Development of Low Cost Filter Banks for Audio Applications

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Abstract

An approach for low cost uniform and octave filter banks for Audio applications is proposed. The analysis stages of these filterbanks are based on IIR allpass second order sections with modified response. The modification is performed by forcing phase non-linearity at the analysis stage to be out of region of interest. By oversampling, the non-linear segments near the band edges are removed through subsequent synthesis filtering. Compared to existing literature designs, the new approach offers a substantial lowering in computational power, and a lower input/output delay.The simulation and testing of this technique was performed using MATLAB software package.

Keywords: Hearing Aids, Filterbanks, IIR filters, polyphase.

تطوير مصفوفة مرشحات ذات جهد حسابي قليل مناسبة للاغراض السمعية

الخلاصة

يقترح هذا البحث مصفوفة مرشحات رقمية مناسبة للاغراض السمعية التي تتطلب جهد حسابي قليل .تتضمن هذه المصفوفة مرحلة تحليل ومرحلة تركيب على افتراض ان عمليات المعالجة تتم مابين هاتين المرحلتين. مرحلة التحليل التي ترتكز على استخدام المرشحات الكلية التمرير (IIR ssaplla) ذات المقاطع ثنائية الاس . وتحور هذه المرشحات بحيث تكون الاستجابة الغير خطية خارج منطقة الاهتمام يتم زيادة تردد احد العينات للاشارة الداخلة اكثر من المطلوب بحيث يتم ازالة الاجزاء ذات الاستجابة الغير خطية والتي تسبب تشويش الاشارة الخارجة في مصفوفة مرشحات التركيب . اعطى هذا الاسلوب تخفيض كبير في الجهد الحسابي وتاخير قليل مقارنة بالطرق المستخدمة تقليدياً كما في مرشحات FIR .

1. Introduction

Digital signal processing for audio applications provides new signal processing strategies to compensate for hearing loss. However, the digital circuitry intended for this purpose needs to fulfill requirements of low power consumption, low supply voltage and small size to be usable in such applications. It is well known that hearing loss compensation is a function of both frequency and input power [1]. Filter bank design for hearing aids must address the limited memory available, low delay requirement and the flexibility for accurate fitting which may be processed independently to best compensate for the hearing loss. However, high order filters for selective channel separation is required. Modulating a single prototype filter to produce a filterbank provides a natural decomposition of the input signal into frequency bands. This is achieved by a suitable

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transform such as DFT (Discrete Fourier Transformer) or DCT (Discrete Cosine Transformer), which have a great computational efficiency [2]. Filters produced this way are uniformly shifted versions of the low pass prototype filter. Non-uniform auditory critical bands are a better fit to hearing physiology. However, fast modulation techniques are not directly applicable to this kind of filter banks. An improvement in computational cost is achieved by polyphase decomposition so that computational complexity can be reduced by the number of decompositions.

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 requi
res 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 not applicable 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. A filterbank that is suitable for hearing aid applications would allow exact fitting of prescriptive targets with short time delay. It was stated in [3] that delays longer than 20 ms may cause interference between speech and visual integration, clearly, less delay is better. In addition to that the filterbank should be computationally efficient and uses a minimal amount of memory. An even/odd stacking strategy was implemented in [4].

Also a design uses an oversampled, weighted overlap-add (WOLA) DFT filterbank is proposed in [2]. This filterbank uses modulation of a single FIR prototype filter into 32 complex finite impulse response FIR filter bands.

Furthermore, a window based method for near-perfect reconstruction prototype FIR filters that uses highly oversampled, complex modulated filterbanks has been proposed in [5], this extends the idea of critically sampled filterbanks to the oversampled case with different length analysis/synthesis FIR filters. In all literature mentioned above, the filterbanks are based on FIR filters which are computationally expensive and need fast processing to avoid mismatching. In addition to that, design methods which are proposed in a very recent publication [6] are still FIR filtering dependent. On the other hand, non-uniform filterbanks have been treated well in literatures [7, 8, 9, and 10]. These designs are also FIR class filterbanks. In this paper a lower cost implementation is proposed. Our implementation is based on oversampled IIR analysis filterbanks. Distortion due phase non-linearity was removed at the synthesis stage through simple strategy.

