An Over-Sampling Subband Adaptive Filter with the Optimal Real Filter Bank

Yoshito Higa
Digital Video Research Group
Texas Instruments,
Tsukuba RD Center Ltd.
Tsukuba, Japan
E_mail: higa@trdc.ti.com

Hiroshi Ochi
Department of
Information Engineering,
University of the Ryukyus
Okinawa Japan
ochi@ie.u-ryukyu.ac.jp

Shigenori Kinjo
Department of Electrical and
Electronic Engineering,
University of the Ryukyus
Okinawa Japan
kinjo@ie.u-ryukyu.ac.jp

Abstract

The over-sampling subband ADFs are highly demanded because of shortening the convergence of adaptation algorithms and being able to avoid aliasing. However, the conventional methods must use the complex coefficients filter banks, which require a lot of computation compared to the real filter banks. The aim of this correspondence is to develop a new over-sampling subband ADF which does not require complex coefficient filter banks but real filter banks. The filter banks are optimized such that the MSE can be minimized. This turns out that the proposed scheme can minimize the aliasing effect due to decimation.

1 Introduction

In many adaptive filter applications, such as acoustic echo cancellers, an FIR ADF(adaptive digital filters) requires more than several thousand weights. In this case, the use of SADF(subband ADF) is suitable[1] -[5]. This technique decomposes a very long impulse response of the adaptive filter into a short length; so it can achieve parallel processing of adaptive filters. The SADF can be implemented by either the maximal decimation or over-sampling. The maximal decimation is suitable for saving the load of calculation. On the other hand, the over-sampling technique is also highly demanded for shortening the convergence of adaptation algorithms .

The conventional subband algorithms are not sufficient to implement SADFs. This is mainly because they cannot perfectly avoid the degradation effect of aliasing due to decimation. Gilloire et al. have improved the effect of aliasing by allowing additional adaptive cross terms between the subbands[4]. Somayazulu et al. have proposed an over-sampling adaptive structure with auxiliary subbands [1]. However, Gilloire's method increases the computational complexity and Somoyazulu's method cannot carry out the maximum decimation. With regards to the over-sampling SADFs, the conventional methods must use complex coefficients filter banks, which require a lot of computation compared to the real filter banks, due to the complex frequency shift of the prototype low-pass filter in order to avoid aliasing.

We have also proposed another type of SADF in which the analysis filter bank has been optimized by minimizing the LMSE(least mean squared error)[6] - [7]. The MSE in terms of this technique has drastically reduced compared with the conventional SADF.

However, the methods in[6]-[7] have not dealt with the over-sampling case. Therefore, the aim of this correspondence is to develop a new adaptation algorithm which does not require complex coefficient filter banks but real filter banks. The filter banks are optimized such that the MSE can be minimized. This turns out that the proposed scheme can minimize the aliasing effect due to decimation. Firstly, the design methodology for real coefficient filter banks for the over-sampling SADF is investigated in Section 2. Section 3 develops a new adaptation algorithm based on the real coefficient SADF. Finally, some computer simulations are examined in order to show the effectiveness of the proposed method in Section 4.

2 Design of Over-Sampling Filter Bank

First, we design a prototype low-pass filter $h_p(n)$, whose cut-off frequency is $\frac{\pi}{2N}$, based on the N-channel pseudo-QMF filter banks[8]. Then the *ith* bin's impulse response $h_p(n)$ of the real coefficient filter bank can be obtained by the cosine modulation according to the following manner:

$$\begin{aligned} h_i(n) &= h_p(n) \\ &\cdot \cos\left(\frac{\pi}{2N}(2i+1)\left(n-\frac{L-1}{2}\right)+\theta_{(i)}\right) \\ i &= 0 \sim N-1, n = 0 \sim L-1 \end{aligned} \quad (1)$$

where $\theta_{(0)} = \frac{\pi}{4}, |\theta_{(i)} - \theta_{(i+1)}| = \frac{\pi}{2}$.

Figure 1 shows an N-channel filter bank in which $N = 4$. The decimation factor is $D = 4$. In the case of over-sampling, i.e. $N \neq D$, the alias-free condition can be written as

$$N = \tau D \quad (2)$$

where τ is the integer number.

Figure 2 illustrates the spectra of the synthesis filter output signals in terms of distinct decimation factor. The shaded parts of the spectra are degraded by the aliasing.

When $D = 4$, the filter bank is the maximally decimated filer bank. The signal spectra in every bin are degraded due to the aliasing; however, they are all canceled by the synthesis filter bank.

Next we consider the case of $D = 2$, which satisfies the condition in Eq.(2). Although the bins 0 and 3 are free from aliasing, the spectra in bins 1 and 2 are degraded by the aliasing. Since they cancel out each other as shown in Fig.2 (c), the synthesized output is alias free.

