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Delay Efficient Cosine Modulated Reconfigurable Filter Bank for Digital Hearing Aids Targeting Noise Induced Hearing Loss

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Abstract: In this paper, an efficient architecture for cosine modulated reconfigurable filter bank with sectional reconfigurability is proposed for Digital Hearing Aids targeting Noise Induced Hearing Loss. Polyphase implementation of the prototype filter with shared modulation coefficients reduces the computational complexity and increases the tuning flexibility. The proposed architecture can be used for frequency compensation of different classes of audiograms while keeping the group delay minimum. The proposed filter bank can provide 1296 various frequency band distribution schemes based on external control signals according to the nature of the audiogram. The architecture is based on a 16-band cosine modulated filter bank with dynamic merging of bands based on the audiogram. The flexibility of the filter bank allows matching most of the audiograms, including low or moderately-sloping SNHL and NIHL, with a notch at the center of the frequency range. For all the standard audiograms considered for evaluation, the proposed design has acceptable matching errors. The group delay is much less compared to existing reconfigurable filter banks, which provide room for accommodating other DSP algorithms in sophisticated Digital Hearing Aids. The hardware resources required for implementation are comparable with other reconfigurable filter banks designed for the same application.

Keywords: Filter bank, Cosine Modulated Filter Bank, Digital Hearing Aid, Delay Efficient Architecture

Zakasnilno učinkovita arhitektura nastavljive filtrirne banke s kosinusno modulacijo za digitalne slušne pripomočke ob izgubi sluha zaradi hrupa

Izvleček: V tem dokumentu je predlagana učinkovita arhitektura za nastavljivo filtrirno banko s kosinusno modulacijo z možnostjo sekcijske rekonfiguracije za digitalne slušne pripomočke ob izgubi sluha zaradi hrupa. Več fazno izvajanje prototipnega filtra s skupnimi modulacijskimi koeficienti zmanjšuje računsko zapletenost in povečuje prilagodljivost nastavljanja. Predlagano arhitekturo je mogoče uporabiti za frekvenčno kompenzacijo različnih razredov avdiogramov, pri čemer je skupinska zakasnitev minimalna. Predlagana banka filtrov lahko zagotovi 1296 različnih shem porazdelitev frekvenčnih pasov na podlagi zunanjih kontrolnih signalov glede na naravo avdiograma. Arhitektura temelji na 16-pasovnem kosinusno moduliranem filtru z dinamičnim združevanjem pasov na podlagi avdiograma. Prožnost banke filtrov omogoča prilagajanje večini avdiogramov, vključno z nizko ali zmerno nagnjenimi SNHL in NIHL, z zarezo na sredini frekvenčnega območja. Pri vseh standardnih avdiogramih, ki so bili upoštevani pri ocenjevanju, ima predlagana zasnova sprejemljive napake ujemanja. Skupinska zakasnitev je v primerjavi z obstoječimi nastavljivimi filtrirnimi bankami veliko manjša, kar zagotavlja prostor za namestitev drugih algoritmov DSP v zahtevnih digitalnih slušnih aparatih. Strojna sredstva, potrebna za izvedbo, so primerljiva z drugimi nastavljivimi filtrirnimi bankami, zasnovanimi za isto uporabo.

Ključne besede: filtrirna banka, kosinusna modulacija, digitalen slušni aparat, zakasnilno učinkovita arhitektura

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1 Introduction

According to recent research by the World Health Organization (WHO) [1], hearing-related health concerns are rising globally, including among younger generations. Digital Hearing Aids (DHA) are one of the prominent treatment options for Hearing Impairment (HI). Despite this, the usage of hearing aids is only at 17% globally, with low-income countries having limited access to DHAs. As a result, enhancing the effectiveness of DHAs is a crucial area of investigation that could improve their accessibility and value. Research towards incorporating sophisticated DSP algorithms in DHA for noise reduction, compression and intelligibility faces various challenges like increased chip area, delay, design complexity, and power consumption [2]. Similarly, the findings from audiology pose strict design constraints for DHAs [3].

Audiograms are graphs that show an individual's hearing threshold at different frequencies, measured in decibels (dB). Audiograms are used to determine the type of Hearing Impairment (HI) of a person. The audiologist measures a Patient's hearing thresholds against the frequencies between 250 Hz and 8000 Hz and marks it on a logarithmic graph, as seen in Figure 1. Hearing Impairment is broadly classified as Age-Related Hearing Impairment (ARHI), also known as presbycusis, and Noise-Induced Hearing Loss (NIHL) [4]. ARHI is considered the most common hearing disorder [5], leading to chronic disability in older age and is also referred to as Sensory Neural Hearing Loss (SNHL). On the other hand, NIHL primarily affects the working class, usually in factories with considerable noise and the young generation who habitually listen to music at high volume [6,7]. Low or moderate slopes in the low-frequency regions of the audiogram characterize SNHL [8]. However, NIHL is characterized by frequency notching in the mid-frequency regions of the audiogram [9]. Two sample audiograms [10] demonstrating SNHL and NIHL are given in Figure 1.

The critical component in a DHA is the Filter Bank, which separates the individual frequency bands in the hearing spectrum for frequency compensation based on Hearing Loss. Hence, it is a straightforward approach to explore the possibilities of designing efficient filter banks to accommodate other DSP circuits in sophisticated DHAs. The DHA amplification characteristics are programmed by the hearing aid technician based on the audiogram measurements through a wired or wireless interface between the Computer and DHA. This process is known as audiogram matching. When the slope of the audiogram is more at a particular frequency region, a large number of narrow bands are required for error-free audiogram matching and frequency compensation. However, this may increase the computational complexity and group delay of the filter bank.

Traditional approaches like Frequency Response Masking (FRM) [11] focus on the design of a computationally efficient filter bank. However, it suffers from delay problems restricting its usage in advanced DHAs. In earlier works on Digital Audiology Technology [12], it is shown that compromise in the Linear Phase of the filter bank, up to a certain extent, is not sensible to the human ear. Based on this finding, various approaches to designing filter banks using modulation techniques have been explored. One of the popular methods is called the Cosine Modulated Filter Bank (CMFB) [13]. Over the years, the CMFB technique evolved as an attractive solution for filter banks in DHAs because of its real-valued coefficients, polyphase implementation possibilities, ease of design and less delay.

