

Design of the Audio Coding Standards for MPEG and AC-3

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ABSTRACT

ISO MPEG 1/2 and Dolby AC-3 are widely used in the network, wireless, multimedia system and video industry. This dissertation studies the design of audio standards: MPEG-1/2 and AC-3.

The perceptual audio coder like MPEG-1/2 and AC-3 can be analyzed through filterbank, psychoacoustic model, stereo matrix, bit allocation/quantization, and packing block. This dissertation considers the design for the filterbank, psychoacoustic model, stereo matrix, and bit allocation/quantization. This dissertation summarizes the filterbanks adopted in coding standards and presents a unified fast algorithm for these filter banks. On the psychoacoustic models, the hybrid filterbank is proposed to replace to original frequency analyzer for MPEG audio standards to have efficient computing. On the bit allocation, we analyze the issues in bit allocation and present the efficient method. This dissertation also studies the stereo irrelevancy and presents the new method to achieve good quality.

Keywords: MPEG, AC-3, Audio coding, Filterbank, Bit allocation, Intensity/coupling coding, Layer 3.

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Chapter 1

Introduction

During the last decade, analog audio has been wholly replaced by the CD-quality digital audio. The demand for digital audio compression with constraint bandwidth, limit storage is rapidly increased for the network, wireless, multimedia system and video industry. In response to this need, considerable researches for the perceptually transparent coding of high-fidelity (CD-quality) digital audio have been developed. Several algorithms have now become international standards or commercial products. ISO MPEG-1/2 layer 1/2/3 and Dolby AC-3 are the most widely adopted among the standards such as- HDTV, DVD, VCD, and Internet audio.

MPEG-1 [24] comprises a flexible hybrid coding technique that incorporates several methods including subband decomposition, filterbank analysis, transform coding, entropy coding, dynamic bit allocation, non-uniform quantization, adaptive filterbank, and psychoacoustic analysis. MPEG coders accept 16-bit PCM input data at sample rates of 32, 44.1, and 48 kHz. MPEG-1 offers separate modes for mono, stereo, dual independent mono, and joint stereo. Available bit rates are 32-192 kb/s for mono and 64-384 kb/s for stereo.

The MPEG layer 3 achieves quality improvements by adding several important mechanisms on the foundation of the layer 1/2. A hybrid filterbank is

introduced to increase frequency resolution and thereby better approximate critical band behavior. The hybrid filterbank includes adaptive filterbank to improve pre-echo control. Sophisticated bit allocation and quantization strategies that rely upon non-uniform quantization, analysis-by-synthesis, and entropy coding are introduced to allow reduced bit rates and improved quality. First, a hybrid filterbank is constructed by following each subband filter with an adaptive MDCT. This practice allows for higher frequency resolution and pre-echo control. Use of an 18- point MDCT, for example, improves frequency resolution to 41.67 Hz per spectral line. Adaptive MDCT block sizes between 6 and 18 points allow improved pre-echo control. Using shorter blocks during rapid attacks in the input sequence allows pre-masking to hide pre-echoes, while using longer blocks during steady-state periods reduces side information and hence bit rates. Bit allocation and quantization of the spectral lines is realized in a nested loop procedure that uses both non-uniform quantization and Huffman coding. The inner loop adjusts the non-uniform quantizer step sizes for each block until the number of bits required to encode the transform components falls within the bit budget. The outer loop evaluates the quality of the coded signal (analysis-by-synthesis) in terms of quantization noise relative to the JND thresholds.

MPEG-2 [23] extends the capabilities offered by MPEG-1 to support the so called 3/2 channel format with left, right, center, and left and right surround channels. The first MPEG-2 standard is backward compatible with MPEG-1 in the sense that 3/2 channel information transmitted by an MPEG-2 encoder can be correctly decoded for 2-channel presentation by an MPEG-1 receiver. The

second MPEG-2 standard sacrifices backwards MPEG-1 compatibility to eliminate quantization noise unmasking artifacts that are potentially introduced by the forced backward compatibility.

Algorithm	Transform	Channels	Applications	References
MPEG-1 layer 1/2	Subband	1, 2	VCD, DVB	[24]
MPEG-1 layer 3	Hybrid	1, 2	MP3, Network	[24]
MPEG-2 layer 1-3	Hybrid	1-5.1	MP3, Network	[23]
MPEG-2 AAC	Subband/hybrid	1-48	Network, HDTV	[25], [31]
Dolby AC-3	Transform	1-5.1	DVD, HDTV	[43], [27]

Table 1.1 Audio coding standards and applications.

AC-3 perceptual audio coder [43], [27] is developed for the 320 kb/s for High-Definition Television (HDTV) standard and also widely adopted in DVD film. AC-3 carries 5.1 channels of audio (left, center, right, left surround, right surround, and a subwoofer), but it has also been designed for compatibility with conventional mono, stereo, and matrixed multi-channel sound systems. A modified Discrete Cosine Transform (MDCT) filterbank is used to decompose audio signal. Transform spectrums are quantized using a psychoacoustically derived dynamic bit allocation scheme. Spectral information obtained from the MDCT is encoded using a novel mantissa/ exponent coding scheme. First, the spectral stability is evaluated. All transform coefficients are transmitted for stable spectra, but time updates occur only every 32 ms. Fewer components are encoded for transient signals, but time updates occur frequently, e.g., every 5.3 ms. A spectral envelope is formed from exponents corresponding to log spectral line magnitudes. These exponents are differentially encoded. Psychoacoustic

quantization masking thresholds are derived from the decoded spectral envelope for 64 non-uniform subbands that increase in size proportional to the ear's critical bands. The thresholds are used to select appropriate quantizers for transform coefficient mantissas in a bit allocation loop. If too few bits are available, high-frequency coupling (above 2 kHz) between channels may be used to reduce the amount of transmitted information. Exponents, mantissas, coupling data, and exponent strategy data are combined and transmitted.

As shown in Fig. 1.1, a perceptual audio coder is composed of filterbank, psychoacoustic model, stereo matrix, bit allocation/quantization, and packing block. The filterbank splits the input signals into subbands. Stereo matrix reduces the stereo irrelevancy. Then, samples in the subbands are quantized and coded under the control of a psychoacoustic model. This dissertation considers the design of these blocks as follows. Chapter 2 summarizes the filterbanks adopted in all these coding standards and presents a new unified fast computing algorithm for these filterbanks with variant forms and sizes. The unified algorithm reduces the development period for variant filterbanks and gives a guideline for developing new filterbanks. Chapter 3 presents a hybrid filterbank approach for the psychoacoustic models in MPEG audio standards to replace the original Fourier transform for efficient computing. Chapter 4 analyzes the issues in bit allocation and present the efficient bit allocation method for MPEG layer 3 and Dolby AC-3. For MPEG layer 3, the non-uniform quantizer and variant length coding make the developing efficient bit allocation more difficult. A noise predictor for the non-uniform quantizer for layer 3 is developed and one iteration bit allocation using the noise predictor is presented. For Dolby AC-3, it

adapts its range according to the specified exponent strategy. These strategies affect the temporal resolution and the spectral resolution of the quantization ranges. These encoded exponents also affect the analysis result of the psychoacoustic model. The exponents and the resultant psychoacoustic results determine the quantization results and thus has led to high complexity. This dissertation present the criteria to decide the strategies for the exponent coding and psychoacoustic model parameter and propose a efficient bit allocation algorithm for AC-3. Chapter 5 studies the stereo irrelevancy and presents the design method. KL (Karhunen-Loève) transform is introduced to design and analyze the intensity/coupling schemes to reduce stereo irrelevancy. With integrating the KL transform into intensity coding/coupling schemes of MPEG and AC-3, this dissertation presents and compares the algorithms to improve quality. Chapter 6 concludes the dissertation.