2. A low cost filterbank strategy.

Typical implementation of filterbanks for hearing aids purposes is shown in Figure 1, H_k are the analysis filters, G_k are the synthesis filters, D is the subsampling factor, for k=0,1,2,3…..M-1, where M is the number of bands.

The requirements on the analysis filters are stringent, they should have a fairly flat pass band so that the magnitude frequency response is not distorted and a narrow transition bands so that only a small amount of unwanted energy is let in.

Optimal filters satisfying the above requirements such as elliptic filters [11] are optimal in the minimax sense. However, they have a very nonlinear phase response around the band edge. In high quality auditory systems this is considered to be objectionable [12]. A strategy to solve this problem is to oversample the input signal by a factor of two or more. The filter *H(f)* in Figure 2(c), now has a much wider transition band, so that the phase-response nonlinearity is acceptably low. The obvious price paid in this is the increased internal rate of computation.

To reduce implementation cost and lowering power consumption, the analysis stage is implemented with infinite impulse response IIR filterbank. The analysis prototype filter is constructed from second order allpass sections as shown in Figure 3. This implementation can be realized in polyphase arrangements as shown in Figure 4, thus reducing the implementation cost further down by a factor of two, which results in half the number of calculations per input sample and half the storage requirements.

The polyphase structure can be modified by shifting the downsampler to the input to give more efficient implementation [2]. In Figure 4, y_0, y_1 represent lowpass and highpass filter outputs respectively.

$A_0(z)$ and $A_1(z)$ are causal real stable allpass filters. Elliptic filters fall into this class of filters yielding very low-complexity analysis filters. The transfer function of the prototype analysis filter for L sections is given by:

$$H(z) = \sum_{k=0}^{N-1} A_k(z^N) z^k \quad \text{….(1)}$$

Where:

$$A_k(z^2) = \prod_{n=1}^{L_k} A_{k,n}(z^2) = \frac{a_{k,n} + z^{-N}}{1 + a_{k,n} z^{-N}} \quad \text{….(2)}$$

$a_{k,n}$ is the coefficient of the *kth* allpass section in the *n*th branch, L_k is the number of sections in the nth branch and N is the number of branches in the structure. These parameters can be determined from filter specifications, with a compromise between the number of zero sections and phase deviation from linearity [13]. Because of the small number of calculations required per filter order and very high performance, such a structure is suitable for filtering requiring high speed of operation and high levels of integration. Furthermore the polyphase IIR filter structure used here is not very sensitive to coefficient quantization [2], which makes a fast fixed-point implementation a viable option.

On the other hand, the synthesis filter bank is constructed form a prototype FIR filter that is related to the analysis prototype filter in such a way that the distorted components due to phase non- linearity is removed. These components are assumed to be of negligible importance to the audio band of interest. We have defined a new point on the frequency axis Fla, we called it the end of non linearity point. This relationship is illustrated diagrammatically by Figure 5. A demonstration of this situation is shown in Figure 6. This figure depicts a magnitude and phase responses of filters covering the audio range of interest. Thanks to oversampling we can hopefully recover the band of interest by synthesis filtering without any distortion.

A general relationship in the z-domain is set to model the system as follows:

$$\sum_{k=0}^{M-1}\sum_{m=0}^{D-1} G_k(z)H_{k,m}(z)X_{k,m}(z) = z^{-\partial}\hat{X}(z) \quad \text{….(3)}$$

Where k, is the subband index, and m is component index due to downsampling, i.e. aliasing components, ∂ is the system delay. Since the subband gains will be adjusted individually, then the aliasing issue will be dropped and equation (3) reduces to;

$$\sum_{k=1}^{M-1} H_k(z)G_k(z)X_k(z) \approx cz^{-\partial}\hat{X}(z) \quad \text{…..(4)}$$

Where c is a scalar constant once the analysis prototype filter is designed, the prototype synthesis filter can be optimized in the frequency domain by minimizing an objective function T_0 ($e^{j\omega}$), which represents the distortion function in the individual subbands. T_0 ($e^{j\omega}$) was deduced from (4);

$$T_0(e^{jw}) = \left|\, 1 - \left|H_0(e^{jw})G_0(e^{jw})\right|^2 \,\right| \quad (5)$$

For the prototype analysis / synthesis prototype filter pair, $H_0(e^{jw})$, and $G_0(e^{jw})$, respectively.