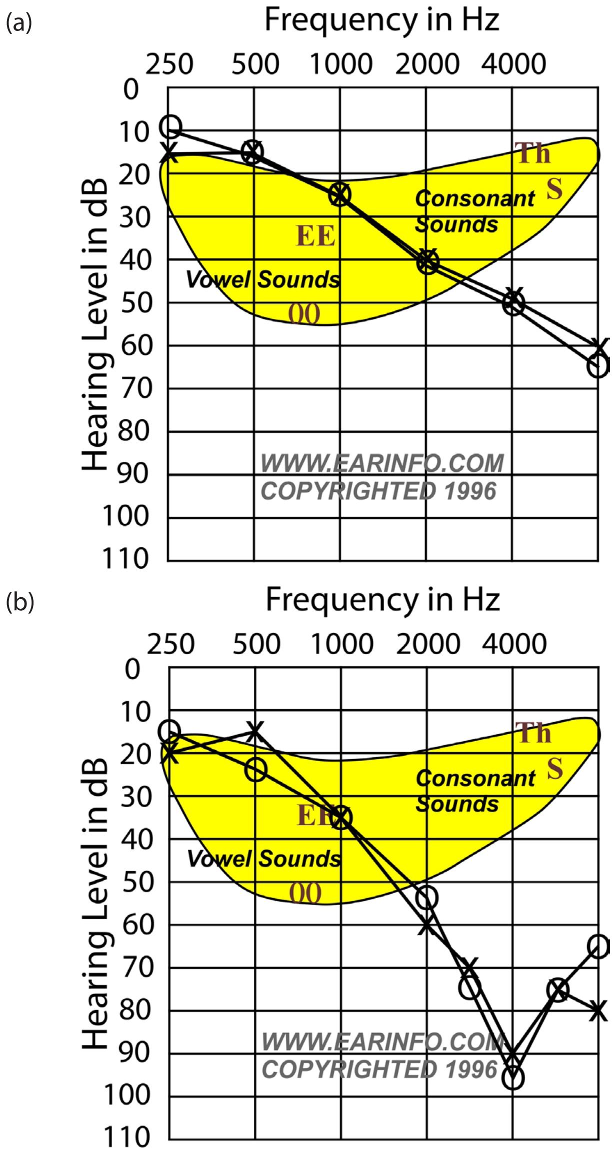

Figure 1: (a) SNHL Audiogram (b) NIHL Audiogram

Various works have been proposed in the literature to address low or moderately-sloping hearing Loss (Figure 1(a)) [3,11]. However, only a few works are available that address the notching NIHL audiograms (Figure 1(b)). There are two types of filter banks, namely fixed and reconfigurable filter banks. In the case of reconfigurable filter banks, the number of bands and band allocation frequencies can be flexibly adjusted based on the patient's audiogram. Fixed audiograms are generally used to correct hearing loss at low-frequency regions by allocating a greater number of bands. As NIHL is characterized by notches at center frequencies, flexible band allocation is essential. Thus, reconfigurable filter banks, which can correct the notched audiograms of NIHL by allocating a greater number of bands at center frequencies, are an important area of research.

Various studies show that [14] the Maximum Matching Error (MME) permissible for audiogram design is ±3dB, and it is required to allocate bands such that the MME is always within limits for the targeted audiograms. Otherwise, the user may feel uncomfortable at certain frequencies while using the DHAs due to unwanted amplification. An ideal filter bank must be able to provide tolerable MME while keeping the delay minimum [15]. The effectiveness and usability of DHAs also depend on the group delay of the device. The acceptable group delay for Closed Hearing Aid Fitting (CHAF) is up to 30 ms [16], and the same for Open Hearing Aid Fitting (OHAF) is 5 - 10 ms [17]. A higher Delay in OHAF devices results in a 'comb filtering effect,' and a higher delay in CHAF devices results in an 'occlusion effect.' The comb filtering effect is due to the superimposition of the actual and delayed sounds. Similarly, the occlusion effect is a perceived echo-like sound generated from one's voice. Even though studies claim that hearing-impaired people have higher group delay tolerance than ordinary people, a delay of more than 30 ms affects the audio-visual integration and results in 'lip reading.' Considering various studies [18], 10 ms group delay remains an unofficial standard among audiologists and hearing aid engineers [19].

This paper presents a reconfigurable CMFB architecture that provides a good trade-off between filtering complexity, implementational efficiency, design complexity, reconfigurability and group delay for DHAs targeting NIHL. The rest of the paper is organized as follows. Section 2 describes the conceptual and mathematical background of Cosine Modulated Filter Bank. Section 3 explores various articles from the literature addressing design approaches in reconfigurable filter banks for DHAs. The design and architecture of the proposed filter bank are detailed in Section 4. Results are presented in Section 5, followed by a thorough discussion in Section 6. Finally, a conclusion is drawn in Section 7.

2 Background

2.1 Choice of Cosine Modulated Filter Bank (CMFB)

CMFB falls under Modulated Filter Banks (MFB), which have many real-time applications in audio, video, transmultiplexers for communication, and biomedical signal processing [20]. MFBs are broadly classified as DFT filter banks and CMFBs. Modulating sequences are complex exponentials for the DFT filter bank and co-sinusoids for CMFB. The main attraction of CMFB compared to other filter bank approaches is that the signals can be decomposed into different bands using a single linear phase low pass Prototype filter. CMFB implementation is further classified as uniform filter banks (UFBs) and non-uniform filter banks (NUFBs) based on applications.

From a design perspective, CMFB can be classified as Perfect Reconstruction (PR) CMFB and Nearly Perfect Reconstruction (NPR) CMFB. NPR nature of the CMFB is due to its nonlinear phase property of sub-bands after modulation of the Prototype filters. Research findings show that PR is a desirable but not necessary condition for audio applications [21]. Hence, the NPR filter bank eases the complexity of implementation and provides more economical solutions for the implementation. Generally, the CMFB designs are based on FIR filters due to their guaranteed stability and reduced design complexity.

In most of the applications, uniform CMFB designs dominate. However, for applications like DHAs, it is required to use non-uniformly allocated reconfigurable band spacings. Non-uniform reconfigurable band allocation is preferable when the signal energy exhibits bandwidth-dependent distribution among frequency bands. Thus, it is important to explore the possibilities of designing non-uniform reconfigurable filter banks for DHAs using CMFB techniques. Only a few works [21-24] appear in this direction, especially as architectures for implementation in DHAs. For reconfigurable CMFB design, various approaches, namely, interpolation methods, merging bands methods, and transition band methods, can be adopted. This paper proposes the design of a non-uniform reconfigurable filter bank architecture using a merging method.

2.2 Mathematical background of CMFB

In a Modulated Filter Bank, let us assume that the filter coefficients $h_k(n)$ of k^{th} filter in the filter bank are generated through exponential modulation. ie,

$$h_k(n) = h_0(n) e^{j2\pi k/M} \tag{1}$$

where $h_0(n)$ is the filter coefficient of the Prototype filter. Now, it is required to design a new class of filter banks with real coefficients using the Cosine Modulation technique. This can be achieved by generating $2M$ complex filters using exponential modulation and linking the appropriate pairs of filter bands to create the sub-bands. Therefore, the prototype filter $P_0(z)$ is a low-pass filter with a cut-off frequency $\pi/2M$ as, shown in Figure 2.

