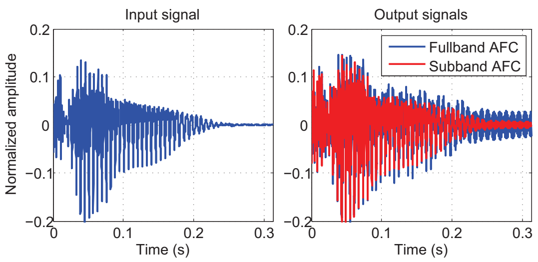

Fig. 6. Convergence of fullband and subband AFC

delayed relatively to the desired signal $D(z)$ by Δ samples where $\Delta = da + ds$. The delay is compensated by prediction of the feedback signal. It is done by adding the compensation delay into the coefficients estimation branch. The feedback signal can be predicted with high accuracy because of its deterministic (periodic) nature.

Using the scheme presented above the subband AFC system is proposed (figure 5). Subband decomposition is carried out using 8-channel oversampled WCMFB. The filter bank is obtained by applying all-pass transform to uniform CMFB [7]. The magnitude response of the employed filter bank is given in figure 7.

The WCMFB provides a great flexibility in design. The bandwidth of filters depends on the single all-pass transformation parameter. Also the WCMFB has an efficient implementation based on polyphase representation of the prototype

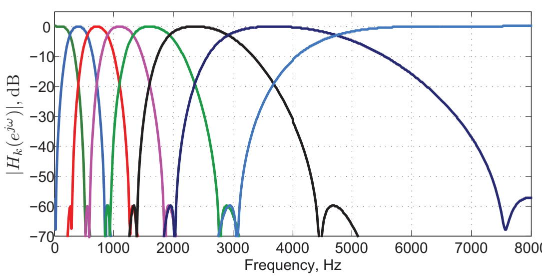

Fig. 7. Magnitude response of oversampled WCMFB

filter and fast DCT-4 algorithm [9]. Using the rule presented in [7] the following subsampling ratios for the filter bank were obtained $S_k = \{19, 9, 6, 4, 2, 1, 1, 1\}$. The total oversampling factor $\mathcal{O} = \sum_{k=0}^{7} \frac{1}{S_k} \approx 4.08$ that is comparable with the design presented in [3].

The output signals of fullband and subband AFC systems are shown in figure 6. It can be seen that subband decomposition increases convergence speed and reduces the level of residual oscillations.

2.4. Spectral gain shaping method

2.4.1. *Noise reduction algorithm*

Noise reduction is implemented using the psychoacoustically motivated spectral weighting rule [10]. The algorithm has adjustable parameter $\zeta = 10^{-\mathrm{RL}/20}$ that defines the desired residual noise level RL in dB. Usually power spectral density (PSD) of noisy speech is estimated using smoothed version of the squared amplitude spectrum. We perform estimation of PSD using the cochlear FB outputs and assume that PSD within each passband of cochlear filters is constant.

In order to estimate PSD of noise $\mathbf{R}_n$ a computationally efficient and robust noise estimation algorithm based on modified version of the Minima Controlled Recursive Averaging (MCRA) is used [10]. PSD is estimated in the subbands by averaging past spectral power values $\mathbf{R}_e(n)$ and using smoothing parameters that are adjusted by the signal presence probability. Weighting factors are updated every 4 ms using the algorithm outlined in table 1.

Table 1. Noise reduction algorithm

Sample processing iterations (for $n = 1, 2 \dots$)

$$\mathbf{R}_e(n) = [R_e(0, n), R_e(1, n) \dots R_e(K-1, n)]^T$$
$$\mathbf{E}(n) = [e_0^2(n), e_1^2(n) \dots e_{K-1}^2(n)]^T$$
$$\mathbf{R}_e(n) = \gamma \mathbf{R}_e(n-1) + (1-\gamma)\mathbf{E}(n)$$

Block processing iterations ($n = R, 2R \dots$)

$$\mathbf{R}_n(n) = \mathrm{MCRA}(\mathbf{R}_e(n))$$
$$\mathbf{R}_T(n) = \mathbf{R}_n(n) \odot (\mathbf{H}^{\mathrm{opt}}(n) - \zeta)^2$$
$$\mathbf{H}^{\mathrm{JND}}(n) = \min(\sqrt{\mathbf{R}_T(n)/\mathbf{R}_n(n)} + \zeta, 1)$$

- $\mathbf{H}^{\mathrm{opt}}(n)$ is equal to the Wiener-filter solution
- $\odot$ denotes element-by-element multiplication
- JND stands for Just Notable Distortion
- R is block size in samples