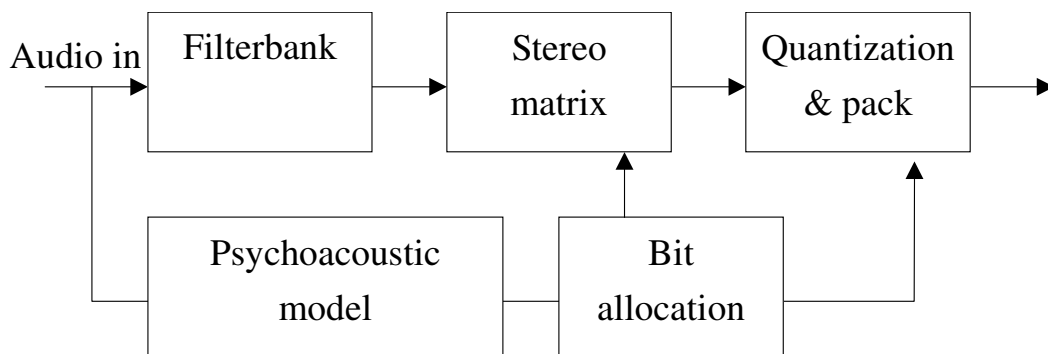


Fig. 1.1 Block diagram for perceptual audio coder.

Chapter 2

Unified Algorithm for Fast Filterbank Computing

Current audio coding standards such as MPEG-1 layers 1-3, MPEG-2 layers 1-4, MPEG-4, and AC-3, have adopted for compression various forms of the filterbank (CMFBs). This chapter demonstrates that all these MCTs can be derived into two modules: the permutation and the discrete cosine transform. The derived DCTs are either type-II, type-III, or type IV. On the three types of the DCT, this chapter proposes a fast computing algorithm to uniformly compute all the three types of the DCTs. The new fast algorithm has good features in regularity, complexity, and general applicability.

2.1 Introduction

In current audio coding standards such as MPEG-1 layers 1-3, MPEG-2 layers 1-4, MPEG-4, and AC-3, the cosine-modulated filterbanks (CMFBs) [41] have been widely adopted to transform an audio sequence from time-domain to transform domain or subband domain for compression. However, all the CMFBs' formulae vary with not only the standards but also with the standard layers, block length, encoding, and decoding process. For real-time applications,

these various formulae need to be individually designed and tuned for precision, complexity, and memory movements. This chapter will develop the unified fast algorithm for these formulae.

As shown in Fig. 2.1, the process of CMFBs can be considered from two steps: the window-and-overlapping addition (WOA) and the modulated cosine transform (MCT). The WOA performs a windowing multiplication and addition with overlapping audio blocks. The complexity of this step is $O(k)$ for an audio sample, where k depends on the overlapping factors of the forms. For example, the factor k is 16 for the MPEG-1 layer 2 and 2 for the AC-3. The second step, MCT, has a complexity $O(W)$ per audio sample, where W is the windowing length and is quite different for various CMFBs. The range of W is from 36 for MPEG-1 layer 3 to 4096 for the MPEG-4. For WOA, direct implementation has been generally adopted and the design is straightforward. On the contrary, the complexity of the MCT is high, and fast algorithms have been developed based on the similar concepts developed for the fast Fourier transform. It has been widely known that developing fast algorithms like the fast Fourier transform and the fast cosine transform needs to consider the tradeoff between arithmetic complexity, structure regularity, modularity, and numerical precision. Hence, it is always a critical issue for designing hardware or software for the fast MCTs.

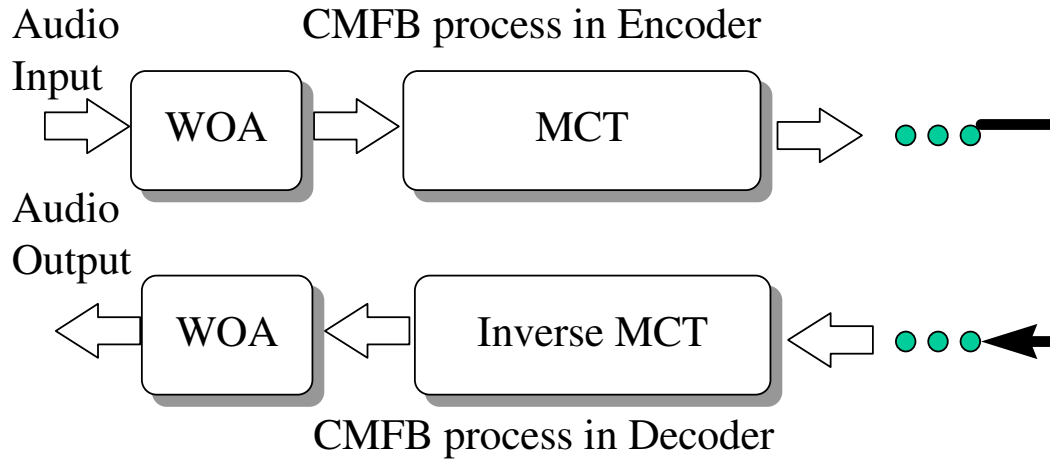


Fig. 2.1 The cosine-modulated filterbanks in the audio encoder and the decoder.

As illustrated in Fig. 2.1, this section demonstrates that all the various MCTs can be derived into two modules: the pre- (or post-) permutation and the discrete cosine transform (DCT). The DCT derived from the MCTs can be one of the three types of DCTs generally referred to as type-II, type-III, and type-IV [34]. On the results, this chapter further develops a fast algorithm which recursively decomposes a type of DCT with length N into other types of DCTs with length $N/2$. Recursive decomposition is the vehicle adopted in developing fast algorithms for sinusoidal transforms such as the discrete Fourier transform and the discrete cosine transforms. However, the main difference of the recursive decomposition is the decomposition of one type of DCT into type-II, type-III, or type-IV. The difference leads to two important benefits. First, the approach has a data regularity that is a property of the fast Fourier transform but not for the fast cosine transform. The regularity is important for the data path design in VLSI chip design [5], [22] and the memory addressing in software programming. The second merit is that the fast algorithm can be optimally implemented for all the MCTs in audio standards. Since this algorithm

recursively and regularly decomposes the long length transforms into short length ones through three types of the DCTs, the unrolling of the recursive decomposition from length N into length 2 will be the interleaving of the three types of the DCTs. In other word, the fast algorithm is applicable to all the three types of the DCT, and the computing vehicle for the three DCT types is the same. Hence, this section demonstrates that all the various CMFBs in the audio coding standards lead to different pre-permutation or post-permutation but will have the same computing vehicle for the DCTs. Through the same computing vehicle, the software modules or hardware modules can be generally developed for all these audio compression standards.

