

# A FLEXIBLE FILTERBANK STRUCTURE FOR EXTENSIVE SIGNAL MANIPULATIONS IN DIGITAL HEARING AIDS

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## ABSTRACT

Filterbanks for digital hearing aids must use significantly different criteria than those designed for coding applications. For digital hearing aids, the filterbank channel gains must be adjustable over a large dynamic range to compensate for the hearing loss. This adjustability violates the alias cancellation properties of critically sampled filterbanks designed for coding. This paper describes a filterbank designed exclusively for hearing aid applications. Consideration will be given to the extremely limited memory, low delay and low power requirements that must be met in a typical hearing aid application.

## 1. INTRODUCTION

It has long been known that hearing loss is a function of both frequency and input level. A filterbank (Fig. 1) provides a natural decomposition of the input signal into frequency bands, which may be processed independently to best compensate for the hearing loss and meet prescriptive targets. Although filterbanks have traditionally been constructed using analog techniques, digital filterbanks have a great number of advantages including precise control over the phase response, which greatly facilitates signal reconstruction.

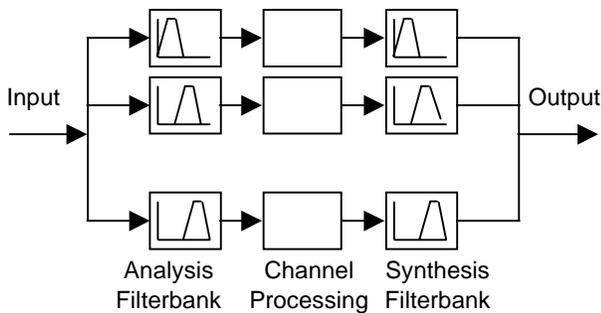


Figure 1. Filterbank signal processing

In this paper we describe an extremely flexible framework for separating the input signal into frequency bands which forms the basis for a multichannel compression hearing aid developed by the authors [1]. Filterbank design for hearing aids must address the extremely limited memory available, low delay requirement and the flexibility for accurate fitting. An oversampled, weighted overlap-add filterbank, designed to meet these requirements, will be presented in the paper.

Although non-uniform auditory critical bands are a better fit to hearing physiology, fast modulation techniques are not directly applicable. Much greater computational efficiency is obtained when the filter response shapes are realized as a series of modulations of a low-pass prototype filter covering the entire frequency range. This modulation produces identical filter shapes resulting in a uniformly spaced filterbank. Although a greater number of bands are needed to achieve sufficient resolution at low-frequencies, this additional computational expense is more than compensated for by the use of a fast modulation technique.

To achieve a further improvement in low-frequency fitting, an *even/odd* stacking modification was implemented which allows a selectable shift of the filterbank center-frequencies by one-half band (Fig. 2). This doubles the number of potential band edges compared to the number of bands. This filterbank has been implemented and operates on a Motorola DSP56301-based portable platform. It is suitable for low-power, real-time operation.

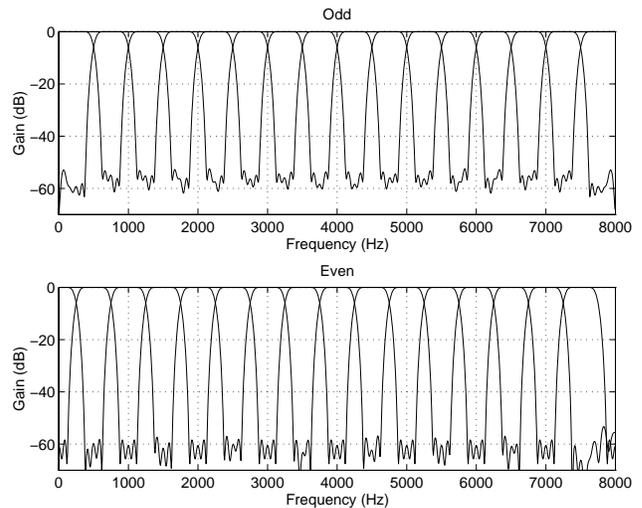


Figure 2. Frequency response of filter bank channels for odd and even channel stacking arrangements

## 2. CRITERIA

### 2.1 Coding Background

A filterbank decomposes the input signal into a series of separate frequency bands. By using designs that minimize the overlap between adjacent bands, the resulting representation is approximately orthogonal. It is natural that filterbanks have found extensive use in the low bit-rate coding of speech and image signals and have been heavily optimized for this case. The MPEG coder is an example used for high fidelity speech coding [2, 3].