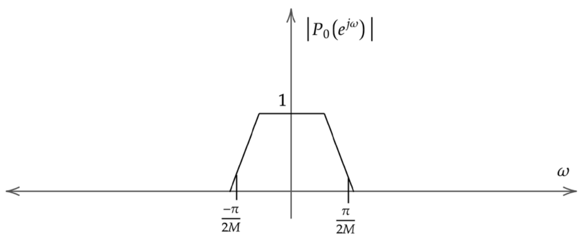

Figure 2: Prototype Filter in MFB

Now, the polyphase components of $P_0(z)$ are $P_k(z)$, $0 \leq k \leq 2M - 1$. Thus, equation (2) gives the frequency responses of the polyphase components.

$$P_k(e^{j\omega}) = P_0\left(e^{j(\omega - \frac{k\pi}{M})}\right) \tag{2}$$

Where $P_k(e^{j\omega})$ are $k\pi / M$ shifted versions of the Prototype filter response as shown in Figure 3.

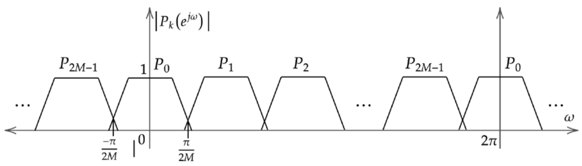

Figure 3: Magnitude Response of the Polyphase components in MFB

From Figure 3, it is evident that $P_k(e^{j\omega})$ and $P_{2M-k}(e^{j\omega})$ are images with respect to the zero frequency, so they can be combined to create the prototype filter with only real coefficients. When the images are combined using the method mentioned above, the pass band width of the combined filters will be $2\pi / M$, which is twice that of the Prototype Filter $P_0(z)$. Now, to make the filter bank suitable for amplitude compensation in DHAs, the initial prototype filter needs to be shifted by an amount of $2\pi / M$. Let us call the shifted version of the filters $Q_k(z)$. This is achieved by replacing z with $zW^{1/2}$ in the actual prototype filter $P_0(z)$ as given in Figure 4 and as per equation (3).

$$Q_k(z) = P_0(zW^{k+0.5}); 0 \leq k \leq 2M - 1 \tag{3}$$

In the case of a uniform Cosine Modulated filter bank, the bandwidth of all the M filters is equal to π / M. An equation that can be derived to compute the filter coefficients $h_k(n)$ of the Prototype filter $H_k(z)$ as given in equation (4).

$$h_k(n) = 2p_0(n)\cos\begin{bmatrix} \pi / M(k+0.5) \\ [n - N/2] + \theta_k \end{bmatrix} \tag{4}$$

Where, $\theta_k = (-1)^k \frac{\pi}{4}, 0 \leq k \leq M-1$ and N is the order of the Prototype filter. Figure 4 shows the magnitude response of an M channel non-uniform cosine modulated filter bank.

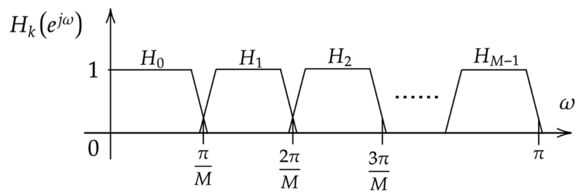

Figure 4: Magnitude Response of an M Channel Uniform CMFB

To match the logarithmic nature of the human auditory system, it is required to convert the above uniform CMFB filter bank to a Non-uniform CMFB filter bank for the actual implementation. The non-uniform filter bands are generated by merging the adjacent bands based on the narrow band and wide band requirements at different regions of the spectrum. Now, the transfer function of the resulting non-uniform CMFB filter is given by equation (5).

$$\tilde{H}_i(z) = \sum_{k=n_i}^{n_i + l_i - 1} H_k(z) \tag{5}$$

Where, $i = 0,1,2 \ldots \tilde{M} - 1$ and $\tilde{M}$ is the number of bands in the non-uniform CMFB filter bank. In equation 5, l_i is the number of adjacent bands to be merged and n_i is the corresponding upper band edge frequency of the i^{th} band. The magnitude response of an $\tilde{M}$ band non-uniform Cosine Modulated Filter Bank is given in Figure 5.

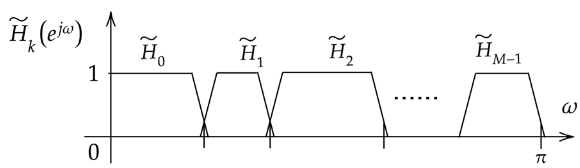

Figure 5: Magnitude Response of M Channel Non-uniform CMFB

3 Literature review

The Cosine Modulated Filter bank was originally derived from the Quadrature Mirror Filter Bank (QMF). The concept of QMF was later extended to create M Channel Filter Banks [25]. However, the exponential modulation in M Channel Filter banks results in complex filter coefficients. The CMFB is proposed to eliminate the complex filter coefficients [26]. Efficient polyphase implementation of uniform CMFB [27] using linear phase structure of the prototype filter has shown that the hardware savings are significant in such implementations when the ratio of the length of the filter to the number of filter bands is large. In Nearly Perfect Reconstruction (NPR) CMFB, the reconstruction criteria are relaxed to obtain efficient and faster polyphase implementations. The main objective of work related to NPR CMFB [28] is to design NPR CMFB exploiting the symmetry of the Linear Phase Prototype filter to reduce the number of multiplications in Polyphase implementation.

The design of non-uniform CMFB [10] from uniform filter banks by merging method gained the attention of researchers working in audio-related technology. It is shown [29,30] that such a strategy using interpolation techniques offers simpler designs and high stopband attenuation. Comparatively, a different approach [31] for designing non-uniform CMFB using a transition filter shows that non-uniform filter banks are designed from uniform filter banks by incorporating a transition filter to satisfy the aliasing cancellation conditions. However, individual filters need to be implemented for the non-uniform bands in this approach, which increases the computational complexity.

The first work reported [32] on the use of a CMFB filter bank for DHAs uses a uniform CMFB. In that work, it is shown how the typical design constraints for DHAs are achieved using appropriate filter bank types and efficient prototype filter design. Another important work [21] explaining the design of a non-uniform CMFB for DHA uses NPR non-uniform CMFB with a merging approach and transition filter approach. In this work, different FBs are designed for different audiograms and audiogram matching is performed. However, the architecture of the filter bank is not included in the work, and the design is non-reconfigurable for different audiograms. Hence, an efficient implementation strategy needs to be further explored for non-uniform Reconfigurable filter banks.

