EFFICIENT DESIGN OF LOW DELAY IIR QMF BANKS FOR SPEECH SUBBAND CODING

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ABSTRACT

Speech subband coding offers resources for wideband speech processing due to the utilization of masking effects of the human ear. The effectiveness of this method depends on a proper splitting of the signal frequency band. In this paper we propose an efficient design of low delay QMF banks using IIR prototype filters. These filterbanks allow sharp bandsplitting operations with a sufficient amount of subband channels and distinctively low perceptive phase distortions.

1 INTRODUCTION

Subband coding techniques take advantages of specific properties of the human ear, like critical bandwith and temporal masking. Useful filterbanks for splitting the signal frequency band of up to 7 kHz should establish 10 or more channels similar to the distribution of the critical bands.

Conventional methods are using FIR QMF banks to take advantage of the linear phase characteristic of nonrecursive filters and the possibility of perfect recontruction (PR) [1]. But sharp cutoff slopes of subband channels require high order FIR filters associated with high system delay of the entire filterbank.

This paper presents QMF banks in tree structure using IIR filters of distinctively low order and sharp cutoff slopes which allow different bandsplitting characteristics with up to 24 channels in regular and nonregular structures. For implementation a single AT&T DSP32C realtime processing environment is used. Variable filter orders inside the tree structure and various compensative methods are evaluated on the effect of minimizing perceivable phase distortions.

2 TWO CHANNEL FILTERBANKS IN POLYPHASE STRUCTURE (REVIEW)

The two channel filter bank (Fig. 1) is characterized by a low and high pass filtering process in the analysis bank

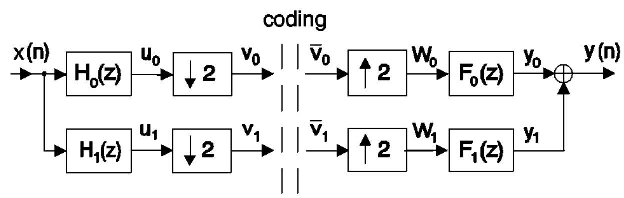


Figure 1: Two-Channel Filterbank

followed by a decimation with factor 2. The synthesis bank recombines the subband signals after interpolation.

By proper choice of the synthesis filters $F_0(z)$ and $F_1(z)$ aliasing due to decimation is perfectly cancelled. The constraints are as follows:

$$F_0(z) = H_1(-z) \quad (1)$$
$$F_1(z) = -H_0(-z) \; . \quad (2)$$

This leads to the transfer function $T(z)$ of the entire filterbank:

$$T(z) = \frac{Y(z)}{X(z)} = \frac{1}{2}\left[H_0(z)H_1(-z) - H_1(z)H_0(-z)\right] \; . \quad (3)$$

If we assume $H_0(z)$ and $H_1(z)$ to be quadrature mirror filters ($H_1(z) = H_0(-z)$) this equation is simplified to

$$T(z) = \frac{1}{2}\left[H_0(z)^2 - H_0(-z)^2\right] \; . \quad (4)$$

The above mentioned assumption can be used in recursive and nonrecursive filter structures and ensures stable transfer functions in the analysis and synthesis bank.

In order to give more effectiveness to the signal processing decimation and interpolation should be conducted before analysis filtering and after synthesis filtering, respectively. This can be achieved by the polyphase decomposition of $H_0(z)$:

$$H_0(z) = E_0(z^2) + z^{-1}E_1(z^2) \; . \quad (5)$$

The decomposition can be applied to all nonrecursive and a special class of elliptic recursive filters.