For sub-band coding purposes, reliance is placed on the fact that typical spectra are not flat. This enables the coder to allocate more bits to the more perceptually important high-energy regions and fewer bits to the less perceptually important low energy regions of the spectrum.

At first glance, an  $M$ -band filterbank would appear to increase the data rate because the single input data stream has been split into  $M$  separate bands generating  $M$  times more data. It is possible, however, to *decimate* the data streams because of their reduced bandwidth. In the important case of *critical* sampling, only 1 in  $M$  samples is used (for an  $M$ -band uniform filterbank) in each band. The total data rate is thus unchanged through the filterbank, which is intuitively satisfying since no new information was added. The problem is that overlap between adjacent bands results in the generation of *aliasing* distortion, since any residual response in the adjacent bands is folded back into the original band by the decimation procedure.

In general, the most concise alias-free representation is only obtained if the filter bands directly abut each other with complete frequency coverage and no overlap. Although this is impossible, developments have led to filterbanks with slightly overlapping bands designed in such a way that aliasing distortion generated in the analysis stage is exactly canceled by imaging distortion in the synthesis stage [4, 5].

In the absence of coding (quantization of the bands), these filterbanks produce transparent results. Under increasingly coarse quantization, the effective gain changes (from the quantization step) result in the imaging distortion not completely canceling the aliasing distortion. In practice, this does not degrade the perceptual audio quality because the coarsely quantized bands are lower in energy and the noise floor masks the uncanceled aliasing distortion.

### 2.2 Hearing Aid Application

For hearing aid use, the frequency splitting is performed for the purpose of modifying the spectral shape of the input signal. Hearing aid fitting typically requires a wide gain adjustment range. In a compression system, the input signal level, which can be measured as the overall level, channel level or a combination, controls these gains [1].

Given the requirement for wide gain adjustment, the alias cancellation theory is invalid and critical sampling is insufficient. This problem necessitated the development of an oversampled

filterbank. Although oversampling increases the data rate, it is the price that must be paid for gain adjustability without aliasing.

In a compression system, gain changes are dynamic. This may cause anomalies in the overall frequency response if phase differences exist between adjacent bands. To avoid these undesirable frequency response notches or peaks at the band edges (which frequently occur in analog systems), it is necessary to constrain the filter channel impulse responses to be linear phase and of equal delay.

Thus, an ideal filterbank for a hearing aid application would: (1) allow precise fitting of prescriptive targets, (2) have short delay (3) be computationally efficient and (4) use a minimal amount of memory.

## 3. OVERSAMPLED WOLA FILTERBANK

The criteria mentioned in the latter part of the last section are contradictory. For example, better filter responses are obtained only at the expense of delay. Both cannot be simultaneously satisfied. Thus, the filterbank is an engineering compromise best arrived at through careful consideration of the performance parameters.

Initial experiments were conducted with critically sampled filterbanks. As mentioned before these filterbanks achieve good performance only in cases with mild channel gain changes. To achieve high-quality reproduction under widely varying gain changes, it was necessary to oversample by at least a factor of two to reduce the level of uncanceled aliasing generated when band gains differ greatly.

The selected design uses an oversampled, weighted overlap-add (WOLA) DFT filterbank [6, 7, 8] to split the input signal into 16 frequency bands. This filterbank uses modulation via the DFT to replicate a single prototype filter into 32 complex filter bands. This modulation produces identical filter shapes and results in a uniform filterbank. At a sampling frequency of 16 kHz, the resulting bands are 500 Hz wide. Total computational complexity is 46 multiply-accumulates per output point. Group delay is 12.5 ms.

To achieve sufficient frequency resolution at low frequencies, a uniform filterbank requires a larger number of bands than would be required by a non-uniform filterbank. Fortunately, the DFT method described above generates a large number of bands at low computational expense. This large number of bands is required to achieve a good fit to audiometric data at low frequencies (which is normally given on a log-frequency scale). As a result of the linear frequency spacing and large number of channels, the high frequency band spacing may be greater than necessary. Thus, it may be advantageous to group bands at high frequencies for gain adjustment purposes.