2.4.2. *Noise reduction performance evaluation*

In performance evaluation the AFC and DRC algorithms were disabled. Four male and female speech signals of duration

Table 2. Performance of the noise reduction system

Before noise reduction					
SNR	20 dB	15 dB	10 dB	5 dB	0 dB
SII	0.73	0.62	0.52	0.47	0.37
After noise reduction (RL = 15)					
SNR	28.5 dB	23.1 dB	19.0 dB	14.1 dB	7.9 dB
SII	0.74	0.66	0.59	0.51	0.41

25 s were used in the experiment. White noise was added to create noisy signals with five different signal-to-noise (SNR) ratios. Speech intelligibility index (SII) measures were obtained for each processed speech segment. SII provides values between 0.0 and 1.0 that are highly correlated with true intelligibility of speech [11]. The average values of SII and SNR are listed in Table 2. The results show that the noise reduction scheme provides a significant SNR boost with a slight increase of intelligibility index.

2.4.3. *Hearing loss compensation*

The hearing loss threshold is frequency depended and determined at specified frequencies (250, 500, 1000, 2000, 3000, 4000 and 6000 Hz) using a pure tone audiogram. The shape of the correspondent insertion gain depends on prescription method and does not necessarily follow the shape of the hearing loss thresholds. Figure 8 shows how a given insertion gain matches with the proposed SGSM (the insertion gain is obtained with the NAL-RP prescription [11]).

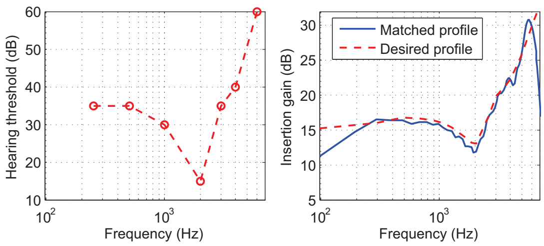

Fig. 8. Hearing loss profile and insertion gain matched with the proposed SGSM

2.4.4. *Dynamic range compression*

The basic idea of DRC is to automatically control the gain in each subband based on the current signal level. The main DRC parameters are the input/output (I/O) function and attack and release times. The I/O function maps the input signal level to the output.

The SGSM block uses the DRC algorithm outlined in table 3. The target spectral gains g_k are determined from the current signal level $P(n,k)$, noise reduction weighting factor $H^{\mathrm{JND}}(n,k)$ and hearing aid insertion gain $D(k)$.

Table 3. Spectral gain shaping method

Sample processing iterations (for $n = 1, 2 \ldots$)
  for $k = 0, \ldots, K-1$ **do**
    if $P(k, n-1) < e_k^2(n)$ **then**
      $P(k,n) = \alpha P(k,n-1) + (1-\alpha)e_k^2(n);$
    else
      $P(k,n) = \beta P(k,n-1) + (1-\beta)e_k^2(n);$
    end if
  end for

Block processing iterations ($n = R, 2R \ldots$)
  for $k = 0, \ldots K-1$ **do**
    $P_{out}(k) = \mathrm{IOFunc}(10\log_{10}(P(k,n)H^{\mathrm{JND}}(k,n)D(k)));$
    $G(k) = P_{out}(k) - P(k,n);$
    $g_k = 10^{G(k)/10};$
  end for

- IOFunc – input/output function of DRC
- $D(k)$ – insertion gains of k-th channel
- α, β – smoothing constants of attack and release times

3. COMPUTATIONAL COMPLEXITY

The prototype of the HA hardware is shown in figure 9. The prototype combines a host processor and reconfigurable part. The host processor performs the following tasks: adaptive filters' coefficients update, noise reduction, DRC and hearing loss compensation. The reconfigurable part includes the filter banks (WCMFB and cochlear FB) and can be adjusted to match target requirements of the HA application (e.g. desired sampling frequency and number of channels).

Table 4 shows the algorithmic complexity of the proposed HA system in terms of real multiplications per sample. WCMFB implies the fast DCT-4 algorithm ($O(\frac{M}{2}\log_2(M) + M)$), polyphase implementation of the prototype filter ($O(N)$)

Table 4. HA computational complexity: M – number of WCMFB channels, N – order of WCMFB prototype filter, K – number of cochlear FB channels

Algorithm	Complexity
WCMFB	$O(\frac{M}{2}\log_2(M) + M + 2N)$
Cochlear FB	$O(2rK)$
Subband AFC	$O(\mathcal{O} \cdot (L + 2L))$
NR + DRC	$O(3K + \frac{11}{R}K)$