T_0 ($e^{j\omega}$) is the amplitude distortion, that is minimized over the frequency region of the first subband. Initially, G_0 ($e^{j\omega}$) was designed using Hamming window to meet cut off requirements, then optimized in (5) to achieve minimum amplitude distortion.

Now Consider the arrangement shown in Figure 7, the input signal *x(n)* is decomposed into sub-signals with the aid of analysis filterbank *H(z)*.This filterbank is an octave implementation which closely matches the frequency response of the human perceptual ability.

In each level of the decomposition, a half band infinite impulse response filter IIR, is used for signal splitting. The frequency response of this filter was modified to meet passband constrains defined by Figure 5. The implementation complexity can be reduced by the use of polyphase decomposition in each stage. Another cost reduction can be achieved by shifting the down-samplers to the inputs of the filters utilizing the noble identities. Table 1 shows the frequency allocation in each level. Figure 8 depicts the magnitude and phase responses of analysis and synthesis prototype filters. These filters were designed on normalized frequency bases to fit all stages.

As far as reconstruction of processed signal is concerned, the synthesis filterbank is implemented in reversed manner compared to the analysis stage as shown in Figure 9. The prototype synthesis filterbank is based on FIR filter to ensure stability plus removal of phase distortion due to the use of IIR filters at the analysis stage. The computational complexity of this part is comparable to that suggested in literature. However, computational gains are actually obtained at the analysis stage.

1. Simulation and Results.

In the first case**,** a uniform filter bank with 16 bands providing frequency resolution down to 500 Hz were constructed and simulated. This filterbank is implemented in a tree fashion similar to that mentioned in [14], with a difference in passband frequency specifications. Figure 10 depicts the frequency response of the uniform filterbank. The Analysis stages of both uniform and octave filter banks are based on a prototype filter, designed for 70 dB stopband attenuation and 0.5 dB pass-band ripple.

To measure the reconstruction error, a sinusoidal signal, sweeping linearly within the audio range of interest, here considered from 1 to 8000 Hz, is used to excite the entire filter bank as shown in Figure 11a. Figure11b depicts the reconstructed signal.

Although Figures 11a and 11b are evidence of the goodness of the approach, they show only a small piece the actual signal. Picture of the entire sweep is shown in Figure 12; a severe distortion at the end of each sweep cycle was noticed. This however, should not be discouraging, as the synthesis filters are constrained to have a stringent attenuation near the band edges, avoiding phase non-linearity distortion. In addition to that, frequency components around 8 kHz and above are considered to be of negligible importance to human hearing.

Amplitude variations ranging from 0 to 0.095% of input signal amplitude was observed in the first case. This range of amplitude variation could hardly be detected by human cochlea. A spectral comparison between original and reconstructed signals is shown in Figure 13. It is obvious that the power spectrum of the reconstructed signal is a close match to the original one.

In the second case, a non uniform filter bank has a response as shown in Figure14, was used to decompose the audio signal as in the first case. This filterbank is a close match to the cochlea frequency response, i.e. it is a mimic of human perception. A speech signal as shown in Figure15a was used to test the reconstruction capability of the filterbank which was designed on normalized frequency bases.

Apart form the attenuation that the processed signal has suffered from, the signal was reconstructed fairly accurately as shown in Figure 15b. It was found out that amplitude distortions are not easily noticed by average listeners. The attenuation issue can be treated by suitably adjusting the individual subbands at the output of the analysis stage. Since this design is targeted for hearing disables, it is usual to adjust the gain of the individual frequency band to fit the disabled preference.

The main advantage of the new approach is in its low number of multiplications required per input sample at the analysis stage. The existing equivalent two fold oversampled FIR approach requires at least a 128-tap filters FIR for equivalent frequency selectivity. For 16 band filter bank, the overall multiplication requirements are 2048 multiplications for subband filtering at the analysis stage. In comparison, the use of the multirate polyphase IIR structure at the analysis stage allowed a decrease in the number of multiplications per input sample to 24 if downsampling is performed after polyphase decomposition, and only half of that when downsampling is performed before polyphase decomposition .This is the case since half the calculations performed at the odd sample intervals and the rest of them at the even sample intervals according to the noble identities [2]. The total signal delay due to filterbank insertion was calculated to be around 5.5 ms, this delay value is much lower than the maximum acceptable delay in hearing aids applications.