A reconfigurable filter bank design [22] based on CMFB published recently adopts a different approach compared to the above works. In this work, the prototype filter is Cosine Modulated, and then the sub-bands are generated by the nonlinear transformation of the uniform sub-bands. However, the reconfigurability is limited as only four different frequency band decomposition schemes are available. The number of multipliers required is higher than the other methods, and linear phase characteristics are affected due to the introduction of nonlinear transformation of the uniform sub-bands. The literature review of the various articles shows a gap in efficient implementation strategies for non-uniform reconfigurable NPR CMFBs. Also, the value of the filter bank increases if the non-uniform filter bank design strategies can be extended to a reconfigurable filter bank, which may be programmed externally by the audiologist for different audiograms.

4 Proposed non-uniform reconfigurable CMFB

4.1 Architecture of the proposed reconfigurable non-uniform filter bank

Figure 6 shows the architecture of the proposed Reconfigurable filter bank. The main objective of the proposed reconfigurable filter bank is to provide narrow bands and wide bands at different frequency regions of the audiogram. The slope of the audiogram primarily determines the band-splitting scheme at a particular region. The passband width is narrow in regions with steep slopes and wider in areas with low slopes in audiograms. In the proposed architecture, the coefficients of narrower bands are merged to make wider bands, reducing the total multiplications required to implement the filter bank. Figure 7 shows the band allocation in the proposed filter bank. Since the sampling frequency considered for the design of DHAs is 16kHz, in Figure 7 corresponds to 8kHz and $\pi/4$ corresponds to 2kHz. The Proposed filter bank is a sectional reconfigurable filter bank. Here, the entire frequency range, or 8000 Hz, of the audiogram is divided into four equal sections. The width of each section is equal to $\pi/4$ or 2000 Hz in Figure 7. The sectional reconfigurability provides six different band-splitting schemes for each section, which the audiologist can select based on control signals, as given in Table 1. In the figure, the default 500Hz bands are named as C bands, 1000 Hz as B bands and 2000 Hz as A Bands. The D bands are 1000 Hz bands generated by merging the second and third bands at each section.

Table 1: Band Splitting Scheme and Select line logic for a Section

Band Splitting Scheme	Select Line Logic $P_{3k-3}, P_{3k-2}, P_{3k-1}$	Sub band Distribution	Switches to be turned ON
1	000	$\frac{\pi}{16}, \frac{\pi}{16}, \frac{\pi}{16}, \frac{\pi}{16}$	$S_{12k-11}, S_{12k-10}, S_{12k-9}, S_{12k-8}$
2	001	$\frac{\pi}{8}, \frac{\pi}{8}$	$S_{12k-7}, S_{12k-6}, S_{12k-3}, S_{12k-2}$
3	010	$\frac{\pi}{4}$	$S_{12k-7}, S_{12k-6}, S_{12k-3}, S_{12k-2}, S_{12k-1}, S_{12k}$
4	011	$\frac{\pi}{16}, \frac{\pi}{8}, \frac{\pi}{16}$	$S_{12k-11}, S_{12k-5}, S_{12k-4}, S_{12k-3}$
5	100	$\frac{\pi}{8}, \frac{\pi}{16}, \frac{\pi}{16}$	$S_{12k-7}, S_{12k-6}, S_{12k-9}, S_{12k-8}$
6	101	$\frac{\pi}{16}, \frac{\pi}{16}, \frac{\pi}{8}$	$S_{12k-11}, S_{12k-10}, S_{12k-3}, S_{12k-2}$

The fundamental idea of the proposed filter bank architecture is that it uses a parallel structure of a differently modulated FIR filter. The major components of the filter bank are the polyphase substructures of the Prototype filter marked as $A_0 – A_{16}$, the modulator blocks marked as $M_0 – M_{16}$ and the adder network. Initially, the real-valued modulation coefficients of the filter bank are stored in memory. The prototype filter coefficients and the modulation coefficients are given to the Modulator circuit as per the control signals to generate the required bands for a section.

Each modulator consists of 16 multipliers, as shown in Figure 6. The output from the modulator blocks is added inside the adder network to generate the required non-uniform bands. The adder networks 1 to 16 correspond to the 16 bands present in the filter bank. The routing of the adder network ensures that the modulated prototype filter coefficients are correctly rearranged to generate the required bands for a section. Finally, the gain of each band is varied according to the audiogram. The overall response of the filter bank is obtained by adding all the gain-adjusted sub-bands of all the sections.

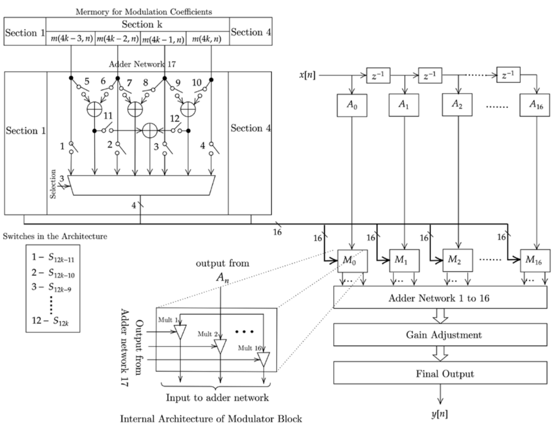

Figure 6: Architecture of the Proposed Non-Uniform Reconfigurable CMFB

The prototype filter of order 64 was first designed using the Blackman window method and implemented in a polyphase structure. This implementation requires only 32 multipliers because of the symmetry property of a Linear phase FIR filter. The choice of order 64 for the prototype filter is a trade-off between matching error and symmetry of the modulation coefficients. From the careful analysis carried out during the design phase, it is observed that maximum multiplier sharing is possible for the modulation coefficients when the order is 64, and the matching error for the targeted Audiograms is within the acceptable limits at order 64. Increasing the order does not significantly improve the matching error but increases the number of multipliers required for modulation coefficients. Decreasing the order of the filter affects the symmetry of the modulation coefficients and thus increases the number of multipliers and deteriorates the audiogram matching performance. The multipliers in the Modulator blocks may be dynamically turned ON or OFF based on the patient's Audiogram, which can reduce the power consumption of the Filter Bank. For example, if the patient has a relatively low slope audiogram, only a few multipliers will be activated in the modulator block, saving much power. Theoretically, the extreme and simplest band distribution consists of four bands and one band in each section, corresponding to 256 and 64 total multipliers for modulation. However, practical audiogram matching requires at least four bands in one or two sections, which requires almost 50% of the maximum multipliers to be turned ON. Through coefficient sharing and modulation coefficient symmetry, the total number of multipliers needed for the implementation of the proposed filter bank is 251.

