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A New Delayless Sub-band Filtering Method for Cancelling the Effect of Feedback Path in Hearing Aid Systems

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Abstract— **Performance of commonly used Hearing aid systems is degraded by the presence of acoustic feedback between loudspeaker and microphone. Prediction Error Method Adaptive Feedback Canceller (PEM-AFC) has been proposed recently that could attenuate the feedback effect. In this paper, we present a new delayless frequency-based sub-band filtering method for alleviating the effect of feedback path for the Hearing aid systems. The proposed method avoids sub-band distortions and has low computational complexity making it suitable for low-power cost-effective hearing aid system designs. Performances of the two methods are compared and simulation results are presented.**

I. INTRODUCTION

Currently small size hearing aids suffer from acoustic feedback path (noise) effects due to the short distance between their loudspeaker and microphone. Presence of this feedback hinders the use of desired high gain in the forward path (Fig. 1). On the other hand, the feedback path effect and gain higher than certain threshold level of the hearing aid device produce annoying howling and whistling in the users ears.

Adaptive feedback cancelation techniques are among most commonly used techniques for attenuating such interfering noise signals. In this category of techniques used for hearing aid devices, the continuous AFC, as shown in Fig. 1, is one of the simplest methods. However, performance of this method is degraded considerably when the desired signal is speech or any spectrally colored signal. Such drawback is mainly due to the fact that the signal through the feedback path and the desired signal are correlated [1].

To reduce the mentioned correlation, various methods have been proposed. Among them, the PEM-AFC method has superior performance. In PEM-AFC method, correlation reduction is achieved by whitening technique. In other words, the signal is adaptively whitened at first, then model of the feedback path is estimated by another adaptive filter. The former procedure can be done by using Levinson-Durbin algorithms [1] or other filtering techniques. While the latter can be performed by Filtered-X LMS [2], Filtered-X Recursive Least Square [2], Partitioned Block Frequency Domain Normalized Least Mean Square (PBFD- NLMS) [1], Discrete Fourier Transform MultiDelay block Frequency domain NLMS (DFT-MDF-NLMS) [3], etc. However, the frequency domain implementation and block processing used in PBFD-LMS and DFT-MDF-NLMS seem to be the most desirable approaches in terms of performance.

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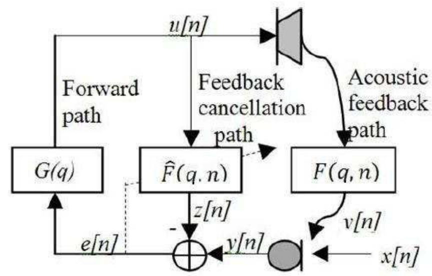

Fig. 1. Continuous AFC Algorithm Including Feedback Path.

In this paper, we present a feedback cancellation method using a new delayless sub-band filtering technique implemented in frequency domain. The proposed method offers unique properties. It avoids data path delays and typical distortion caused by sub-band filtering, especially when the number of sub-bands increases. It also has low computational complexity making the method suitable for low-power and low-cost system designs. We proceed as follows. Section II contains notations used in this paper and references. PEM-AFC method using Levinson-Durbin algorithm and DFT-MDF-NLMS as its adaptive filtering techniques is reviewed in section III. Section IV briefly describes the DFT-MDF-NLMS algorithm. The proposed sub-band filtering is presented in Section V. Section VI briefly compares computational complexity of the two algorithms. Section VII shows the results obtained by simulation and section VIII is the conclusion.

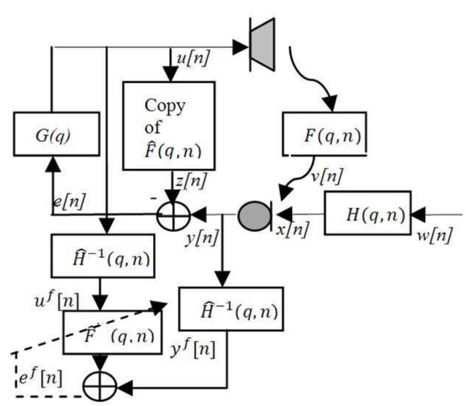

Fig. 2. PEM-AFC Algorithm

TABLE I

DFT-MDF-NLMS PROCEDURE

For each M-sample block of input u^f
Filter length $N = P \times M$ and 50% overlap:

NFFT=2M, Number of sub-filters=2M, Length of each sub-filter=P

$$\mathbf{u}^f_{M,k} = [u^f(kM+M-1) \dots \; u^f(kM)]_{1\times M}{}^T$$

$$\mathbf{u}'_{2M,k} = \mathbf{F}\begin{pmatrix} \mathbf{u}^f_{M,k} \\ \mathbf{u}^f_{M,k-1} \end{pmatrix}_{2M\times 1} \equiv [u'_0[k] \;\; \dots \;\; u'_{2M-1}[k]]^T$$

$$\mathbf{u}'_{j,k} = [u'_j[k] \;\; \dots \;\; u'_j[k-P+1]]_{1\times P}$$

where j denotes frequency bin, j=0:2M-1

$$y'_j[k] = \mathbf{u}'_{j,k}\mathbf{l}^c_{j,k-1} \qquad j = 0 : 2M-1$$

where $\mathbf{l}^c_{j,k}$ is a $P \times 1$ vector of constraint coefficients for sub-filter j

$$\mathbf{e}'_{2M,k} = \mathbf{F}\begin{pmatrix} \mathbf{I}_M \\ \mathbf{0} \end{pmatrix}_{2M\times M}(\mathbf{y}^f_{M,k} - [\mathbf{I}_M \;\; \mathbf{0}]\mathbf{F}^*[y'_0[k] \;\; \dots \;\; y'_{2M-1}[k]]^T)_{M\times 1}$$

$$\equiv [e_0{}'[k] \;\; \dots \;\; e'_{2M-1}[k]]_{1\times 2M}{}^T$$

$$\mathbf{l}_{j,k} = \mathbf{l}_{j,k-1} + \mu[k]\mathbf{u}'^*_{j,k}\mathbf{e}'_j[k] \quad j = 0 : 2M-1$$

where $\mathbf{l}_{j,k}$ is a $P \times 1$ vector of non-constraint coefficients for sub-filter j

$$\mu_j[k] = \frac{\mu_0}{\varepsilon + ||\mathbf{u}'_{j,k}||^2} \qquad \varepsilon and \mu_0 are constant.$$

$$\begin{pmatrix} \mathbf{l}^{cT}_{0,k} \\ \mathbf{l}^{cT}_{1,k} \\ \cdot \\ \cdot \\ \cdot \\ \mathbf{l}^{cT}_{2M-1,k} \end{pmatrix}_{2M\times P} = \frac{1}{2M}\mathbf{F}^*\begin{pmatrix} \mathbf{I}_M & \mathbf{0} \\ \mathbf{0} & \mathbf{0} \end{pmatrix}_{2M\times 2M}\mathbf{F}\begin{pmatrix} \mathbf{l}^T_{0,k} \\ \mathbf{l}^T_{1,k} \\ \cdot \\ \cdot \\ \cdot \\ \mathbf{l}^T_{2M-1,k} \end{pmatrix}_{2M\times P}$$

*F is FFT matrix and * denotes conjugate of matrix or vector*

II. NOTATIONS

The symbols n and q^{-1} denote discrete-time index and discrete-time delay, respectively [1], i.e.

$$q^{-1}u[n] = u[n-1] \tag{1}$$

A discrete-time filter of length L_F can be represented by:

$$F(q,n) = \mathbf{f}^T[n]\mathbf{q} \tag{2}$$

Where $\mathbf{f}[n] = [f_0[n] \;\; f_1[n] \;\; \dots \;\; f_{L_F-1}[n]]^T$ is a vector of filter coefficients and $\mathbf{q} = [1 \;\; q^{-1} \;\; \dots \;\; q^{-L_F+1}]^T$. Based on this notation, filtering of $u[n]$ by $F(q,n)$ is denoted by:

$$F(q,n)u[n] = \mathbf{f}^T[n]\mathbf{u}[n] \tag{3}$$

where $\mathbf{u}[n] = [u[n] \;\; u[n-1] \;\; \dots \;\; u[n-L_F+1]]^T$.

TABLE II

COMPUTATIONAL COMPLEXITY (B: BLOCK SIZE, N: FILTER LENGTH)

Method	Real Multiplication per Sample
DFT-MDF-NLMS	$4(\frac{2N}{B} + (\frac{N}{B} + 3(log_2 2B)))$
Proposed Sub-band	$4log_2(2B) + (1+\frac{1}{b})\frac{4N}{B}(1+2log_2) + \frac{4N}{B}log_2\frac{4N}{B}$