Fig.2 (b) shows the case which does not satisfy the condition in Eq.(2). Since the alias component at $\frac{\pi}{3}, \frac{2\pi}{3}$ are not canceled, we cannot obtain the perfect recontructed output.

From these reasons, we must satisfy the condition in Eq.(2) to obtain the perfect reconstructed signal $\hat{x}(n)$ in case of using the over-sampling filter bank.

3 Proposed Over-Sampling Subband ADF

3.1 Architecture

As we mentioned in section 2, some bins of the real filter banks are degraded by the aliasing even if we use the over-sampling filter banks. This results in the increase of the mean square error of the over-sampling SADF. We shall propose a new method which avoids this difficulty.

Figure 3 shows a proposed over-sampling SADF. It has two kinds of analysis banks. The analysis filter bank $G_k(z)$, which splits reference signals into several bins, consists of a pair of PR(Perfect Reconstruction) filter banks with the synthesis filter bank $K_k(z)$.

It should be noted that we have freedom of design in the analysis filter bank $F_k(z)$ with regard to input signals. This is the advantage of the proposed SADF. If $F_k(z)$ is designed so that $y_k(m) \approx d_k(m)$ after adaptation of $H_k(z)$, $y(n)$ approximately becomes $d(n)$ by use of the PR bank. The proposed scheme adaptively designs the $F_k(z)$.

3.2 Cost Function

First, we define some variables. The $x_k(m)$ and $d_k(m)$ are an input and reference signal with regard to kth bin's ADF. The $h_k(m)$ and $f_k(m)$ are the coefficient of the ADF at the kth bin and that of analysis filter bank with respect to input signals, respectively. L_{h_k} and L_{f_k} denote the length of the $h_k(m)$ and $f_k(m)$, respectively, in which $L = L_{f_k} + (L_{h_k} - 1) \times D$. $\mathbf{R}_{xx}$ denotes the auto-correlation matrix in terms of $x(n)$. The cross-correlation vector $\mathbf{P}_{d'_k x}(\tau)$ is defined with the band-limited reference signals $d'_k(n)$ and input signals $x(n)$.

$$\mathbf{h}_k^T = [h_k(0) \cdots h_k(L_{h_k} - 1)] \quad (3)$$

$$\mathbf{f}_k^T = [\ f_k(0) \cdots f_k(L_{h_k} - 1)\ ] \quad (4)$$

$$\mathbf{0} = [\ 0 \cdots\ 0\ ]^T\ (D \times 1) \quad (5)$$

$$\mathbf{F}_k = \begin{bmatrix} \mathbf{f}_k & \mathbf{0} & \cdots & \mathbf{0} \\ \mathbf{0} & \mathbf{f}_k & \cdots & \mathbf{0} \\ \vdots & \vdots & \ddots & \vdots \\ \mathbf{0} & \mathbf{0} & \cdots & \mathbf{f}_k \end{bmatrix} \quad (6)$$

$$\mathbf{R}_{xx} = \begin{bmatrix} r_{xx}(0) & \cdots & r_{xx}(L-1) \\ r_{xx}(1) & \cdots & r_{xx}(L-2) \\ \vdots & \ddots & \vdots \\ r_{xx}(L-1) & \cdots & r_{xx}(0) \end{bmatrix} \quad (7)$$

$$\mathbf{p}_{d'_k x}^T(\tau) = [\ p_{d'_k x}(\tau) \cdots p_{d'_k x}(\tau + L_{f_k} - 1)\ ] \quad (8)$$

$$\mathbf{P}_{d'_k x} = [\ \mathbf{p}_{d'_k x}(0) \cdots \mathbf{p}_{d'_k x}((L_{h_k} - 1)D)\ ] \quad (9)$$

Now we define the kth bin's cost function as follows:

$$MSE_k = E[e_k^2(m)] = E\left[\left\{d_k(m) - \mathbf{h}_k^T \mathbf{x}_k(m)\right\}^2\right] \quad (10)$$

By substituting equations (3) ~ (9) into (10), we can rewrite MSE_k as follows:

$$MSE_k = \sigma_{d_k}^2 - 2\mathbf{f}_k^T \mathbf{P}_{d'_k x} \mathbf{h}_k + \mathbf{h}_k^T \mathbf{F}_k^T \mathbf{R}_{xx} \mathbf{F}_k \mathbf{h}_k \quad (11)$$