The four sections in the proposed filter bank are reconfigured using three control signals each. Thus, the

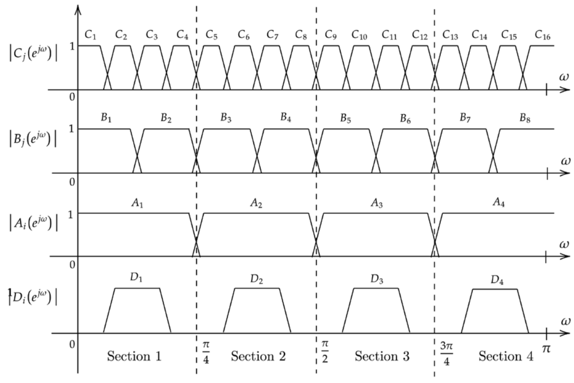

Figure 7: Different Sections and corresponding band Schemes in the Proposed Reconfigurable Filter Bank

entire filter bank can be reconfigured according to the characteristics of the audiogram using only 12 external control signals. Let P_{3k-3}, P_{3k-2}, P_{3k-1} be the selection lines of the k^{th} section. The combination of these select lines, according to the logic given in Table 2, is used to generate the activation signals F_{12k-11} – F_{12k} for the switches S_{12k-11} – S_{12k} that activate or deactivate the adders available in the Merging Circuit. Based on the switching control signals, the modulation coefficients of the proposed filter banks will be merged in the merging circuit to generate the bands for each section as per the schemes given in Table 1.

Table 2: Switch activation Logic for a Section in the Proposed Reconfigurable Filter Bank

Switch Activation Signal	Logic for the 3-bit Select Lines of the k^{th} section
F_{12k-11}	$\overline{P}_{3k-3}(P_{3k-2} \odot P_{3k-1}) + P_{3k-3}\overline{P}_{3k-2}P_{3k-1}$
F_{12k-10}	$\overline{P}_{3k-3}(P_{3k-2} \odot P_{3k-1})$
F_{12k-9}	$\overline{P}_{3k-2}\overline{P}_{3k-1}$
F_{12k-8}	$\overline{P}_{3k-3}P_{3k-2}P_{3k-1} + \overline{P}_{3k-2}\overline{P}_{3k-1}$
F_{12k-7}	$\overline{P}_{3k-3}(P_{3k-2} \oplus P_{3k-1}) + P_{3k-3}\overline{P}_{3k-2}\overline{P}_{3k-1}$
F_{12k-6}	$\overline{P}_{3k-3}(P_{3k-2} \oplus P_{3k-1}) + P_{3k-3}\overline{P}_{3k-2}\overline{P}_{3k-1}$
F_{12k-5}	$\overline{P}_{3k-3}P_{3k-2}P_{3k-1}$
F_{12k-4}	$\overline{P}_{3k-3}P_{3k-2}P_{3k-1}$
F_{12k-3}	$P_{3k-3}\overline{P}_{3k-2}P_{3k-1} + \overline{P}_{3k-3}P_{3k-2}$
F_{12k-2}	$P_{3k-3}\overline{P}_{3k-2}P_{3k-1} + \overline{P}_{3k-3}P_{3k-2}$
F_{12k-1}	$\overline{P}_{3k-3}P_{3k-2}\overline{P}_{3k-1}$
F_{12k}	$\overline{P}_{3k-3}P_{3k-2}\overline{P}_{3k-1}$

4.2 Design of the proposed non-uniform reconfigurable CMFB

Consider an M band uniform Cosine modulated Filter bank derived from a Prototype filter. The impulse response of k^{th} sub-band of the M band uniform Cosine Modulated Filter Bank is given by the following equation.

$$h_k(n) = 2p_0(n)\cos\left(\frac{\pi}{M}\left(n - \frac{N}{2}\right)(k - 0.5)\right) \quad ;1 \le k \le M \tag{6}$$

Where $p_0(n)$ is the impulse response of the prototype filter of order N. To design the uniform CMFB, the cut-off frequency of the prototype filter must be equal to $\pi/2M$. From the above equation, the impulse response of k^{th} band of the band uniform CMFB can be written as follows.

$$h_k(n) = 2p_0(n)\cos\left(\frac{\pi}{16}\left(n - \frac{N}{2}\right)(k - 0.5)\right) \tag{7}$$

In CMFB design, sub-bands of the filter bank are generated by modulating the Prototype filter coefficients with appropriate modulation coefficients. For example, in the above equation, the modulation coefficients of the 16-band uniform CMFB can be computed using the equation (8).

$$m(k,n) = 2\cos\left(\frac{\pi}{16}\left(n - \frac{N}{2}\right)(k - 0.5)\right) \tag{8}$$

These modulation coefficients are used to obtain the uniform bands of bandwidth 500Hz.

If the bands are implemented as direct filters modulated with the modulation coefficients, the computational resources required for the filter bank implementation will be enormous. Hence, in the proposed non-uniform reconfigurable filter bank, the prototype filter is implemented as the polyphase structure to exploit the redundancy in the modulation coefficients. From the modulation coefficient values for an even-order Prototype filter, the following patterns are observed for M = 16, and this redundancy helps to implement the filter bank with optimum computational resources.

In the proposed design, the modulation coefficients for a 16-band uniform filter bank are symmetric with respect to $n = \frac{N}{2}$, $n = \frac{N}{4}$ and $n = \frac{3N}{4}$. This symmetry can be mathematically represented as given in equations (9) and (10).

$$m(k,n) = m(k,N-n); 1 \le k \le M,\ 0 \le n \le N \tag{9}$$

$$m(k,n) = \begin{cases} (-1)^k\, m\left(k, \frac{N}{2} - n\right), & 0 \le n < \frac{N}{2} \\ (-1)^k\, m\left(k, \frac{3N}{2} - n\right), & \frac{N}{2} < n \le N \end{cases} \tag{10}$$

This pattern observed in the modulation coefficients of the filter bank helps us to reduce the multiplier count by implementing the sub-filters as polyphase structures as per equation (11).

$$A_g = \begin{cases} E_g\left(z^{\frac{N}{2}}\right) + z^N p_0(N) & ;g=0 \\ E_g\left(z^{\frac{N}{2}}\right) + z^{-\left(\frac{N}{2}-2g\right)} E_{\frac{N}{2}-g}\left(z^{\frac{N}{2}}\right) & ;1 \le g < N/4 \\ E_g\left(z^{\frac{N}{2}}\right) & ;g = N/4 \end{cases} \quad (11)$$

In equation (11), $E_g\left(z^{\frac{N}{2}}\right)$ represents the g[th] polyphase component of the prototype filter with $\frac{N}{2}$ stages.