There have been many fast computing algorithms developed for the DCT. These algorithms are developed for different transform length and different DCT types. On the audio coding, radix-2 DCT is the main considering length. The development of the radix-2 fast DCT algorithms can be classified into two approaches: (1) the indirect computation of the DCT through the fast Fourier transform or the fast Hartley transform, and (2) the direct computation of the DCT through matrix factorization or recursive decomposition. The first approach needs additional complexity in mapping DCTs into another transform while the second approach in general lacks the modularity and data regularity. As mentioned by Yun [20], the modularity and the regularity are essential for designing hardware and generalizing to higher order transforms. Recently, Kok [16] has developed a fast algorithm for type-II DCT that can recursively decomposes one type-II DCT with length N into two type-II DCTs with length $(N/2)$. The decomposition from one DCT into two DCTs leads to the merit in

modularity and regularity. This section adopts the direct computation approach to achieve low complexity. The complexity analysis shows that the new algorithm can have a complexity matching with the well-known DCT algorithm [16][2][45][37]. Furthermore, we develop the decomposition through the interleaving of three types of DCTs instead of the same type of DCT to improve the regularity and the modularity. Since the decomposition is the interleaving of the three types of the DCTs, the fast algorithm is applicable to all three types of the DCTs instead of just the type II in [16]. The general applicability is the key factor to develop the fast algorithm for the cosine-modulated filterbanks (CMFBs) in the current audio standards.

The rest of this chapter is organized as follows: Section 2.2 illustrates that all the CMFBs can be derived into permutation and the discrete cosine transform. Section 2.3 demonstrates the decomposition of one type of the DCT into the interleaving of the other three DCT types to achieve fast computing. Section 2.4 gives concluding remarks.

2.2 Unified Form for the CMFBS

The modulated cosine transforms (MCTs) used in current audio standards can be classified into three types of filterbanks: the time-domain aliasing cancellation (TDAC) filterbank [30], the variant of the TDAC filterbank [43], and the polyphase filterbank [24]. Table 2.1 illustrates the formulae of the three classes of the cosine-modulated filterbanks (CMFBs) and the correspondence with various audio coding standards. This section demonstrates that all the

CMFBs can be represented as the pre- or post-permutation and the discrete cosine transform (DCT) as shown in Fig. 2.2. The DCT type can be one of the following three types:

Type-II DCT

$$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(2i+1)(k)\right) \text{ for } k = 0, 1, \dots, N-1. \quad (2.1)$$

Type-III DCT

$$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(i)(2k+1)\right) \text{ for } k = 0, 1, \dots, N-1. \quad (2.2)$$

Type-IV DCT

$$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{4N}(2i+1)(2k+1)\right) \text{ for } k = 0, 1, \dots, N-1. \quad (2.3)$$

In equations (2.1)-(2.3), there have been constant terms in front of each formula.

For example the type-IV DCT is

$$X_k = \sqrt{\frac{2}{N}} \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{4N}(2i+1)(2k+1)\right) \text{ for } k=0, 1, \dots, N-1.$$

The constant term $\sqrt{\frac{2}{N}}$ is neglected for ease of description.

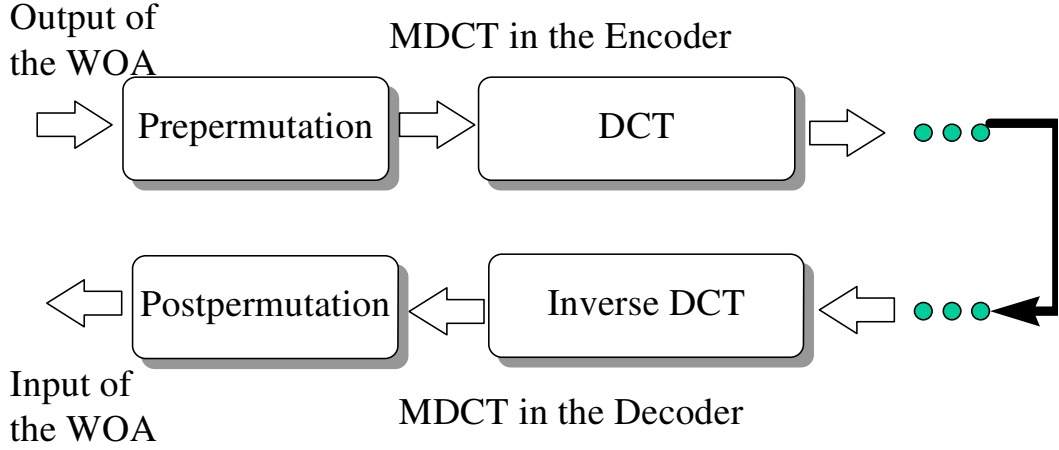


Fig. 2.2 The representation of the MDCT into permutation and the DCT.

2.2.1 Unified form for the MCT in TDAC filterbank

This section illustrates the method to transform the modulated cosine transform (MCT) in time-domain aliasing cancellation (TDAC) filterbank into the permutation and the type-IV DCT. The forward and inverse MCT of the TDAC filterbank are respectively defined as

$$X_k = \sum_{i=0}^{N/2-1} x_i \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(2k+1)\right) \quad \text{for } k=0, 1, \dots, N/2-1. \quad (2.4)$$

and

$$\tilde{x}_i = \sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(2k+1)\right) \quad \text{for } i=0, 1, \dots, N-1. \quad (2.5)$$

where a constant term before each summation is again neglected for representation ease. Also note that, unlike the general transform, the sequence \tilde{x}_i , in (2.5) is in general not equal to sequence x_i in (2.4) given the same X_k .

In the following we proceed with the derivation through three steps. First, we

extend the transform pair in (2.4) and (2.5) to a form which has length N along both indices i and k through **Theorem 1** and **Theorem 2**. Second, the extended transform with length N is represented as a length N transform which is quite similar to the type-IV DCT as illustrated in **Theorem 3** and **Theorem 4**. Finally, the DCT-like transform with length N is reduced to type-IV DCT with length (N/2) with input or output permutation through **Theorem 5** and **Theorem 6**.