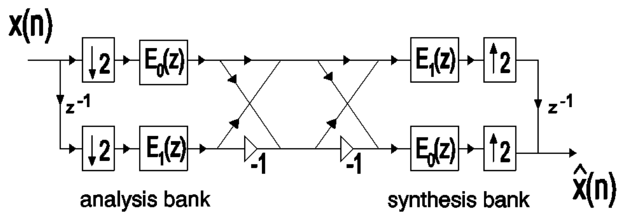


Figure 2: Two Channel QMF bank in polyphase structure

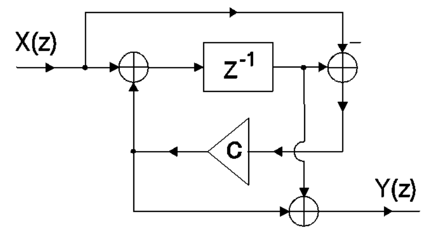


Figure 3: Allpass structure type 1B according to Mitra/Hirano

"Noble Identities" [2] allow the shifting of decimation and interpolation and lead to the two channel QMF bank structure depicted in Fig. 2.

The transfer function $T(z)$ of is given by:

$$T(z) = \frac{\hat{X}(z)}{X(z)} = 2z^{-1}E_0(z^2)E_1(z^2) \ . \tag{6}$$

Conventional QMF bank structures employ nonrecursive (FIR) prototype filters $H_0(z)$ with an optimized polyphase design to ensure linear phase and nearly allpass characteristics of $T(z)$ [1].

Lossless recursive (IIR) filters with Butterworth, Chebyshev and Cauer design principles can always be described as a sum of two allpass transfer functions [2]. If we assume the allpass functions to be of the form $A(z) = a(z^2)$, polyphase decomposition of IIR filters is possible and leads to the equations:

$$H_0(z) = \frac{1}{2}\left[a_0(z^2) + z^{-1}a_1(z^2)\right] \tag{7}$$

$$T(z) = \frac{1}{2}z^{-1}a_0(z^2)a_1(z^2) \ . \tag{8}$$

3 DESIGN PROCEDURE

Lossless elliptic IIR filters with real coefficients of the allpass components can be decomposed in polyphase structure if their transfer function is derived by:

$$H_0(z) = \frac{P(z)}{\prod_k (1 + d_k z^{-2})} \tag{9}$$

($P(z)$ is a polynomial of odd order) [2].

With the QMF relation $H_1(z) = H_0(-z)$ and (??) we can derive:

$$|H_0(e^{j\psi})|^2 + |H_0(e^{j(\pi-\psi)}|^2 = 1 \ , \tag{10}$$

which means that $|H_0(z)|^2$ is a halfband filter and all the poles of $H_0(z)$ are imaginary.

Now the design can be easily conducted by a conventional Cauer filter design under the constraints:

$$a_s = -10\log\left(1 - 10^{-\frac{a_d}{10}}\right) \tag{11}$$

$$\psi_s = \pi - \psi_d \tag{12}$$

(a_s/a_d stopband/passband attentuation; ψ_s/ψ_d stopband/passband edge frequency).

The allpass transfer functions $a_0(z)$ and $a_1(z)$ can be found with [3]:

$$a_0(z) = \prod_{k \ odd}\left(\frac{d_k + z^{-1}}{1 + d_k z^{-1}}\right) \tag{13}$$

$$a_1(z) = \prod_{k \ even}\left(\frac{d_k + z^{-1}}{1 + d_k z^{-1}}\right) \ . \tag{14}$$

37 prototype filters with stopband edge frequencies $0,5\pi \leq \psi_s \leq 0,68\pi$ and stopband attentuations 40 dB $\leq a_s \leq$ 58 dB have been designed (Filter orders 5, 7, and 9).

The resulting filter structures are realized using experimental results on allpass networks by S.K.Mitra and K.Hirano [4]. With allpass coefficients d_k depicted in (13) and (14) best results on multiplication round-off errors and a minimum number of multipliers can be achieved with allpass structure type 1B (Fig. 3, $c = -d_k$) [4].

4 IMPLEMENTATION OF MULTICHANNEL IIR QMF BANKS

Multichannel IIR QMF banks based on the two channel IIR QMF bank have been realized on a real time DSP32C environment establishing a variety of regular, dyadic and hybrid tree structures.

In order to reduce perceivable phase distortions especially in the high level analysis filters, IIR prototype filters with wider transition bands are implemented in higher tree levels. This method is similar to conventional FIR QMF tree structures. Experiments using arbitrary male and female speech signals showed distinctive improvements.