4. Conclusions.

A design method for low complexity filterbanks is proposed. These filterbanks are targeted for hearing aids. The analysis stages of these filterbanks are based on IIR allpass second order sections with modified response. Using the proposed approach, a substantial computational savings can be obtained. The maximum amplitude error was calculated to be less than 0.095% of input signal amplitude. This amplitude variation is tolerable by human hearing. Also, input/output signal delay due to filterbak insertion was calculated to be 5.5 ms. this delay is of no effect on visual integration in hearing aids.

References

[1] Schneider T, Brennan RL, A Multichannel Compression Strategy for a Digital Hearing Aid. Proc. *ICASSP*-97, Munich, Germany, pp. 411-415.

[2] Vaidyanathan PP, Multirate Systems and Filter Banks. Prentice-Hall Inc., 1993.

[3] Agnew J, An Overview of Digital Signal Processing in Hearing Instruments. The Hearing Review, July 1997.

[4] Brennan RL, Schneider T, A Flexible filter structure for extensive signal manipulations in digital hearing aids.. Proceedings of the 1998 IEEE International Symposium on Circuits and Systems, 1998. ISCAS '98.

[5] Herman D , Chau E, Dony RD, Areibi SM, Widwow based prototype filter design for highly oversampled filterbanks in audio applications, Procedings of IEEE ICASSP Honolulu ,16-20 April 2007.

[6] Dumitrescu B, Bregovic´ R, Sarama¨ki T, Design of low-delay nonuniform oversampled filterbanks", ElsevierSignal Processing 88 (2008) 2518– 2525, www.elsevier.com/locate/sigpro

[7] Princen J, The Design of non-Uniform Modulated Filterbanks, IEEE Transactions on Signal Processing, VOL. 43, NO. 11, November 1995.

[8] Cvetkovic Z. Nonuniform Oversampled Filter Banks for Audio Signal Processing. IEEE Transactions on Speech and Audio Precessing, VOL. 11, NO. 5, September 2003.

[9] Mariane R, Petragliaand Paulo, Batalheiro B. Nonuniform Subband Adaptive Filtering With Critical Sampling” IEEE Transactions on Signal Processing, VOL. 56, NO. 2, February 2008.

[10] Dumitrescu B, Bregovi R, Saram T. Simplified Design of Low-Delay OversampledNPR GDFT Filterbanks. EURASIP Journal on Applied Signal Processing Volume 2006, Article ID 42961, Pages 1–11.

[11] Antoniou A. Digital Filters: Analysis and Design. New York: McCraw Hill, 1979.

[12] Special Issue on Digital Audio, *IEEE* ASSf Mag. Vol. 2, Oct. 1985.

[13] Ekiz O, Constantinides AG. Phase linearity in polyphase filters. Proceedings of IEEE ICASSP Volume1,5-9 June2000 Page(s):321-3.

[14] Krukowski A, Kale I. Polyphase IIR Filter Banks for Subband Adaptive Echo Cancellation Applications. Proceedings of the 2003 International Symposium on Circuits and Systems, 2003. ISCAS '03.

Table (1) Subband Frequency Allocation

Band No.	Frequency Allocatior	Bandwidth
0	0 - 125	125
1	125 - 250	125
2	250 - 500	250
3	500 - 1000	500
4	1000 - 2000	1000
5	2000 - 4000	2000
6	4000 - 8000	4000

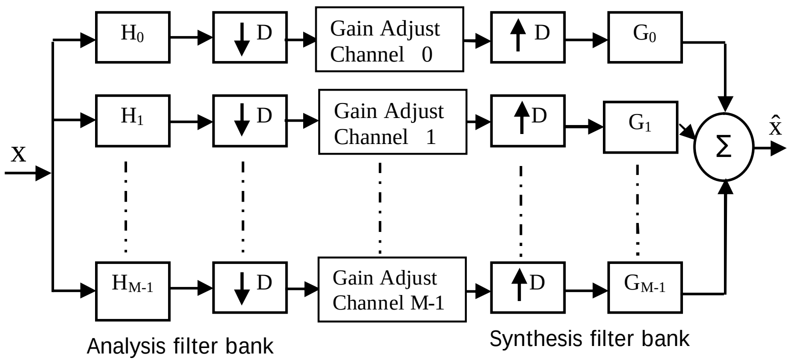


Figure (1) Filterbank for hearing aids, typical audio range is from 1Hz to8 kHz, in 500Hz intervals