Define the following transform pair:

$$X'_k = \sum_{i=0}^{N-1} x'_i \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(2k+1)\right), \text{ for } k=0, 1, \dots, N-1 \quad (2.6)$$

and

$$\tilde{x}'_i = \frac{1}{2} \sum_{k=0}^{N-1} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(2k+1)\right), \text{ for } i=0, 1, \dots, N-1. \quad (2.7)$$

The following two theorems illustrate the relation between the extended transform and the TDAC transform.

Theorem 1: The sequence X'_k in (2.6) is anti-symmetric in the sense that

$$X'_k = -X'_{N-1-k} \quad \text{if } N \text{ is a multiple of } 4.$$

<proof>: Representing X'_{N-1-k} as

$$X'_{N-1-k} = \sum_{i=0}^{N-1} x'_i \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(2(N-1-k)+1)\right) \quad \text{for } k=0 \dots N-1$$

which can be reformulated as

$$X'_{N-1-k} = \sum_{i=0}^{N-1} x'_i \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(-1-2k) + \pi \left(2i+1 + \frac{N}{2}\right)\right) \quad \text{for } k=0 \dots N-1$$

Since the transform length N is a multiple of four,

$$\begin{aligned}
X'_{N-1-k} &= -\sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(-1-2k)\right) \\
&= -\sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(-1-2k)\right) = -X'_k
\end{aligned}$$

Theorem 2: Let N be an integer with the multiple of four. Assume that the sequence X'_k with length N is obtained by extending the sequence X_k with length (N/2) according to $X'_k = -X'_{N-1-k}$ for $k=N/2, \dots, N-1$. Given (2.5) and (2.7) the sequence \tilde{x}_i computed from (2.5) is equivalent to the sequence \tilde{x}'_i computed from (2.7).

<Proof>: From (2.7),

$$\tilde{x}'_i = \frac{1}{2} \sum_{k=0}^{N-1} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(2k+1)\right) \tilde{x}'_i$$

Separating the summation into two parts yields

$$\tilde{x}'_i = \frac{1}{2} \left\{ \sum_{k=0}^{\frac{N}{2}-1} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(2k+1)\right) + \sum_{k=\frac{N}{2}}^{N-1} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(2k+1)\right) \right\}$$

Replacing the index k in the second summation as N-1-k yields

$$\tilde{x}'_i = \frac{1}{2} \left\{ \sum_{k=0}^{\frac{N}{2}-1} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(2k+1)\right) + \sum_{k=0}^{\frac{N}{2}-1} X'_{N-1-k} \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(2(N-1-k)+1)\right) \right\}$$

Since $X'_k = -X'_{N-1-k}$ and N is a multiple of four, the formula can be further rewritten as

$$\begin{aligned}\tilde{x}'_i &= \frac{1}{2} \left\{ \sum_{k=0}^{\frac{N-1}{2}} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(2k+1)\right) - \sum_{k=0}^{\frac{N-1}{2}} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(2(N-1-k)+1)\right) \right\} \\ &= \sum_{k=0}^{\frac{N-1}{2}} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1+\frac{N}{2}\right)(2k+1)\right) = \tilde{x}_i\end{aligned}$$

Through **Theorem 1** and **Theorem 2**, we can compute the MCT transform in (2.4) and (2.5) through (2.6) and (2.7), respectively. Define the DCT-like transform as follows

$$X_k = \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right), \text{ for } k=0, 1, \dots, N-1 \quad (2.8)$$

and

$$\tilde{u}_i = \sum_{k=0}^{N-1} X_k \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right), \text{ for } i=0, 1, \dots, N-1 \quad (2.9)$$

The following theorem sets the fundamental to compute TDAC transform through (2.8) and (2.9).

Theorem 3: Given (2.6) and (2.8), the sequence X'_k computed through (2.6) is equivalent to the sequence X_k computed through (2.8) if N is a multiple of 4 and the sequence u_i in (2.8) is permuted from the sequence x'_i in (2.6) through the following form

$$u_i = -x_{i+\frac{3N}{4}}, \text{ for } i=0, 1, \dots, \frac{N}{4}-1, \quad u_i = x_{i-\frac{N}{4}}, \text{ for } i = \frac{N}{4}, \frac{N}{4}+1, \dots, N-1 \quad (2.10)$$

<Proof>: Substituting $j=i+N/4$ into (6) gives

$$X'_k = \sum_{j=N/4}^{5N/4-1} x'_i \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) \text{ for } k=0, 1, \dots, N-1$$

Representing the summation into two terms yields

$$X'_k = \sum_{j=N/4}^{N-1} x'_i \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) + \sum_{j=N}^{5N/4-1} x'_j \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) \text{ for } k=0, 1, \dots, N-1 \quad (2.11)$$

Let $m=j-N$. Since

$$\cos\left(\frac{\pi}{2N}(2(m+N)+1)(2k+1)\right) = -\cos\left(\frac{\pi}{2N}(2m+1)(2k+1)\right),$$

(2.11) can be reformulated as

$$\begin{aligned} X'_k &= \sum_{j=N/4}^{N-1} x'_i \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) + \sum_{m=0}^{N/4-1} -x'_i \cos\left(\frac{\pi}{2N}(2m+1)(2k+1)\right) \\ &= \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) = X_k \end{aligned}$$

Theorem 4: Given (2.7) and (2.9) with $X'_k = X_k$, the sequence \tilde{x}'_i computed from (2.7) can be obtained from the sequence \tilde{u}_i computed from (2.9) through the following permutation:

$$\begin{aligned} \tilde{x}'_i &= \frac{1}{2} \tilde{u}_{i+\frac{N}{4}}, \text{ for } i=0, 1, \dots, \frac{3N}{4}-1 \\ \tilde{x}'_i &= \frac{1}{2} \tilde{u}_{\frac{7N}{4}-1-i}, \text{ for } i=\frac{3N}{4}, \frac{3N}{4}+1, \dots, N-1 \end{aligned} \quad (2.12)$$

<Proof>: From (2.7)

$$\tilde{x}'_i = \frac{1}{2} \sum_{k=0}^{N-1} X'_k \cos\left(\frac{\pi}{2N}\left(2i+1+\frac{N}{2}\right)(2k+1)\right) \text{ for } i=0, 1, \dots, N-1.$$

Consider the summation from two separate parts. For $0 \leq i < \frac{3N}{4}$,

$$\tilde{x}'_i = \frac{1}{2} \sum_{k=0}^{N-1} X'_k \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(2k+1)\right) = \frac{1}{2} \sum_{k=0}^{N-1} X'_k \cos\left(\frac{\pi}{2N} \left(2\left(i + \frac{N}{4}\right) + 1\right)(2k+1)\right) = \frac{1}{2} \tilde{u}_{i+\frac{N}{4}}$$

For $\frac{3N}{4} \leq i < N$

$$\tilde{x}'_i = \frac{1}{2} \sum_{k=0}^{N-1} X'_k \cos\left(\frac{\pi}{2N} \left(2\left(i + \frac{N}{4}\right) + 1\right)(2k+1)\right)$$

Since $\cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right) = \cos\left(\frac{\pi}{2N} (2(2N-1-i)+1)(2k+1)\right)$

$$\begin{aligned} \tilde{x}'_i &= \frac{1}{2} \sum_{k=0}^{N-1} X'_k \cos\left(\frac{\pi}{2N} \left(2\left(2N-1 - \left(i + \frac{N}{4}\right)\right) + 1\right)(2k+1)\right) \\ &= \frac{1}{2} \sum_{k=0}^{N-1} X'_k \cos\left(\frac{\pi}{2N} \left(2\left(\frac{7N}{4} - 1 - i\right) + 1\right)(2k+1)\right) = \frac{1}{2} \tilde{u}_{\frac{7N}{4}-1-i} \end{aligned}$$

To further derive the relation with type-IV DCT, we consider the following

Lemma:

Lemma 1: The sequence X_k computed through (2.8) is anti-symmetric in the sense that $X_{N-1-k} = -X_k$, for $k=0, 1, \dots, N-1$.