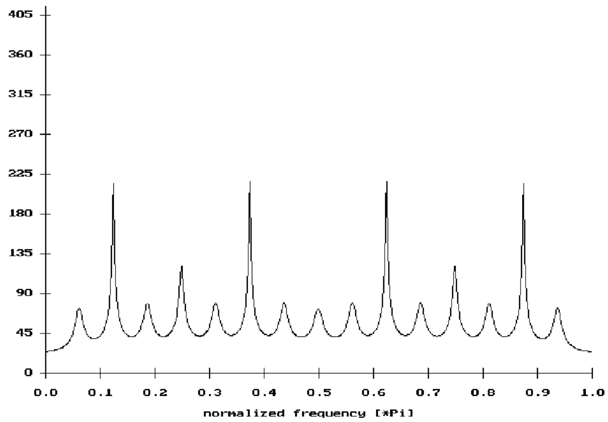


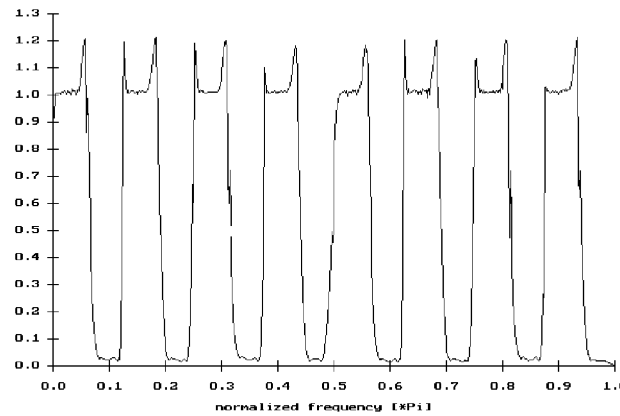
Figure 4: Group delay of a regular 16 channel IIR QMF bank (Prototype filters: $\psi_s = 0,52\pi$ (order 9) in level 1-3, $\psi_s = 0,65\pi$ (order 5) in level 4)

The most noticeable result of these experiments is the effect that phase distortions of high order IIR filters in the lower tree levels are highly reduced by the implementation of low order IIR filters only in the highest tree level. The decimation ratio in this level still guarantees sharp cutoff slopes. Fig. 4 shows the group delay as an example of a regular 16 channel IIR QMF bank with prototype filters of order 9 ($\psi_s = 0,52\pi$) in tree level 1-3 and order 5 ($\psi_s = 0,65\pi$) in level 4. This filterbank produces good perceptive reconctruction of the original signal. The group delay peaks at 0.125π, 0.375π, 0.625π and 0.875π are distinctively less noticeable as expected from the figure.

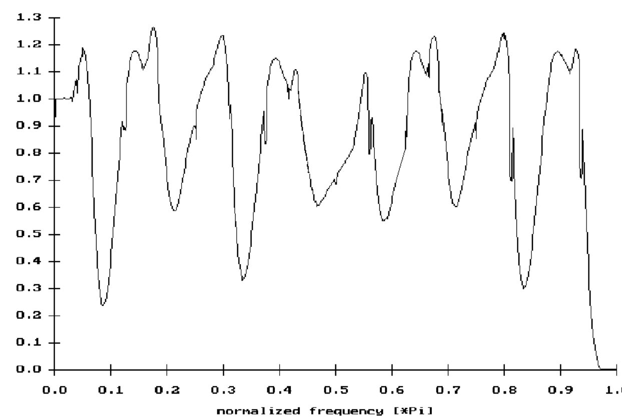
This effect gives rise to the implementation of low delay IIR QMF banks with superior sharp cutoff slopes compared to FIR filterbanks. With these implementations it is possible to exploit masking effects in speech coding by totally suppressing particular subbands and avoiding aliasing of the filterbank. To evaluate these properties, selective frequency responses have been measured using filterbanks with suppressed subbands of every even numbered band. Fig. 5 compares the selective frequency response of the regular 16 channel QMF bank using the IIR prototype filter distribution as in Fig. 4 and FIR prototype filters of order 15 in all tree levels.