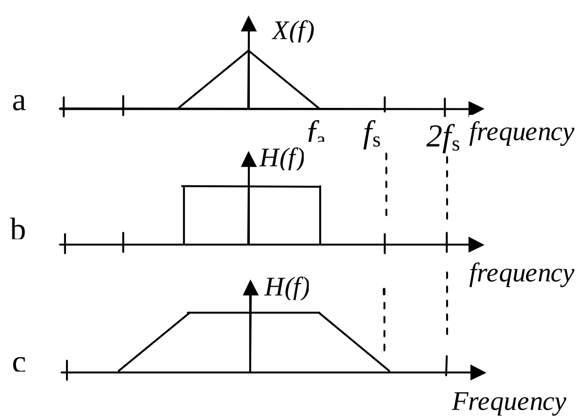


Figure (2) (a) Spectrum of input signal which has no spectral components above f_a, (b) Ideal magnitude response of a band limiting filter, (c) Non ideal filter characteristics. Fs is the sampling frequency

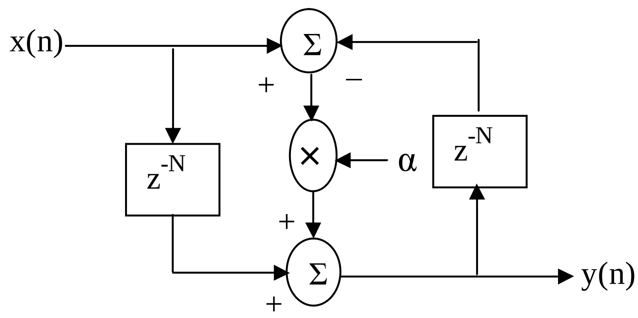


Figure (3) The second order all-pass section

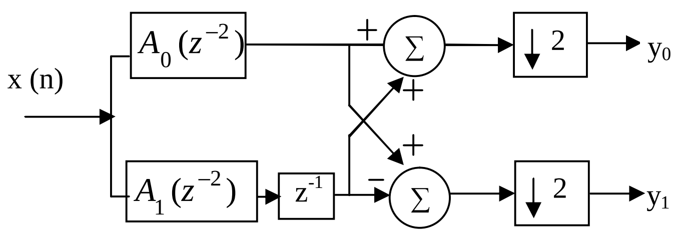


Figure (4) The polyphase implementation

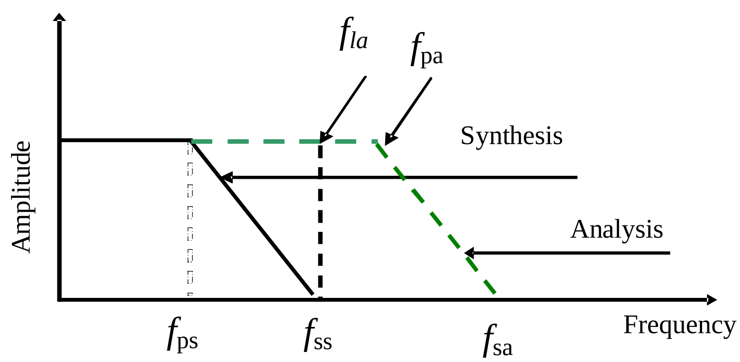
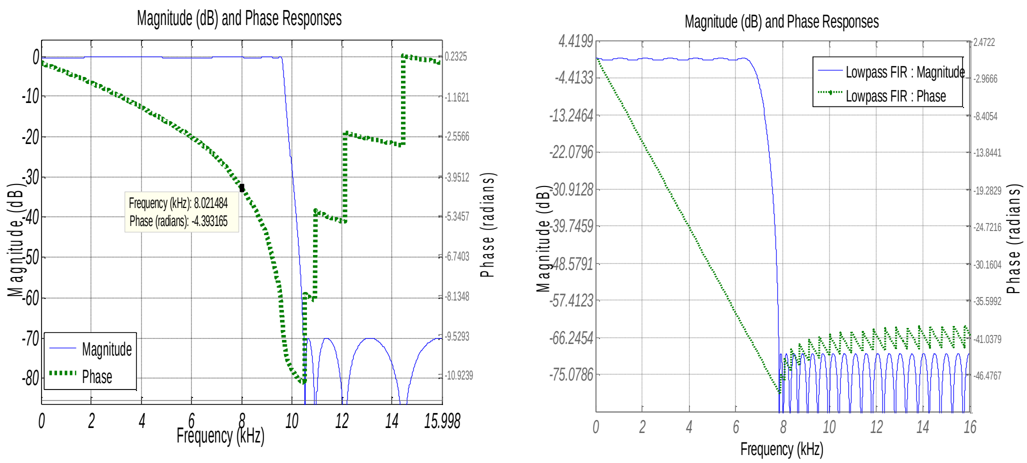