<Proof> From (2.8),

$$\begin{aligned} X_{N-1-k} &= \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2(N-1-k)+1)\right) \quad \text{for } k = 0, 1, \dots, N-1 \\ &= \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2N-2k-1)\right) \\ &= \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(-2k-1) + 2N \frac{\pi}{2N} (2i+1)\right) \\ &= \sum_{i=0}^{N-1} u_i \cos\left(\left(\frac{\pi}{2N} (2i+1)(-2k-1) + \pi\right)\right) \\ &= -\sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(-2k-1)\right) = -\sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right) \\ &= -X_k \end{aligned}$$

Substituting **Lemma 1** into (2.8) yields

$$\begin{aligned}
X_k &= \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right) \text{ for } k=0 \dots N-1 \\
&= \begin{cases} \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right) & \text{for } k=0, 1, \dots, \frac{N}{2}-1 \\ -X_{N-1-k} & \text{for } k=\frac{N}{2} \dots N-1 \end{cases}
\end{aligned} \tag{2.13}$$

Representing type-IV DCT with length (N/2) according to (2.3) gives

$$Y_k = \sum_{i=0}^{\frac{N}{2}-1} s_i \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right) \text{ for } k=0, 1, \dots, \frac{N}{2}-1 \tag{2.14}$$

The following two theorems set the basis to compute (2.8) and (2.9) through type-IV DCT in (2.14).

Theorem 5: Given (2.8) and (2.14), the sequence X_k in (2.8) for $k=0, 1, \dots, (N/2)-1$ will be equivalent to the sequence Y_k in (2.13) if

$$s_i = u_i - u_{N-1-i} \text{ , for } i = 0, 1, \dots, (N/2)-1 \tag{2.15}$$

<Proof> Representing the first term in (2.13) into two summation terms yields

$$\begin{aligned}
X_k &= \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right) \text{ for } k = 0, 1, \dots, N/2 - 1 \\
&= \sum_{i=0}^{N/2-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right) + \sum_{j=0}^{N/2-1} u_{j+\frac{N}{2}} \cos\left(\frac{\pi}{2N} (2(j+\frac{N}{2})+1)(2k+1)\right) \\
&= \sum_{i=0}^{N/2-1} u_i \cos\left(\frac{\pi}{2N} (2i+1)(2k+1)\right) - \sum_{j=0}^{N/2-1} u_{j+\frac{N}{2}} \cos\left(\frac{\pi}{2N} (2(\frac{N}{2}-1-j)+1)(2k+1)\right)
\end{aligned}$$

Let $N-1-m=j+(N/2)$

$$\begin{aligned}
X_k &= \sum_{i=0}^{N/2-1} u_i \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) - \sum_{m=0}^{N/2-1} u_{N-1-m} \cos\left(\frac{\pi}{2N}\left(2\left(\frac{N}{2}-1-\left(\frac{N}{2}-1-m\right)\right)+1\right)(2k+1)\right) \\
&= \sum_{i=0}^{N/2-1} u_i \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) - \sum_{m=0}^{N/2-1} u_{N-1-m} \cos\left(\frac{\pi}{2N}(2m+1)(2k+1)\right) \\
&= \sum_{i=0}^{N/2-1} (u_i - u_{N-1-i}) \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) = Y_k
\end{aligned}$$

To proceed with the following derivation, (2.14) is rewritten by interchanging the indices i and k as follows

$$Y_i = \sum_{k=0}^{N/2-1} s_k \cos\left(\frac{\pi}{2N}(2k+1)(2i+1)\right) \quad \text{for } i=0, 1, \dots, \frac{N}{2}-1
\tag{2.16}$$

Theorem 6: Given (2.9) and (2.16) with $s_k = 2X_k$ and X_k has the anti-symmetric property described in **Lemma 1**, the sequence \tilde{u}_i in (2.9) for $i=0, 1, \dots, (N/2)-1$ is equivalent to the sequence Y_i of type-IV DCT in (2.16).

<Proof> From (2.9) $\tilde{u}_i = \sum_{k=0}^{N-1} X_k \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right)$, for $i=0, 1, \dots, N-1$

From **Lemma 1**, $X_{N-1-k} = -X_k$. Hence

$$\begin{aligned}
\tilde{u}_i &= \sum_{k=0}^{N/2-1} (X_k - X_{N-1-k}) \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) \\
&= 2 \sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right)
\end{aligned}$$

To summarize, the forward MCT in (2.4) can be computed through the type-IV DCT in (2.14) with the input permutation through (2.10) and (2.15) in **Theorem 3** and **Theorem 5**. From **Theorem 2**, **Theorem 4**, and **Theorem 6**, the inverse MCT in (2.5) can be computed through the type-IV DCT in (2.16)

with the output permutation in (2.12).

2.2.2 Unified form for the variant of TDAC filterbanks

Two variants of time-domain aliasing cancellation (TDAC) filterbank have been adopted in the Dolby AC-3 coder to provide the perfect reconstruction property between different block sizes [43]. The first transform pair is defined as

$$X'_k = \sum_{i=0}^{\frac{N}{2}-1} x'_i \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right), \text{ for } k=0,1, \dots, N/2-1 \quad (2.17)$$

$$\tilde{x}'_i = \sum_{k=0}^{\frac{N}{2}-1} X'_k \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right), \text{ for } i=0,1, \dots, N-1 \quad (2.18)$$

The second transform pair is

$$X''_k = \sum_{i=0}^{\frac{N}{2}-1} \tilde{x}''_i \cos\left(\frac{\pi}{2N}(2i+1+N)(2k+1)\right), \text{ for } k=0,1, \dots, N/2-1 \quad (2.19)$$

$$\tilde{x}''_i = \sum_{k=0}^{\frac{N}{2}-1} X''_k \cos\left(\frac{\pi}{2N}(2i+1+N)(2k+1)\right), \text{ for } i=0,1, \dots, N-1 \quad (2.20)$$

This section demonstrates that (2.17)-(2.20) can be derived as permutation and type-IV DCT. First we set the relation between the transform pair in (2.17)-(2.18) and that in (2.8)-(2.9).