It is important to realize that better frequency resolution is available only at the expense of greater signal delay for *any* filterbank method. Long delays are not desirable—delays of 6-8 ms are reported to be just noticeable. Delays longer than 20 ms may cause interference between speech and visual integration [9]. Clearly, less delay is better.

For audiometric fitting, a useful compromise is available in the form of the *even* or *odd* stacking of the filterbank (Fig. 2). Effectively, this procedure doubles the effective length of the FFT to  $2N$  points but selects only half the number of bins.

Mathematically, the forward and inverse even and odd DFT transform pairs (Equations 1 and 2) are given by:

$$1) \quad Y_k = \sum_{n=0}^{N-1} y(n) W_N^{-(k+\nu)n}, k = 0, 1, \dots, N-1$$

$$2) \quad y(n) = \frac{1}{N} \sum_{k=0}^{N-1} Y_k W_N^{(k+\nu)n}, n = 0, 1, \dots, N-1$$

Where

$$W_N = e^{-k(2\pi/N)}$$

FFT algorithms were developed for both cases. For even stacking,  $\nu$  is zero and the conventional  $N$ -point FFT may be used. For odd stacking,  $\nu$  is 0.5 and the conventional FFT must be extended to be an odd FFT [10]. This additional choice for  $\nu$  doubles the number of bin locations without requiring a longer analysis and allows band edges to be placed within one-half the width of the filterbank channel.

## 4. FILTERBANK OPERATION

### 4.1 Overview

The frequency response of the analysis window (i.e., the prototype low-pass filter) is modulated by the odd (even) FFT to produce channel responses as illustrated in Fig 2. The individual channel signals are decimated by  $N/OS$  where  $N$  is the FFT size and  $OS$  is the oversampling factor. It is critical that the analysis lowpass filter be sufficiently sharp to minimize the aliasing distortion generated by the decimation step.

The spectral shape of the input signal is modified at this point by applying suitable gains to the frequency channel signals. This is followed by the corresponding inverse odd (even) FFT, interpolation, synthesis window weighting and the overlap-add procedure. This window (a low-pass filter which is the counterpart to the analysis window), minimizes the spectral imaging distortion created during the interpolation step.

To conserve memory, it is highly desirable that the synthesis window be derived from the analysis window. Fortunately, the oversampling of the channel signals relaxes the requirements on the synthesis window. The channel images are spaced at intervals of  $OS$  rather than 1 as in critical sampling. Since the synthesis window low-pass function need only reject images  $OS$ -channels away it can be set as a decimated version of the analysis window. The decimated window also has the advantage that synthesis delay (half the synthesis window length) is significantly reduced.

### 4.2 Analysis/Synthesis Window Design

Both time and frequency domain constraints must be placed on the analysis and synthesis windows. For a WOLA DFT filterbank with  $N/2$  frequency bands (i.e., one that uses an  $N$ -point FFT) and band outputs that are decimated by  $N/OS$  ( $OS=2$  was used above), *time domain constraints* must be placed on the analysis window coefficients such that a zero appears every  $M_1$  samples (where  $M_1=N$ ) [6]. The synthesis window must have zeros every  $M_2$  samples where ( $M_2=N/OS$ ). For  $OS=2$ , this constraint (i.e., a synthesis window with zeros at half the spacing of the analysis window) can be met by decimating the analysis window coefficients by a factor of two.

*Frequency domain constraints* must be placed on the combined analysis/synthesis window frequency response such that both windows are “good”  $M$ -band filters. The analysis window must have a cutoff frequency of  $\pi/M_1$  and the synthesis window must cutoff at  $\pi/M_2=2\pi/M_1$ . This constraint can be met by designing an analysis window, that does not “droop” at half it’s cutoff frequency (i.e.,  $\pi/2M_1$ ), which becomes the cutoff frequency of the (decimated) synthesis filter.

A recently developed method of designing  $M$ -band filters that allows simultaneous time and frequency constraints to be placed on a design is the eigenfilter approach [11]. This approach was used to design the combined analysis/synthesis window.