Similarly, equations (12) and (13) give the polyphase components of the prototype filter.

$$E_g\left(z^{\frac{N}{2}}\right) = P_0(g) + z^{-\frac{N}{2}} P_0\left(g + \frac{N}{2}\right) \quad (12)$$

$$E_{\frac{N}{2}-g}\left(z^{\frac{N}{2}}\right) = P_0\left(\frac{N}{2} - g\right) + z^{-\frac{N}{2}} P_0(N-g) \quad (13)$$

Thus, the polyphase components for the actual implementation can be denoted using equation (14).

$$A_g = P_0(g) + z^{-\frac{N}{2}} P_0\left(g + \frac{N}{2}\right) + z^{-\left(\frac{N}{2}-2g\right)}\left(P_0\left(\frac{N}{2} - g\right) + z^{-\frac{N}{2}} P_0(N-g)\right) \quad (14)$$

where $P_o(g)$ is the g^{th} coefficient of the prototype filter.

Since FIR filter with even order and symmetric impulse response is used for implementation, the above equation can be modified as given in equation (15).

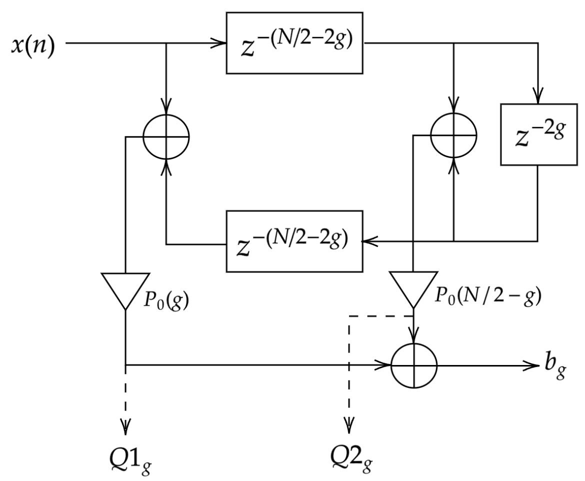

Figure 8: Implementation of the Sub filter component A_g

$$A_g = P_0(g)\left(1 + z^{-(N-2g)}\right) + z^{-\left(\frac{N}{2}-2g\right)} P_0\left(\frac{N}{2} - g\right)\left(1 + z^{-2g}\right) \quad (15)$$

The architectural implementation of the polyphase component of the sub-filter A_g is illustrated in Figure 8.

Similarly, a pattern as given in equation (16) is observed for the modulation coefficients corresponding to 16 bands in the proposed filter bank when arranged into columns with the same indices.

$$|m(k,r)| = |m(k,17-r)| \quad (16)$$

This pattern also helps in reducing the number of multipliers required for the filter bank implementation by adopting the multiplier-sharing scheme between the sub-bands. The number of multipliers inside each block depends on the number of unique modulator coefficients required to create different bands for that polyphase component A_g of the Prototype filter.

Considering all the factors mentioned above, the number of multipliers required to implement modulator coefficients of a 16-band uniform filter bank is 219. Because of the symmetry in the filter coefficients, the number of multipliers needed to implement the prototype filter is 32. In total, 251 multipliers are required to implement the proposed non-uniform reconfigurable filter bank.

4.3 Delay analysis of the proposed non-uniform reconfigurable CMFB

The group delay of the proposed non-uniform Cosine Modulated Filter Bank is given by Equation (17).

$$t = \left\lfloor \frac{N-1}{2} \right\rfloor \frac{1}{f_s} = \left\lfloor \frac{65-1}{2} \right\rfloor \frac{1}{16 \times 10^{-3}} = 2ms \quad (17)$$

Where N is the length of the Prototype filter and f_s is the sampling frequency. The delay of the CMFB filter bank is much less compared to other types of filter banks. This is because the entire filter bank can be implemented using a single Prototype filter and efficiently implemented using the Polyphase structure by exploiting the redundancy in the modulation coefficients to shift the bands.

5 Experimental Results

The simulation results for the proposed Reconfigurable CMFB are shown in Figure 9. Unlike fixed filter banks, Reconfigurable CMFB must provide bands with different resolutions and band combinations for a particular region of the audiogram. That means sub-bands with different bandwidths are available throughout the frequency region for frequency compensation, and the audiologist may be able to merge or unmerge bands based on the characteristics of the audiogram. Figure 9 shows the different band resolutions available at different regions of the audiogram. Suppose an audiologist needs a narrow band in the middle region alone to compensate for the frequency notch; they can provide narrow bands in that region using external control signals for Section 2 or Section 3. Similarly, a steep slope can be compensated at high or low-frequency regions using narrow bands in Sections 1 and 4. A bands are enough for frequency compensation when the audiogram is almost linear with a lesser slope at all the frequency regions. Only six bands are required for frequency compensation of low or moderately-sloping audiograms, which reduces the computational resources used in the filter bank for a particular audiogram. The audiogram matching curve and audiogram matching error depicted in Figure 10 show that the proposed Reconfigurable filter bank can accommodate all kinds of audiograms, including low or moderately-sloping SNHL audiograms and NIHL audiograms. The audiogram matching performance of the proposed reconfigurable filter bank for Standard IEC 60118-15(IEC2012) audiograms and the FPGA synthesis results are given in Table 3 and Table 4, respectively. The Proposed architecture is synthesized using Zynq FPGA kits to evaluate parameters like computational resource utilization in terms of LUTs and delay in terms of logic delay. The proposed design is first converted to synthesizable Verilog code using the MATLAB HDL coder and synthesized using FPGA. The results of the proposed Reconfigurable CMFB are compared with a fixed non-uniform CMFB targeting NIHL. The results show that incorporating reconfigurability improves audiogram matching flexibility but comes with increases in computational overhead and logic delay compared to fixed CMFB.