Theorem 7: Let the sequence X_k in (2.9) with length N be obtained by extending the sequence X_k^1 with length $(N/2)$ according to $X_k = -X_{N-k-1}$ for

$k=N/2, \dots, N-1$. Given (2.9) and (2.18), the sequence \tilde{u}_i computed from (2.9) is two times the sequence \tilde{x}_i^1 computed from (2.18).

<Proof>: From (2.9),

$$\begin{aligned} \tilde{u}_i &= \sum_{k=0}^{N-1} X_k \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) \\ &= \left\{ \sum_{k=0}^{\frac{N-1}{2}} X_k \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) + \sum_{k=\frac{N}{2}}^{N-1} X_k \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) \right\} \end{aligned}$$

Since $X_k = -X_{N-1-k}$,

$$\begin{aligned} \tilde{u}_i &= \sum_{k=0}^{\frac{N-1}{2}} X_k \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) + \sum_{k=0}^{\frac{N-1}{2}} X_{N-1-k} \cos\left(\frac{\pi}{2N}(2i+1)(2(N-1-k)+1)\right) \\ &= \sum_{k=0}^{\frac{N-1}{2}} X_k \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) - \sum_{k=0}^{\frac{N-1}{2}} X_k \cos\left(\frac{\pi}{2N}(2i+1)(2(N-1-k)+1)\right) \\ &= 2 \sum_{k=0}^{\frac{N-1}{2}} X_k \cos\left(\frac{\pi}{2N}\left(2i+1+\frac{N}{2}\right)(2k+1)\right) = 2\tilde{x}_i' \end{aligned}$$

Lemma 1 and **Theorem 7** set the fundamental to derive the TDAC-variant in (2.17) and (2.18) through DCT-like transform in (2.8) and (2.9). From **Theorem 5** and **Theorem 6**, the DCT-like transform can be computed through the type-IV DCT. Hence, the first form of the TDAC-variant transform can be derived into the permutation and type-IV DCT.

The following two theorems illustrate the relation between the MCT of the TDAC-variant in (2.17)-(2.18) and that in (2.19)-(2.20).

Theorem 8: Given (2.17) and (2.19), the sequence X_k^2 in (2.19) is equivalent to X_k^1 if

$$x_i'' = -x'_{i+\frac{N}{2}} \text{ for } i=0, 1, \dots, \frac{N}{2}-1, x_i'' = -x'_{i-\frac{N}{2}} ; i = \frac{N}{2}, \frac{N}{2}+1, \dots, N-1 \quad (2.21)$$

<Proof>: Substituting $j=i-N/2$ into (2.19) yields

$$\begin{aligned} X_k'' &= \sum_{j=N/2}^{3N/2-1} \dot{x}_{j-\frac{N}{2}}'' \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) \text{ for } k=0, 1, \dots, N-1 \\ &= \sum_{j=N/2}^{N-1} \dot{x}_{j-\frac{N}{2}}'' \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) + \sum_{j=N}^{3N/2-1} \dot{x}_{j-\frac{N}{2}}'' \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) \end{aligned}$$

Let $m=j-N$

$$X_k'' = \sum_{j=N/2}^{N-1} \dot{x}_{j-\frac{N}{2}}'' \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) + \sum_{m=0}^{N/2-1} \dot{x}_{m+\frac{N}{2}}'' \cos\left(\frac{\pi}{2N}(2m+2N+1)(2k+1)\right)$$

Since

$$\begin{aligned} \cos\left(\frac{\pi}{2N}(2m+2N+1)(2k+1)\right) &= -\cos\left(\frac{\pi}{2N}(2m+1)(2k+1)\right) \\ X_k'' &= \sum_{j=N/2}^{N-1} \dot{x}_{j-\frac{N}{2}}'' \cos\left(\frac{\pi}{2N}(2j+1)(2k+1)\right) - \sum_{m=0}^{N/2-1} \dot{x}_{m+\frac{N}{2}}'' \cos\left(\frac{\pi}{2N}(2m+1)(2k+1)\right) \\ &= X_k' \end{aligned}$$

Theorem 9: Given (2.18) and (2.20), the sequence \tilde{x}_i^2 in (2.20) is equivalent to

\tilde{x}_i^1 if

$$\tilde{x}_i'' = -\tilde{x}'_{i+\frac{N}{2}} \text{ for } i=0, 1, \dots, \frac{N}{2}-1, \tilde{x}_i'' = \tilde{x}'_{i-\frac{N}{2}} ; i = \frac{N}{2}, \frac{N}{2}+1, \dots, N-1 \quad (2.22)$$

<Proof>: From (2.20),

$$\tilde{x}_i'' = \sum_{k=0}^{N/2-1} X_k'' \cos\left(\frac{\pi}{2N}\left(2\left(i+\frac{N}{2}\right)+1\right)(2k+1)\right), \text{ for } i=0, 1, \dots, N-1$$

For $0 \leq i < N/2$

$$\tilde{x}_i'' = \sum_{k=0}^{N/2-1} X_k'' \cos\left(\frac{\pi}{2N}\left(2\left(i+\frac{N}{2}\right)+1\right)(2k+1)\right) = \tilde{x}'_{i+\frac{N}{2}}$$

For $N/2 \leq i < N$

$$\begin{aligned}\tilde{x}_i'' &= \sum_{k=0}^{N/2-1} X_k'' \cos\left(\frac{\pi}{2N} \left(2\left(i + \frac{N}{2}\right) + 1\right)(2k + 1)\right) \\ &= - \sum_{k=0}^{N/2-1} X_k'' \cos\left(\frac{\pi}{2N} \left(2\left(i + \frac{N}{2} - N\right) + 1\right)(2k + 1)\right) = -\tilde{x}'_{i-\frac{N}{2}}\end{aligned}$$

The computation of the two variants of TDAC filterbank defined in equations (2.17)-(2.20) can be computed with the following remarks:

Computing Process for (2.17): From **Theorem 5**, the MCT of the first TDAC-variant in (2.17) can be computed directly through the type-IV DCT in (2.14) with the input permutation

$$s_i = x'_i - x'_{N-1-i}, \text{ for } i = 0, 1, \dots, (N/2)-1 \quad (2.23)$$

Computing Process for (2.18): From **Theorem 6** and **Theorem 7**, the inverse MCT of the first TDAC-variant in (2.18) can be computed directly through the type-IV DCT in (2.16).

Computing Process for (2.19): From **Theorem 5** and **Theorem 8**, the MCT of the second TDAC-variant in (2.19) can be computed directly through the type-IV DCT in (2.14) with the input permutation in (2.21) and (2.23).

Computing Process for (2.20): From **Theorem 6**, **Theorem 7** and **Theorem 9** the inverse MCT of the second TDAC-variant in (2.20) can be computed directly through the type-IV DCT in (2.16) through the output permutation in (2.22).