From equation (11), the optimal adaptive coefficients and the $LMSE$ can be obtained as follows:

$$\mathbf{h}_{k,opt} = \left\{\mathbf{F}_k^T \mathbf{R}_{xx} \mathbf{F}_k\right\}^{-1} \mathbf{P}_{d'_k x}^T \mathbf{f}_k. \quad (12)$$

$$LMSE_k = \sigma_{d_k}^2 - \mathbf{f}_k^T \mathbf{P}_{d'_k x} \left\{\mathbf{F}_k^T \mathbf{R}_{xx} \mathbf{F}_k\right\}^{-1} \mathbf{P}_{d'_k x}^T \mathbf{f}_k. \quad (13)$$

Equation (13) is the function in terms of $F_k(z)$; therefore, we can design the optimal filter coefficient for the analysis filter bank $F_k(z)$ by minimizing (13). In real applications, $LMSE_k$ is minimized by an adaptive fashion such as the LMS algorithm.

4 Simulation Result

We show a computer simulation under the following conditions.

Unknown system	:	64-tap FIR LPF
Input signal $x(n)$	:	1st-order AR
	:	($pole = 0.9$)
Algorithm	:	NLMS
Step size	:	1.0
Filter bank 1	:	24-tap (4-band)
Filter bank 2 (QMF)	:	24-tap (4-band)
Decimation factor	:	$D = 2$

Figure 4 shows the amplitude responses of the filter bank 2, $G_k(z)$. The optimized filter bank 1, $F_k(z)$, whose amplitude responses is shown in Figure 5, can be obtain by minimizing the cost function in Eq.(13).

Figure 6 shows learning curves of the proposed scheme and that of the conventional over-sampling SADF with real coefficients filter banks. As we can see, the noise floor is drastically improved by using the proposed method.

5 Conclusion

We have proposed a new type of over-sampling subband adaptive filter, which can reduce the LMSE compared with that of the conventional schemes, by using two different types of analysis filter banks.

Our future study is to analyze the convergence property of the proposed algorithm such as the convergence rate and how to define the step size.

References

[1] V.S.Somayazulu, S.K.Mitra, and J.J.Shynk, "Adaptive line enhancement using multirate techniques," Proc. ICASSP, Glasgow, pp.928-931, May 1989.

[2] M.R.Asharif, F.Amano, S.Unagami and K.Murano, "Acoustic echo canceller based on frequency bin adaptive filter(FAF)," Proc. IEEE Int. Conf. Global Telecommun., Tokyo, pp.1940-1944, Nov. 1987.

[3] W.Kellermann, "Analysis and design of multirate systems for cancellation of acoustical echoes," Proc. ICASSP, NewYork, pp.2570-2573 Apr. 1988.

[4] A.Gilloire, "Experiments with subband acoustic echo cancellers for teleconferencing," Proc. ICASSP, Dallas, pp.2141-2144, Apr. 1987.

[5] A.Gilloire and M.Vetterli, " Adaptive filtering in subbands," Proc. ICASSP, NewYork, pp.1572-1575, Apr. 1988.

[6] H.Ochi, Y.Higa and S.Kinjo, "A Subband Adaptive Filter with Two Types of Analysis Filter Bank," Proc.ICASSP95, pp.993-996, Detroit, May 1995

[7] Y.Higa, H.Ochi and S.Kinjo, "A Subband Adaptive Filter with a Variable Analysis Filter Bank," Proc.ICASSP96, pp.1727-1730, Atlanta, May 1996

[8] Trong Q.Nguyen, A.K.Soman and P.P.Vaidyanathan,"A Quadratic-constrained least-squares approach to linear phase orthonormal filter bank design", Proc.ICASSP93,pp.383-386, 1993

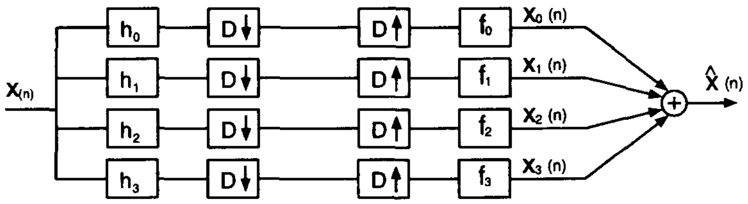


Fig.1. 4 channel filter bank.