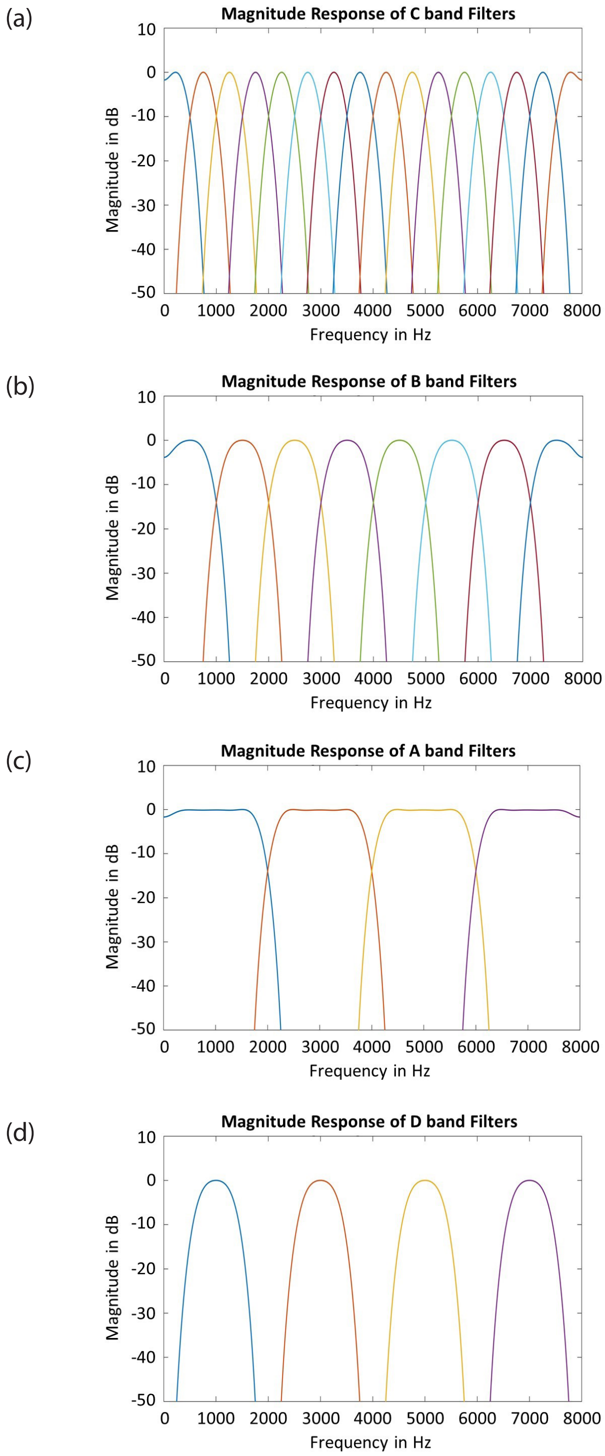

Figure 9: (a) C (b) B (c) A (d) D bands of the proposed non-uniform reconfigurable CMFB

6 Performance comparison and discussion

A detailed comparison of the Proposed Non-uniform Reconfigurable CMFB with other state-of-the-art filter banks for DHA is given in Table 5. The comparison is made with FRM [8,33-35], Farrow [36,37] and CMFB-based filter banks [21,22]. From the comparison results, it is observable that the matching performance of the Proposed filter banks is satisfactory for different kinds of real and standard audiograms. The proposed reconfigurable CMFB can flexibly match both SNHL and NIHL audiogram within the tolerable limits. The delay of the proposed filter banks is much less, making them suitable candidates for sophisticated DHAs, which provides

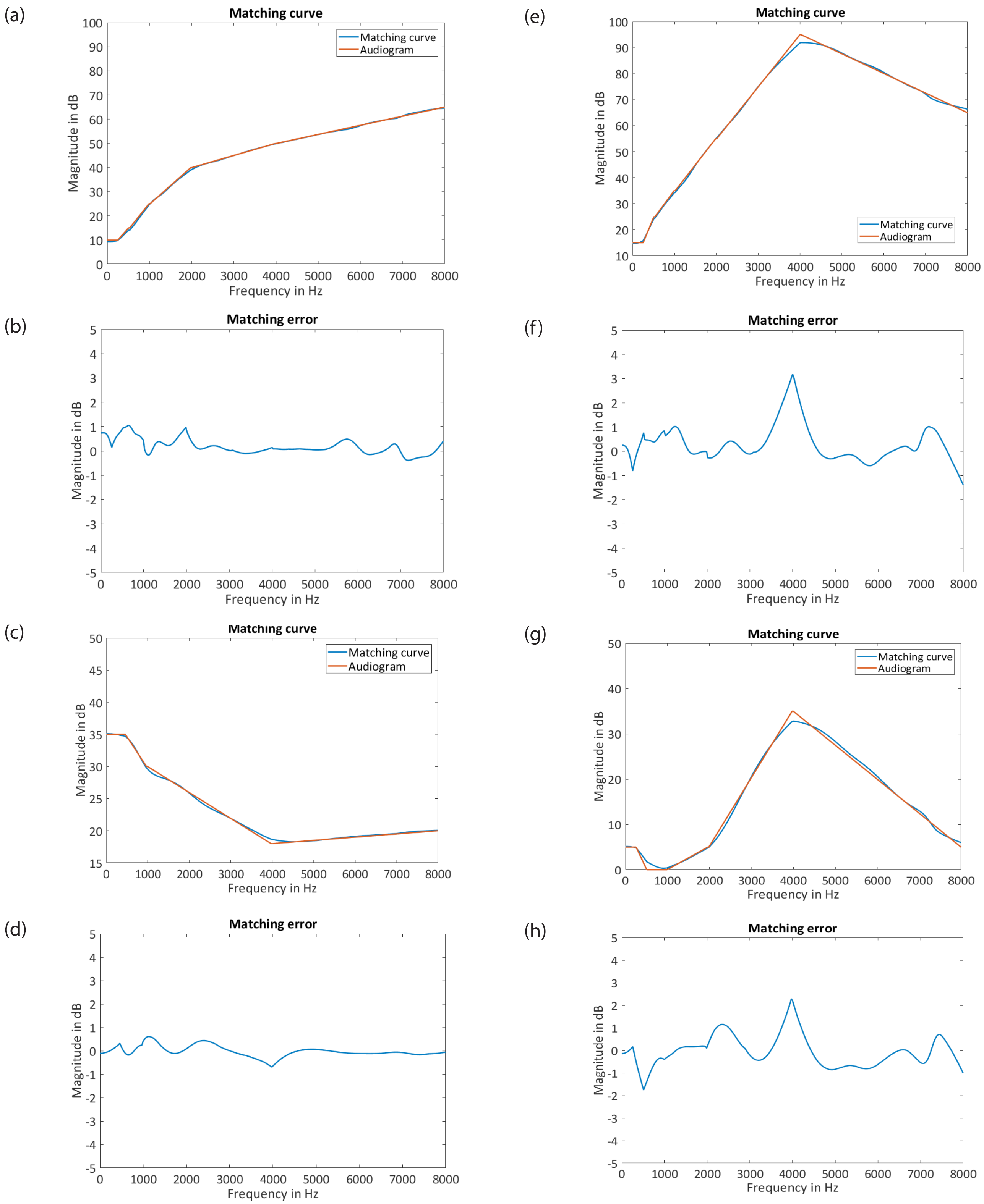

Figure 10: Matching performance of non-uniform and reconfigurable CMFB (a) & (b) Type 1 SNHL, (c) & (d) Type 2 SNHL, (e) & (f) Type 1 NIHL, (g) & (h) Type 2 NIHL

room for incorporating other complex DSP algorithms. The Computational resources required to implement the proposed filter banks are slightly higher than FRM methods and comparable with CMFB and Farrow methods. When architectural advantages and matching performance of the proposed filter bank are considered, it is arguable that the slight increase in the computational resources is reasonable for practical implementation.