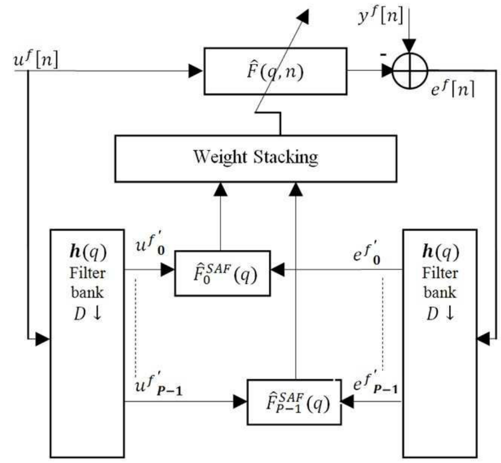

Fig. 3. The Sub-band Filtering Method

III. PEM-AFC

Presence of the closed loop in Fig. 1, makes feedback signal and desired signal be correlated with each other when the desired signal is a spectrally colored signal, like speech. Common AFC algorithm does not exploit any method to decrease the aforementioned signals correlation. As a consequence, this algorithm results in a biased estimation of the compensating filter coefficients, inadequate attenuation of the feedback path effect and low system performance. PEM-AFC is one of the most robust methods for decreasing this amount of bias [1]; in which correlation between desired signal and feedback signal is reduced by means of whitening filter. Fig. 2 depicts the block diagram corresponding to this method. In hearing aid the desired signal is generally a speech signal which can be (piece-wise) modeled by an AR process, i.e.:

$$x[n] = H(q,n)w[n] \tag{4}$$

Where $H(q,n)$ is an AR model and $w[n]$ is an impulse train or a zero-mean white noise sequence for voiced or unvoiced phonemes, respectively.

Loudspeaker and microphone signals are whitened by $\hat{H}^{-1}(q,n)$ whose transfer function is the inverse of AR model estimated by Levinson-Durbin algorithm. Whitened signals $u^f[n]$ and $y^f[n]$ are considered as the input and desired signals of adaptive filter $\hat{F}(q,n)$.

Applying LMS algorithm to $e^f[n]$ (or orthogonality principle) leads to normal equations below, by which estimated coefficients can be found [1].

$$\hat{f}[n] = E\{u^f[n]u^{f,T}[n]\}^{-1}E\{u^f[n]y^f[n]\} \tag{5}$$

$\hat{f}[n]$ is the vector of estimated coefficients. $u^f[n]$ contains last $L_{\hat{F}}$ samples of $u^f[n]$; where

$$u^f[n] = \hat{H}(q,n)^{-1}u[n] \tag{6}$$

$$y^f[n] = \hat{H}(q,n)^{-1}y[n] \tag{7}$$

Assuming $\hat{H}(q,n) = H(q,n)$, $y^f[n]$ can be replaced by $w[n] + F(q,n)u^f[n]$ in Eq. 5; which leads to an unbiased estimation of feedback coefficients [1].

TABLE III

SIMULATION CONDITIONS

Method	Feedback Path	Forward Path	Desired Signal	AR Length for L-D	Frame Length for L-D	Length of Full-band Filter	Blocksize Blocksize	Evaluation Criterion
DFT-MDF-LMS	100-tap FIR filter	$10e^{1}60$	AR(3)	3	10 ms	128	16,32,64,128	Misalignment
			Speech	20	10 ms	128	64	PESQ
Proposed	100-tap FIR filter	$10e^{1}60$	AR(3)	3	10 ms	128	16,32,64,128	Misalignment
Sub-band			Speech	20	10 ms	128	64	PESQ

IV. DFT-MDF-NLMS

Filter $\hat{F}(q,n)$ can be implemented by different adaptive algorithms. One of them is DFT-MDF-NLMS, which possesses advantages of block processing of LMS algorithm in frequency domain. This method partitions $\hat{F}(q,n)$ into smaller sub-filters based on polyphase definition to process the data block by block in the frequency domain. Corresponding procedure is represented in Table I [4].

V. THE SUB-BAND FILTERING

Fig. 3 illustrates the sub-band technique, in which input and error signals are decomposed into sub-bands by analysis filter bank $h(z)$. Decomposed or sub-band signals are each used in sub-band adaptive filters to find corresponding filter coefficients or sub-band weights, which are then stacked together in order to make full-band filter $\hat{F}(q,n)$.

For each sub-band filter, weight updating can be done by NLMS algorithm. Providing $y(n)$ by a full-band filter instead of having it obtained by combining the outputs of sub-band adaptive filters, makes this structure delayless [5].