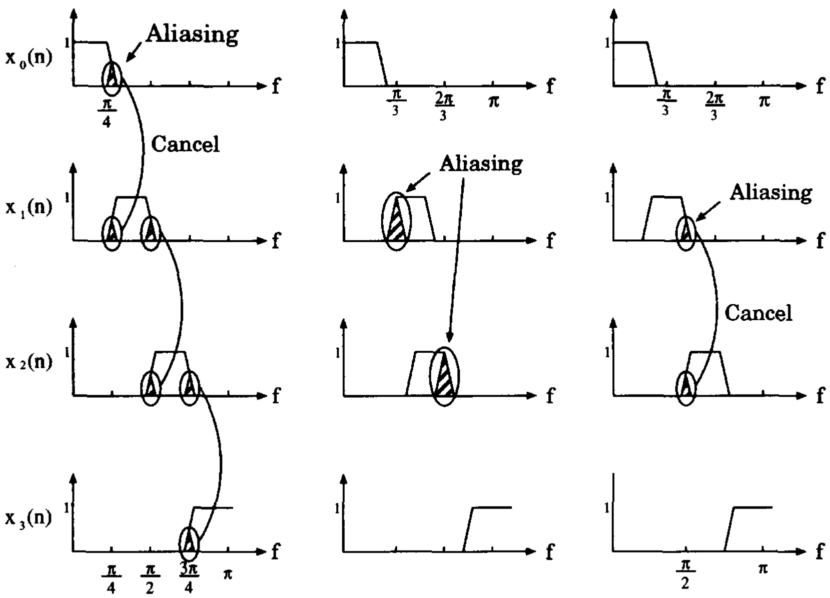


Fig.2. Aliasing canceling on each decimation factor .

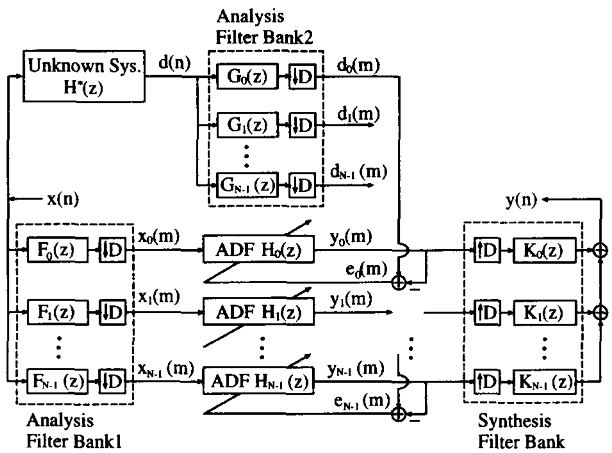


Fig.3. Proposed subband ADF.

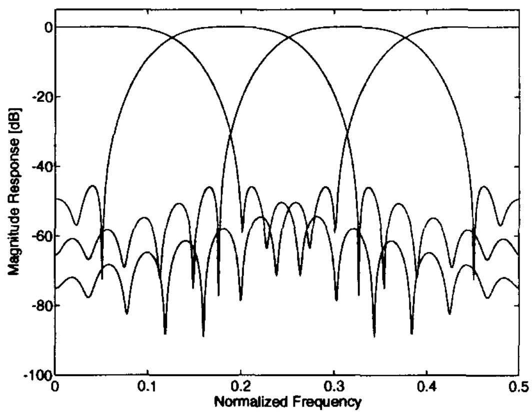


Fig.4. QMF bank($G_k(z)$).

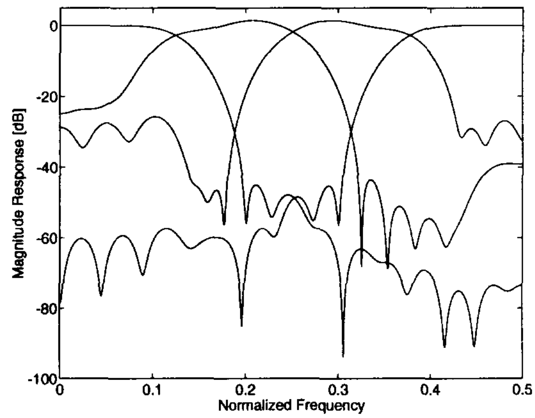
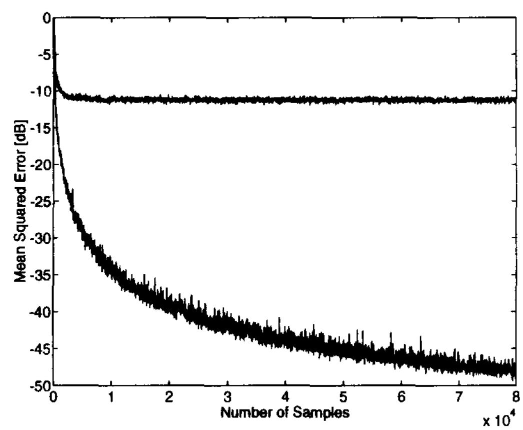


Fig.5. Optimized filter bank($F_k(z)$).



The upper curve : Conventional method
The lower curve : Proposed method

Fig.6. Learning curves.