The existing fixed CMFB design [21] presents the merging method and transition filter methods. This work does not discuss the actual implementation of the filter bank, and it proposes the design of the filter bank by direct modulation of the individual bands. Also, this approach does not provide a common filter bank structure for different kinds of audiograms; instead, the design of a filter bank with different number of bands is proposed for different audiograms. This is not a practical approach, as different filter banks cannot be realized for different classes of audiograms. The matching performance of the existing work is presented with an arbitrary number of bands for different audiograms. However, the proposed filter banks offer a practically feasible implementation approach.

Table 3: Audiogram Matching performance of the proposed reconfigurable CMFB for standard IEC 60118-15(IEC2012) audiograms

Class of the Audiogram	Name of the Standard Audiogram	Maximum Matching Error (MME) in dB	
		Fixed Non-Uniform CMFB	Reconfigurable CMFB
Flat or Moderately Sloping Standard Audiograms	Very Mild	0.45	0.49
	Mild	0.62	0.67
	Moderate	0.59	0.64
	Moderate/ Severe	0.94	0.71
	Severe 1	2.31	0.93
	Severe 2	2.23	0.97
	Profound	2.67	1.28
Steep Sloping Standard Audiogram	Very Mild	3.21	2.38
	Mild	3.94	2.85
	Moderate/ Severe	4.23	2.73

Table 4: FPGA Synthesis results of the proposed CMFB filter banks

Architecture	Zynq Evaluation and Development (ZED) FPGA (XC7Z020CLG484-1)				
	LUTs	IOBs	Total Delay (ns)	Logic Delay (ns)	Net Delay (ns)
Non-uniform CMFB for NIHL	21756	208	279	262	17
Proposed Reconfigurable CMFB	36987	282	454	420	34

The existing reconfigurable CMFB approach [22] uses an indirect design approach. The Prototype filter is cosine modulated to generate uniform sub-bands, and these sub-bands undergo non-linear transformation to produce the required bands. Only four different band allocation schemes are available in the filter bank, which makes bands narrower in some regions and wider in other regions at the same time; hence, there is a dependency between different bands. The MME is beyond the tolerable limit for notched NIHL audiograms. Compared to this method, the proposed reconfigurable CMFB provides sectional independent reconfigurability in which the bands at one region do not affect the other region. For example, a narrow band can be simultaneously allocated for low and high-frequency regions, which is not available in the existing work. Also, the proposed Reconfigurable CMFB outperforms the existing Reconfigurable filter bank in matching performance.

When the matching performance of the proposed filter banks is compared with non-CMFB approaches like FRM and Farrow, the matching performance is comparable and within the tolerable limit. However, the delay performance of the proposed filter bank is the best among these methods. The delay of the existing CMFB reconfigurable filter bank [22] is higher than that of the proposed method as it uses the non-linear transformation architecture in the final stage of the filter bank. The only disadvantage of the proposed method compared to both the fixed and reconfigurable FRM approach is the slight increase in the computational resources. However, the delay performance, reconfigurability through external control signals and tolerable MME for all classes of audiograms, make it an attractive candidate for real-time implementation in modern DHAs.

The efficiency of the Proposed filter banks lies in the following aspects. Polyphase implementation of the reconfigurable CMFB keeps the computational resources within acceptable limits. Design of the entire filter bank using a Single Prototype filter reduces the design complexity. The efficient architecture of the modulator structure by exploiting the coefficient redundancy makes the filter bank realizable for practical DHA applications. The dynamic activation and deactivation of the modulator multipliers in the reconfigurable filter bank based on audiograms reduces the power consumption of the filter bank. Sectional reconfigurability provided to the reconfigurable filter bank enables tuning of the DHAs through the control signals independent of the frequency region. Merging the adjacent bands to generate wider bands reduces computational complexity.

7 Conclusions

This paper presents the architecture and design of a 16 bands sectional non-uniform reconfigurable filter bank based on the Cosine Modulation technique. The proposed reconfigurable CMFB is obtained by merging the modulation coefficients of 16 band uniform CMFB. Polyphase implementation of the Prototype filter and grouping of modulation coefficients reduced the computational resources required for the real-time implementation. The group delay of the filter bank is only 2ms, which enables it to be used in DHAs, which requires delay relaxation for the implementation of other advanced DSP algorithms. About 1296 different schemes may be obtained using this structure, making it an attractive solution for matching all classes of audiograms. It uses the merging method to generate reconfigurable sub-bands. Initially, the modulation coefficients are stored in a memory. The merging of modulation coefficients is carried out based on the external select line and is fed to multiplier blocks to generate the required bands. This helps the audiologist select wider or narrower bands for each region of the audiogram to reduce the matching error while performing frequency compensation. The unwanted multipliers may be disconnected in the modulator block to save power.

A maximum resolution of 500Hz can be obtained using this method. The number of multipliers required to implement this filter bank is 251, which is 34% higher than the existing reconfigurable CMFB approach. However, this method provides better matching performance for all classes of audiograms and has a very low group delay of 2 ms.

8 Conflict of Interest

The authors declare no conflicts of interest.

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Table 5: Audiogram Matching Result Comparison of Various Filter Bank Techniques with Proposed NUCMFB and Reconfigurable CMFB

Filter bank Design Method	CMFB Based	Maximum Matching Error (MME) for Different Audiograms (in dB)								Computational Resource		Delay (ms)
		Type 1	Type 2	Type 3	Type 4	Type 5	Type 6	Type 7	Type 8	#Multipliers	#Adders	fs=16kHz
FRM 8 band (Optimized) [8]	N	0.83	2.11	14.13	9.64					18	36	12.8
Reconfigurable FRM 2 [33]	N	-	-	-	-	5.63	1.84	-	-	76	170	12.1
Reconfigurable FRM 4 [34]	N	1.49	2.54	2.72	-	-	-	1.49	-	84	196	18.75
Reconfigurable Farrow 1 [36]	N	2.45	1.67	-	-	-	2.11	-	-	216	432	1.1
Reconfigurable Farrow 2 [37]	N	2.60	1.51	2.96	-	2.0	1.61	-	-	138	276	1.3
FRM 16 band [35]	N	0.42	0.27	4.33	1.63	2.5	0.59	0.49	1.17	33	64	6
Fixed Non-Uniform CMFB (Merging) [21]	Y	2.49	2.19	-	-	-	-	-	-	192	384	2
Fixed Non-Uniform CMFB (Transition) [21]	Y	1.9	2.12	-	2.23	2.3	-	2.7	2.4	147	142	2
Reconfigurable CMFB [22]	Y	-	1.65	-	-	3.75	1.35	-	-	187	156	7.77
Reconfigurable CMFB (Proposed)	Y	1.95	0.69	3.18	2.85	2.57	1.95	2.28	1.88	251	274	2

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