2.2.3 Unified form for the polyphase filterbank

The transform pair for the cosine modulation transform in the polyphase

filterbank [43] is

$$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{N}\left(i - \frac{N}{4}\right)(2k+1)\right) \quad k=0, 1, \dots, \frac{N}{2}-1 \quad (2.24)$$

$$\tilde{x}_i = \sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{N}\left(i + \frac{N}{4}\right)(2k+1)\right) \quad i=0, 1, \dots, N-1 \quad (2.25)$$

To proceed with the derivation, we define the following two transform formulae

$$X'_k = \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{N}(i)(2k+1)\right) \quad \text{for } k=0, 1, \dots, \frac{N}{2}-1 \quad (2.26)$$

$$\tilde{u}_i = \sum_{k=0}^{N/2-1} X'_k \cos\left(\frac{\pi}{N}(i)(2k+1)\right) \quad \text{for } i=0, 1, \dots, N-1 \quad (2.27)$$

The derivation proceeds with two steps. First, we show (2.24) and (2.25) can be computed through (2.26) and (2.27) with permutation through **Theorem 10** and **Theorem 11**. Second, we show (2.26) and (2.27) can be computed through type-III DCT through **Theorem 12** and **Theorem 13**.

Theorem 10: Let N be an integer that is a multiple of four. Given (2.24) and (2.26), the sequence X'_k computed through (2.26) is equivalent to the sequence

X_k computed through (2.24) if

$$u_i = x_{i+\frac{N}{4}} \quad i=0, 1, \dots, \frac{3N}{4}-1 \quad u_i = -x_{i-\frac{3N}{4}} \quad i=\frac{3N}{4}, \frac{3N}{4}+1, \dots, N-1 \quad (2.28)$$

<Proof>: Let $j=i-N/4$. Rewrite (2.24) as

$$X_k = \sum_{j=-\frac{N}{4}}^{-1} x_{j+\frac{N}{4}} \cos\left(\frac{\pi}{N}(j)(2k+1)\right) + \sum_{j=0}^{3N/4-1} x_{j+\frac{N}{4}} \cos\left(\frac{\pi}{N}(j)(2k+1)\right), \text{ for } k=0,1,\dots,(N/2)-1$$

Let $m=j+N$. Since

$$\cos\left(\frac{\pi}{N}(m-N)(2k+1)\right) = -\cos\left(\frac{\pi}{N}(m)(2k+1)\right)$$

$$X_k = \sum_{m=\frac{3N}{4}}^{N-1} -x_{m-\frac{3N}{4}} \cos\left(\frac{\pi}{N}(m)(2k+1)\right) + \sum_{j=0}^{3N/4-1} x_{j+\frac{N}{4}} \cos\left(\frac{\pi}{N}(j)(2k+1)\right) = X'_k$$

Theorem 11: Let N is a multiple of four. Given (2.25) and (2.27) and $X_k = X'_k$,

the sequence \tilde{x}_i computed through (2.25) can be permuted from the sequence \tilde{u}_i

computed through (2.27) with the following form

$$\tilde{x}_i = \tilde{u}_{i+N/4} \text{ for } i=0,1, \dots, \frac{3N}{4}-1 \text{ and } \tilde{x}_i = -\tilde{u}_{i-3N/4} \text{ for } i=\frac{3N}{4}, \frac{3N}{4}+1, \dots, N-1$$

(2.29)

<Proof >: Rewrite (2.25) as

$$\tilde{x}_i = \sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{N}\left(i+\frac{N}{4}\right)(2k+1)\right) \quad i=0, 1, \dots, N-1$$

For $0 \leq i < 3N/4$

$$\tilde{x}_i = \sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{N}\left(i+\frac{N}{4}\right)(2k+1)\right) = \tilde{u}_{i+\frac{N}{4}}$$

For $3N/4 \leq i < N$

$$\tilde{x}_i = \sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{N}\left(i+\frac{N}{4}\right)(2k+1)\right) = -\sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{N}\left(i+\frac{N}{4}-N\right)(2k+1)\right) = -\tilde{u}_{i-3N/4}$$

According to(2.1), type-II DCT with length $(N/2)$ is

$$X_i = \sum_{k=0}^{\frac{N}{2}-1} x_k \cos\left(\frac{\pi}{N}(2k+1)(i)\right) \quad \text{for } i=0,1, \dots, \frac{N}{2}-1$$

(2.30)

Theorem 12: Given (2.27) and (2.30), let $X'_k = x_k$. The sequence \tilde{u}_i computed through (2.27) can be obtained from sequence X_i through

$$\begin{cases} \tilde{u}_i = 0 & \text{for } i=N/2 \\ \tilde{u}_i = X_i & \text{for } i=0, 1, \dots, \frac{N}{2}-1 \\ \tilde{u}_i = -X_{N-i} & \text{for } i=\frac{N}{2}+1, \frac{N}{2}+2, \dots, N-1 \end{cases} \quad (2.31)$$

<Proof>: Rewrite (2.27) as

$$\tilde{u}_i = \sum_{k=0}^{N/2-1} X'_k \cos\left(\frac{\pi}{N}(i)(2k+1)\right) \text{ for } i=0, 1, \dots, N-1$$

For $0 \leq i < \frac{N}{2}$

$$\tilde{u}_i = \sum_{k=0}^{N/2-1} X'_k \cos\left(\frac{\pi}{N}(i)(2k+1)\right) = X_i$$

For $i = \frac{N}{2}$,

$$\tilde{u}_i = \sum_{k=0}^{N/2-1} X'_k \cos\left(\frac{\pi}{2}(2k+1)\right) = 0$$

For $\frac{N}{2}+1 \leq i < N$, since

$$\cos\left(\frac{\pi}{N}(i)(2k+1)\right) = -\cos\left(\frac{\pi}{N}(N-i)(2k+1)\right) \text{ for } i=1, 2, \dots, \frac{N}{2}-1, k=0, 1, \dots, \frac{N}{2}-1,$$

we have the following relation

$$\tilde{u}_i = \sum_{k=0}^{N/2-1} X'_k \cos\left(\frac{\pi}{N}(i)(2k+1)\right) = -\sum_{k=0}^{N/2-1} X'_k \cos\left(\frac{\pi}{N}(N-i)(2k+1)\right) = -X_{N-i}$$

According to (2.2), type-III DCT with length (N/2) is

$$X_k = \sum_{i=0}^{\frac{N-1}{2}} x_i \cos\left(\frac{\pi}{N}(i)(2k+1)\right) \text{ for } k=0, 1, \dots, \frac{N}{2}-1 \quad (2.32)$$

Theorem 13: Given (2.26) and (2.32), the sequence X_k in (2.32) is equivalent to the sequence X'_k in (2.26) if the sequence x_i is computed from the sequence u_i through

$$\begin{cases} x_0 = u_0 \\ x_i = u_i - u_{N-1-i} \text{ for } i=1, \dots, \frac{N}{2}-1 \end{cases} \quad (2.33)$$