### 4.3 Analysis

An illustration of the analysis portion of the filterbank is shown in Fig. 3. The main sequence of events is

- Read  $R$  input block samples
- Read sign from analysis sign table at the sign table pointer
- Apply sign to samples
- Circularly increment input sign table pointer
- shift input FIFO and add  $R$  new samples
- Apply window and time-fold to  $N$  samples
- Apply circular shift of  $(n \bmod N)$
- Take  $N$ -point FFT (even or odd)
- Apply channel gains to (complex) frequency data

### 4.4 Synthesis

The schematic of the synthesis operations is shown in Fig. 4. The operations are

- Take inverse FFT of (complex) input (even or odd)
- Apply circular shift of  $(n \bmod N)$
- Periodically extend to  $L/DF$  samples
- Apply synthesis window
- Accumulate into output FIFO, shift out  $R$  samples
- Read sign from synthesis sign table at the sign table pointer

- Apply sign to the  $R$  shifted out samples
- Circularly increment sign table pointer to next sign value
- Circularly increment  $n \pmod{N}$  by  $R/OS$

Where:

- $R$  is the input block size
- $L$  is the input window size
- $N$  is the FFT size
- $DF$  is the decimation factor
- $OS$  is the oversampling factor

The sign table contains the sign factors in a repeating pattern (+1, +1, -1, -1, +1, +1, -1, -1) used to modulate the input sequence and the output sequence in the case of odd stacking. For even stacking, the sign factors are all +1.

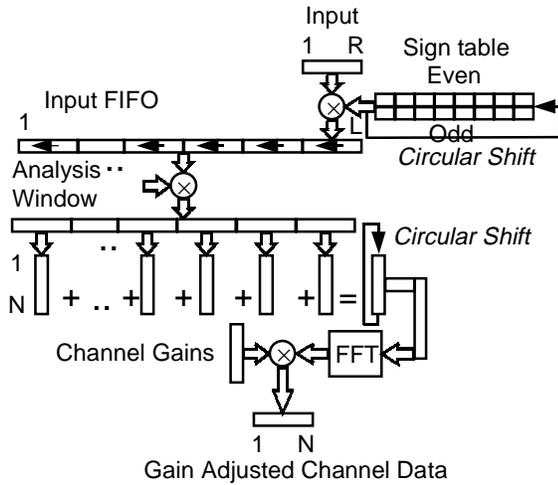


Figure 3. Schematic of filterbank analysis

## 5. CONCLUSIONS

A highly flexible filterbank structure has been described in this paper. Several tradeoffs have been made to make this a practical filterbank that can meet the requirements of a hearing aid application.

The filterbank structure is  $M$ -band with uniform bands. This structure provides many advantages in a hearing aid context. To conserve memory, only a single analysis window is stored. A DFT is used to modulate and replicate this lowpass prototype. To further reduce memory, the synthesis window is created by decimating the analysis window, subject to time and frequency domain constraints that can be satisfied by using an eigenfilter design method. This filterbank was implemented and runs in real-time on a Motorola DSP56301-based portable system. It has been used successfully for a real-time compression system implementation [1].

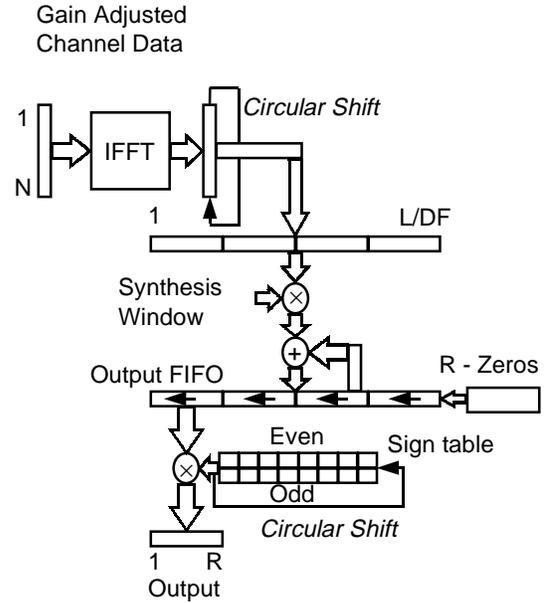


Figure 4. Schematic of filterbank synthesis

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