Filter bank $h(z)$ contains parallel filters $H_k(z)$ $0 \leq k \leq P-1$, which are related to (frequency-shifted version of) the prototype filter $H_0(z)$ as:

$$H_k(z) = H_0(ze^{-j2\pi k/M}) \qquad 0 \leq k \leq P-1 \tag{8}$$

The prototype filter $H_0(z)$ is defined as [5]:

$$H_0(z) = 1 + z^{-1} + \ldots + z^{-M+1} \tag{9}$$

To decrease the spectral distortion (aliasing) introduced by side-lobes of filters $H_k(z)$ $0 \leq k \leq P-1$, decimation factor is assumed to be $D = P/4$ which indicates an oversampling sub-band filtering situation. Instead of having synthesis filter bank, frequency bins of $\hat{F}(e^{j\omega})$ are directly selected from the frequency bins of sub-band Adaptive filters weights $\hat{F}_k^{SAF}(e^{j\omega})$ based on the following formula [5]:

$$\begin{cases} l \in [0,N], \ \hat{F}[l] = \hat{F}^{SAF}_{((\lfloor \frac{lP}{2N} \rfloor))_P}[((l))_{\frac{8N}{P}}] \\ l \in [N,2N], \hat{F}[l] = \hat{F}[2N-1]^* \end{cases} \tag{10}$$

VI. COMPUTATIONAL COMPLEXITY

A Table II summarizes the computational complexity of both methods in terms of real multiplication [4], [5]. For both algorithms calculation of K-point FFT is assumed to have $2Klog_2^K$ real multiplications. Numerical comparison will be shown in the Experimental Results section.

VII. EXPERIMENTAL RESULTS

In this section, PEM-AFC method (Fig. 2) using Levinson-Durbin algorithm for whitening procedure is implemented. DFT-MDF-NLMS and the sub-band filtering are respectively used instead of adaptive filter $\hat{F}(q,n)$ which is responsible for estimation of the feedback path model. Comparison results can be found at the end of this section. FIR filter of length 100 (Fig. 4) which has been measured in [1], is used as the feedback path model in our simulations. Two types of signals are considered as the desired signal, i.e. AR(3) process given below, and speech signal with length 15 s and sampling frequency of 16KHz.

$$H(q,n) = \frac{1}{(1-0.99e^{-j2\pi 0.2}q^{-1})(1-0.99e^{j2\pi 0.2}q^{-1})} \tag{11}$$

Levinson-Durbin algorithm updates its AR coefficients every 10 ms. Forward path transfer function is [1]:

$$G(q) = Ge^{d_G} \tag{12}$$

Where, G and d_G are constants equal to 10 and 160 (samples), respectively. Table III contains the conditions we have assumed in our simulation. To evaluate the performance of the algorithms, two criteria are used. First, Misalignment defined by (13) when using AR process as the desired signal. Second, the Perceptual Evaluation of Speech Quality (PESQ) when using actual speech signal.

$$Misalignment = 10log_{10}((\frac{\int_0^{\pi} |F(e^{j\omega}) - \hat{F}(e^{j\omega})|^2 d\omega)}{(\int_0^{\pi} |F(e^{j\omega})|^2 d\omega))} \tag{13}$$

Based on the above criteria, lower Misalignment and higher PESQ indicate better performance of the algorithm used.

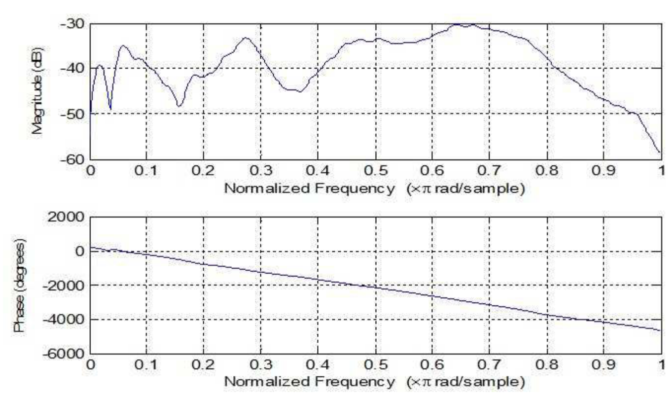

Fig. 4. Feedback Path Transfer Function

Fig.s 5 and 6 show Misalignments of DFT-MDF-NLMS and the proposed sub-band filtering method versus stepsize when the length of full-band filter is 128. In these figures, each curve corresponds to one particular block size. According to Fig. 6, for sub-band method larger block size, which corresponds to larger number of sub-bands, gives better misalignment; while for DFT-MDF-NLMS it is reverse. However, Fig. 5 shows better misalignment for DFT-MDF-NLMS in comparison to the proposed method.