Figure (5) Illustration of the relationship between analysis and synthesis filters f_{pa} is the end of the synthesis pass band, f_{ss} is the beginning of the synthesis stop band, f_{la} is the end of non-linearity in the analysis filter f_{pa} is the end of the analysis pass band .



(a) (b)

Figure (6) a) An elliptic IIR filter deigned to pass up to 8 Hz with approximately linear phase characteristics. b) Prrks-MacLallen FIR filter designed to suppress frequency components above 8 kHz

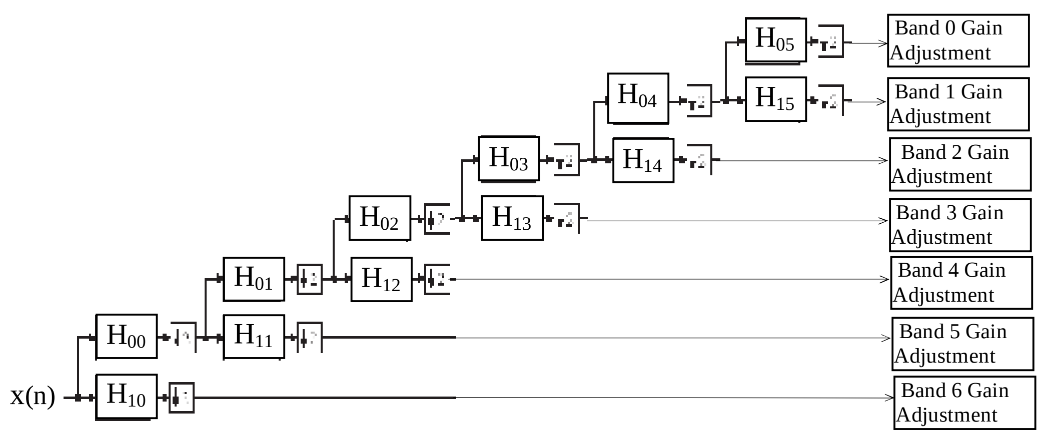
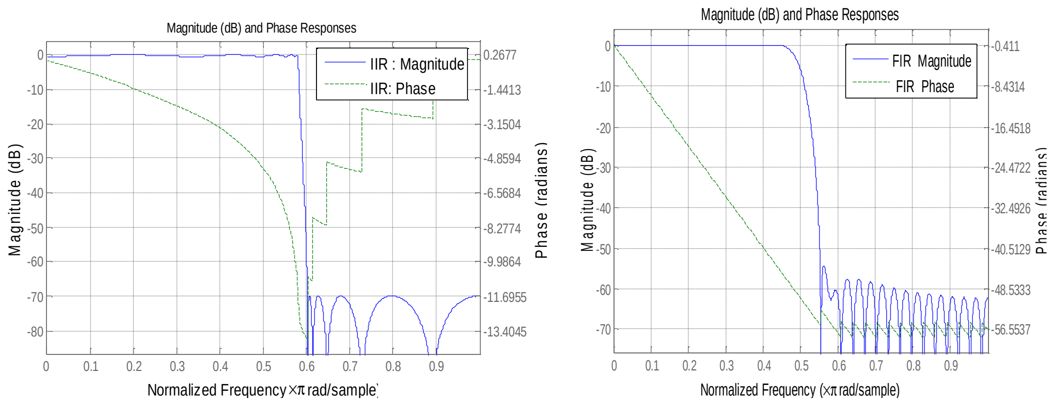