For the adaptation to the logarithmic properties of the human ear it is useful to implement nonregular tree structures. Tab. 1 gives an example of an implemented filterbank using 5 tree levels and splitting the signal frequency band of 8 kHz into 10 subbands with almost logarithmic spaced bandwiths. Tab. 2 depicts a distribution of prototype filters for this structure which shows only hardly noticeable phase distortions.

Further improvements especially for nonregular tree



(a)



(b)

Figure 5: Selective frequency response of regular 16 channel QMF banks (a)IIR prototype filter distribution as in Fig. 4 (b)FIR prototype filters of order 15 in all tree levels

band	frequency
1	0- 250 Hz
2	250- 500 Hz
3	500- 750 Hz
4	750-1000 Hz
5	1000-1500 Hz
6	1500-2000 Hz
7	2000-3000 Hz
8	3000-4000 Hz
9	4000-6000 Hz
10	6000-8000 Hz

Table 1: 5 level tree structured IIR QMF bank with almost logarithmic spacing of the band edges

level	1	2	3	4	5
prototype $[\psi_s]$	$0,53\pi$	$0,53\pi$	$0,53\pi$	$0,53\pi$	$0,59\pi$

Table 2: Useful prototype distribution of the IIR QMF bank according to Tab. 1

structures can be achieved by compensation of phase distortions caused by asymmetrical processing of subbands. This method applies a constant delay G to subbands signals which are not further splitted in the tree structure.

$$G = \sum_{k=1}^{K} 2^k G_k \tag{15}$$

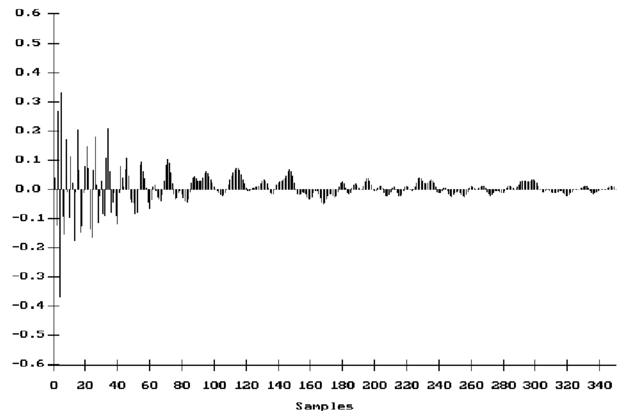
($G_k =$ group delay of prototype filter in level k at $\omega = 0$, $K =$ number of tree levels not used for further processing).

Fig. 6 shows the effect of this compensation on the impulse response of a 6 channel dyadic subband tree structure.

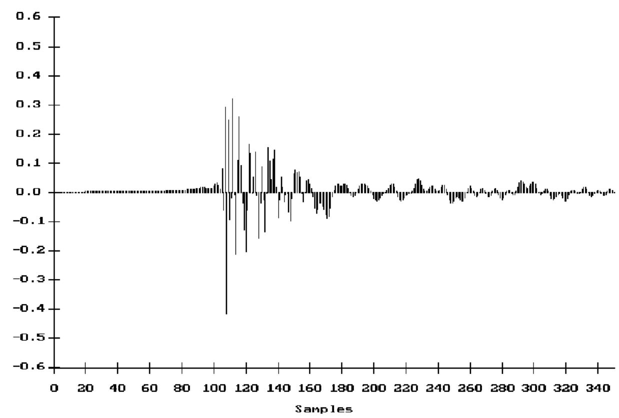
The reduction of phase distortions goes along with a narrow distribution of signal energy compared to non-compensated filterbanks but also higher average group delay.

5 CONCLUSION

Structural and compensative means in low delay IIR QMF banks allow a minimization of phase distortions, which make them more interesting for speech coding. The possibility of multichannel tree structures with a sufficient amount of subbands gives rise to applications in wideband speech processing using principles which are already established in music signal coding.



(a)



(b)

Figure 6: Effect of additional delay in a dyadic tree structure (6 channels, prototype filters with $\psi_s = 0,52\pi$ in all tree levels). (a)Without compensation, (b)compensation with G_k = 1,7 samples

References

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