Table IV indicates the results when the input signal is a speech signal. Simulations have shown that our sub-band filtering method has better performance for variable normalized stepsize instead of having common normalized stepsize. In common normalized stepsize $(\mu_0/power(\mathbf{u}^{\mathbf{f}})), \mu_0$ is a constant, while here, the initial μ_0 for the sub-band filtering method is 0.4 and it is divided by 1.3 per 100 iterations. However, this is a formula found by trial and error which may not be the optimum case. As a result, PESQ represented for the proposed method can improve further. According to this table, PESQ of DFT-MDF-NLMS is slightly better than the PESQ of proposed sub-band method. However, computational complexity of the proposed sub-band for filter length of 128 is less than the computational complexity of DFT-MDF-NLMS. [5] proves that for higher order of the filter and consequently for higher number of the sub-bands, proposed sub-band method will have less complexity compared to DFT-MDF-NLMS. As a result for those applications in which higher number of sub-band is required, proposed sub-band would be more suitable than DFT-MDF-NLMS.

VIII. Conclusion

Two frequency domain LMS algorithms have been used in PEM-AFC method in order to reduce negative effect of acoustic feedback. Both algorithms have been tested by AR process and speech signal. The former algorithm, i.e. DFT-MDF-NLMS has better misalignment and PESQ. While the latter adds improved computational complexity to its benefits specially for higher numbers of sub-bands.

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TABLE IV

Results for Speech Signal

Method	Stepsize	PESQ	Computational Complexity
DFT-MDF-NLMS	$\mu_0 = 0.000005$	4.1308	156
Proposed Sub-band	$initial\mu_0 = 0.4$ $\mu_0 = \frac{\mu_0}{1.3}$ *if* *mod(iteration,100)=0*	3.9049	148

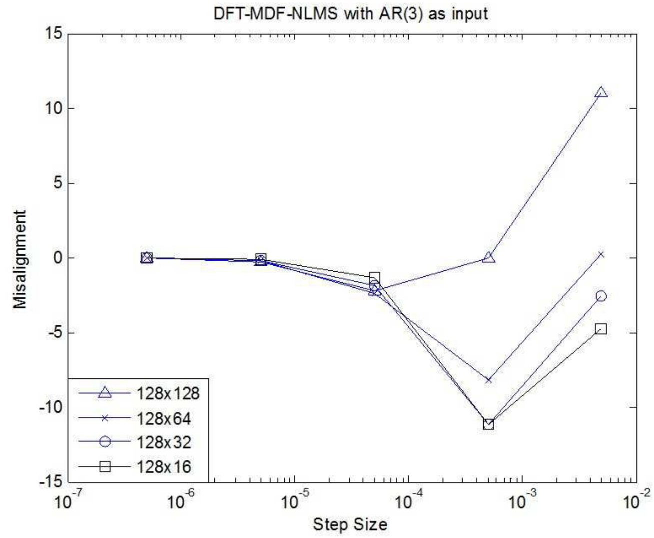

Fig. 5. DFT-MDF-NLMS with 128-tap full-band filter and AR(3) as the desired signal

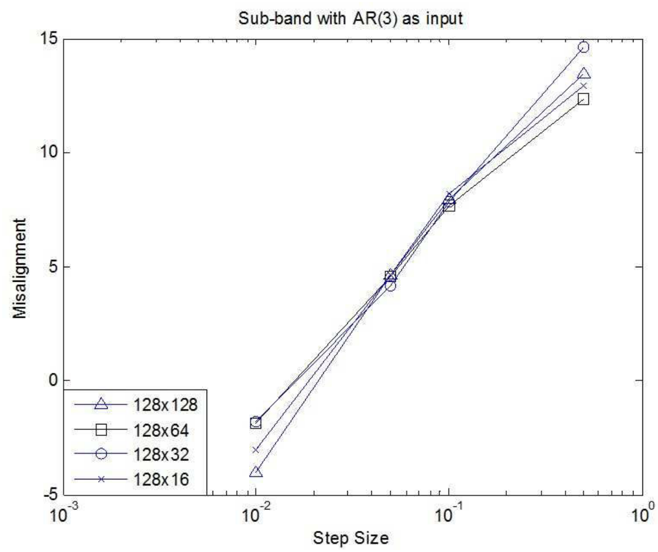

Fig. 6. Sub-band method with 128-tap full-band filter and AR(3) as the desired signal

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