Figure (7) The analysis stage of the octave filterbank arranged for spectral modification in hearing aids



(a) (b)

Figure (8) Magnitude and phase responses of: a) Analysis prototype filter, b) Synthesis Prototype filter

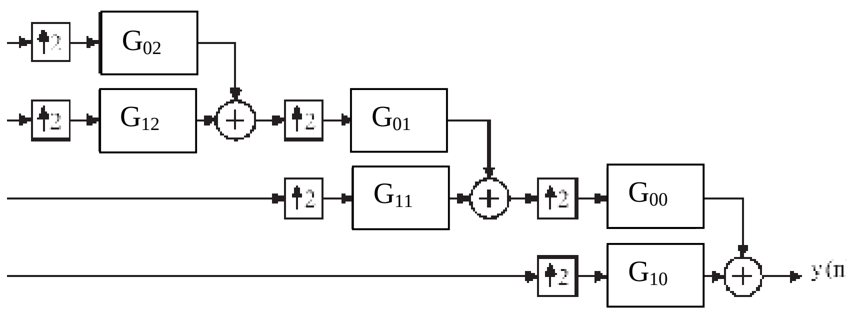


Figure (9) The final stages of the reconstruction process of the non uniform filterbank

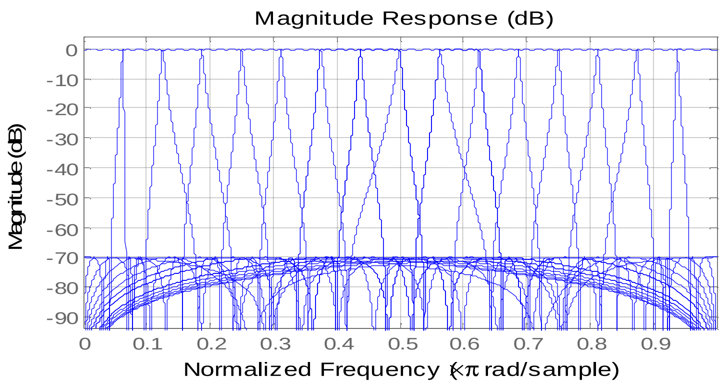
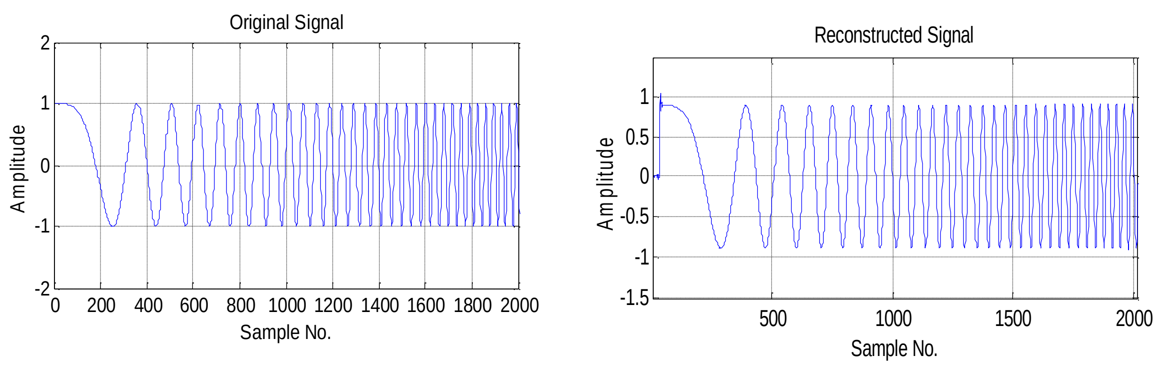


Figure (10) Frequency response of the uniform filterbank



(a) (b)

Figure (11) a) The original audio signal, b) Reconstructed signal by uniform filterbank