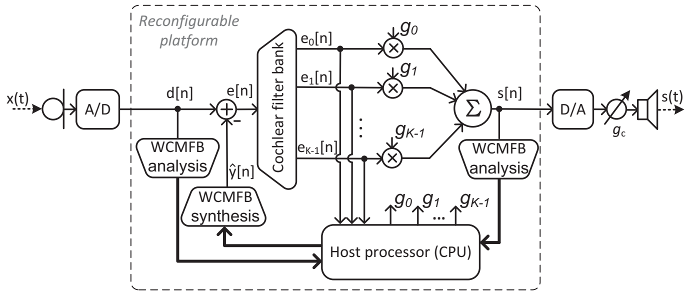

Fig. 9. Prototype of hearing aid system

and all-pass transformation ($O(N)$). As has been stated in section 2.2 in the present HA each cochlear filter is a cascade of two second order sections. Each section requires two real multiplications [6]. In general case if r sections are used the complexity of the cochlear FB is $O(2rK)$.

Considering chosen subsampling ratios the subband AFC system requires $O(\mathcal{O}(L+2L))$ multiplications to implement adaptive filtering, where L is the maximum length of acoustic feedback path in samples and $\mathcal{O}$ is the oversampling factor of WCMFB. PSD estimation in NR and DRC algorithms requires $O(3K)$ multiplications. Recalculation of spectral gains requires approximately $11K$ multiplications once per block of R samples. Hence, total complexity for NR and DRC algorithms equals to $O(3K + \frac{11}{R}K)$. Depending on parameters the overall HA complexity grows linearly or loglinearly.

4. CONCLUSION

A low-delay hearing aid system has been presented. The main component of the system is the spectral gain shaping method that utilizes a cochlear filter bank. The gain shaping method includes noise reduction, hearing loss compensation and dynamic range compression. Acoustic feedback cancellation is carried out in nonuniform frequency subbands. Subband decomposition of the signal is performed by the warped cosine-modulated filter bank. Due to the specific AFC architecture the filter bank does not introduce an additional forward path delay. It has been shown that the proposed HA system has a low computational complexity.

5. REFERENCES

[1] R. W. Bäuml and W. Sörgel, "Uniform polyphase filter banks for use in hearing aids: design and constrainsts," in *Proc. of EUSIPCO'08*, Lausanne, Switzerland, Aug. 2008.

[2] J. M. Kates and K. H. Arehart, "Multichannel dynamic-range compression using digital frequency warping," *EURASIP J. Adv. Sig. Proc*, vol. 2005, no. 18, pp. 3003–3014, 2005.

[3] T. Kurbiel, H. G. Göckler, and D. Alfsmann, "Oversampling complex-modulated digital filter bank pairs suitable for extensive subband signal amplification," in *Proc. of EUSIPCO'09*, Glasgow, Scotland, Aug. 2009, pp. 2658–2662.

[4] A. Pandey and V. J. Mathews, "Low-delay signal processing for digital hearing aids," *IEEE Transactions on Audio, Speech, and Language processing*, vol. 19, no. 4, pp. 699–710, May 2011.

[5] W. Wan, A. Petrovsky, and C. Fan, "A two-dimensional nonlinear cochlear model for speech processing: response to pure tone," in *Proc., 6th International Fase – Congress*, Zurich, Switzerland, 1992, pp. 233–236.

[6] A. Petrovsky, "The synthesis of high order digital bandpass filters with tunable center frequency and bandwidth," in *Proc. of EUSIPCO'96*, Trieste, Italy, Sept. 1996, pp. 1527–1530.

[7] M. Vashkevich, A. Petrovsky, and W. Wan, "Practical design of multi-channel oversampled warped cosine-modulated filter banks," in *Proc. of CCWMC'11*, Shanghai, China, Nov. 2011, pp. 44–49.

[8] M. R. Petraglia and P. B. Batalheiro, "Nonuniform subband adaptive filtering with critical sampling," *IEEE Transactions on Signal Processing*, vol. 56, no. 2, pp. 565–575, 2008.

[9] R. D. Koilpillai and P. P. Vaidyanathan, "Cosine-modulated FIR filter banks satisfying perfect reconstruction," *IEEE Transactions on Signal Processing*, vol. 40, no. 4, pp. 770–783, 1992.

[10] A. Petrovsky, M. Parfieniuk, and A. Borowicz, "Warped DFT based perceptual noise reduction system," in *AES Convention 116*, Berlin, Germany, May 2004, pp. 1–16.

[11] H. Dillon, *Hearing Aids*, Boomerang Press, Australia, 2001.