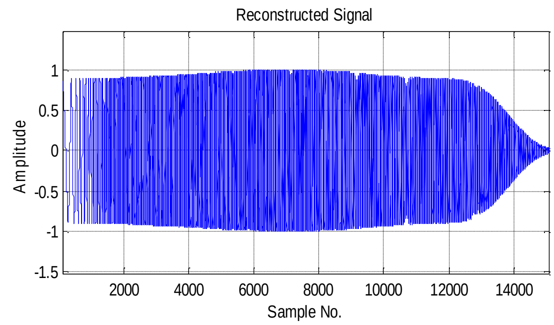


Figure (12) The reconstructed entire sweep

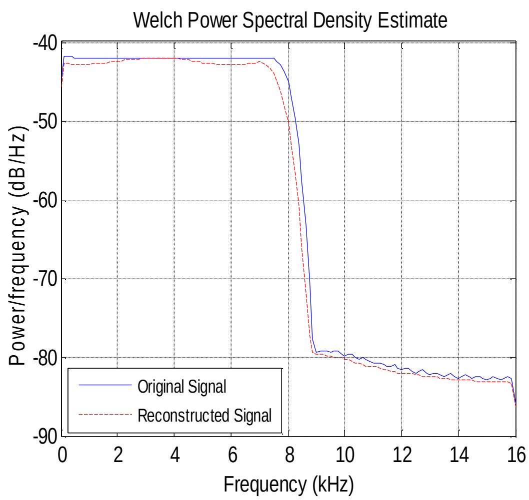


Figure (13) Spectrum estimate of original and reconstructed signals

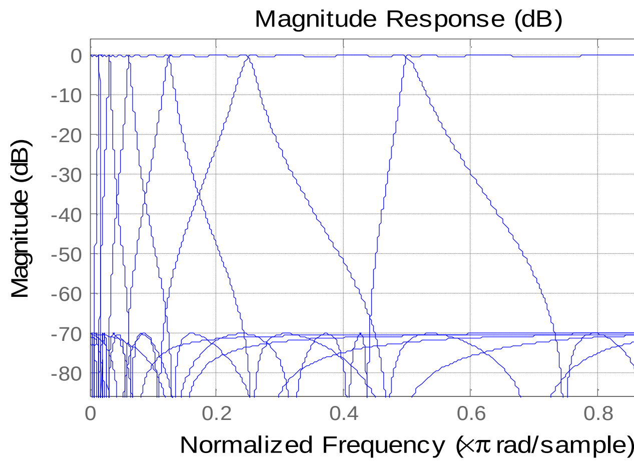
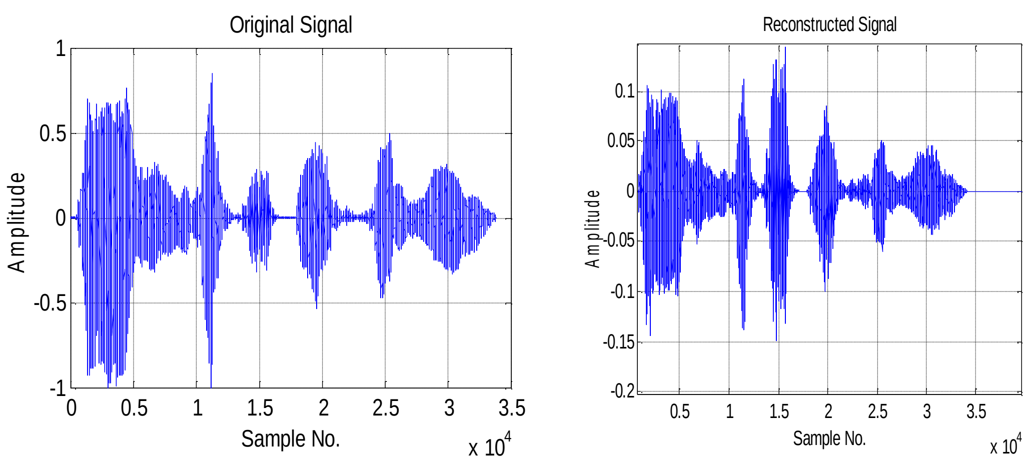


Figure (14) Response of the non-uniform filterbank



(a) (b)

Figure (15) a) Input speech signal used to test the non-uniform filterbank, b) Processed speech signal