<Proof>: From (2.26)

$$\begin{aligned} X'_k &= \sum_{i=0}^{N-1} u_i \cos\left(\frac{\pi}{N}(i)(2k+1)\right), \text{ for } k = 0, 1, \dots, N/2-1 \\ &= u_0 \cos\left(\frac{\pi}{N}(0)(2k+1)\right) + \sum_{i=1}^{N/2-1} u_i \cos\left(\frac{\pi}{N}(i)(2k+1)\right) + \\ &u_{\frac{N}{2}} \cos\left(\frac{\pi}{N}\left(\frac{N}{4}\right)(2k+1)\right) + \sum_{i=1}^{N/2-1} u_{N-i} \cos\left(\frac{\pi}{N}(N-i)(2k+1)\right) \end{aligned}$$

Since

$$u_{\frac{N}{2}} \cos\left(\frac{\pi}{N}\left(\frac{N}{4}\right)(2k+1)\right) = 0$$

and

$$\begin{aligned} \cos\left(\frac{\pi}{4N}(N-i)(2k+1)\right) &= -\cos\left(\frac{\pi}{N}(i)(2k+1)\right) \\ X'_k &= (u_0) \cos\left(\frac{\pi}{N}(0)(2k+1)\right) + \sum_{i=1}^{N-1} (u_i - u_{N-1-i}) \cos\left(\frac{\pi}{N}(i)(2k+1)\right) = X_k \end{aligned}$$

The MCT in (2.24) can be computed through the type-III DCT in (2.32) with the input permutation through (2.28) and (2.33) in **Theorem 10** and **Theorem 13**.

From **Theorem 10** and **Theorem 12**, the inverse MCT in (2.25) can be computed through the type-II DCT in (2.30) with the output permutation in (2.29) and (2.31).

2.3 Fast Algorithm for the Discrete Cosine Transform

Section 2.2 illustrates that the various cosine-modulated transforms used in TDAC, TDAC-variant, and polyphase filterbanks can be divided into two modules: permutation and the DCT. Especially, the forward transform can be represented as the pre-permutation and the DCT while the inverse transform as the DCT and the post-permutation. The DCT can be type-II, type-III, or type-IV. This section develops a method to decompose a type of DCT with length N into two of the three types of the DCT with length $N/2$. The decomposition method will be proved to have the regularity and the modularity in addition to the low complexity. Furthermore, the algorithm is applicable to the cosine-modulated transforms in audio coding standards.

2.3.1 Decomposition for type-II DCT

From (2.1), the k th coefficient of the type-II DCT for an input sequence x_i with length N is

$$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(2i+1)(k)\right) \text{ for } k=0 \dots N-1$$

We first decompose X_k of the type-II DCT into even-indexed and odd-indexed forms. The even-indexed output sequence is

$$X_{2k} = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(2i+1)(2k)\right), \text{ for } k=0, \dots, \frac{N}{2}-1. \quad (2.34)$$

Applying the symmetry property $\cos\left(\frac{\pi}{N}(2(N-1-i)+1)(k)\right) = \cos\left(\frac{\pi}{N}(2i+1)(k)\right)$ gives

$$X_{2k} = \sum_{i=0}^{N/2-1} (x_i + x_{N-1-i}) \cos\left(\frac{\pi}{N}(2i+1)(k)\right) \quad (2.35)$$

which is a type-II DCT with input permutation.

The odd-indexed output sequence is

$$X_{2k+1} = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right), \text{ for } i=0, \dots, \frac{N}{2}-1$$

Applying the anti-symmetry property

$$\cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) = -\cos\left(\frac{\pi}{2N}(2(N-1-i)+1)(2k+1)\right)$$

gives

$$X_{2k+1} = \sum_{i=0}^{N/2-1} (x_i - x_{N-1-i}) \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) \quad (2.36)$$

which is a type-IV DCT with input permutation. From (2.35) and (2.36), a type-II DCT with length N can be decomposed into one type-II DCT and one type-IV with length (N/2) as illustrated in Fig. 2.3.

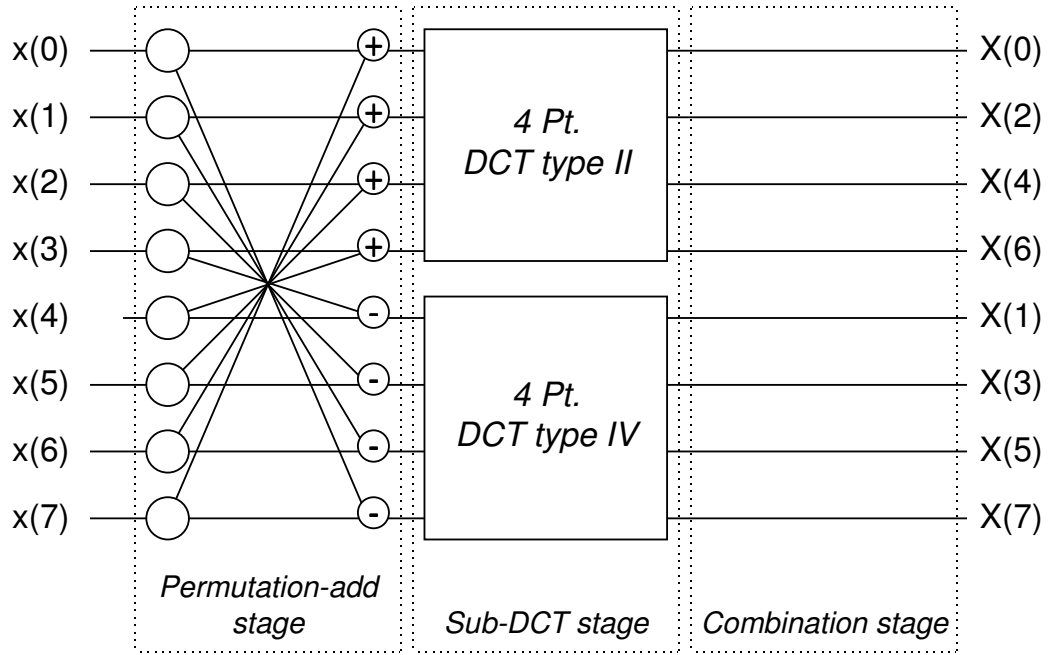


Fig. 2.3 The decomposition of one 8-point type-II DCT into one 4-point type-II DCT and one 4-point type-IV DCT.

2.3.2 Decomposition for type-III DCT

From (2.2), the k th coefficient of type-III DCT for an input sequence x_i with length N is

$$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(i)(2k+1)\right) \text{ for } k=0, 1, \dots, N-1 \quad (2.37)$$

We separate both the input sequence x_i and the output sequence of type-III DCT. The input is separated into even-indexed and odd-indexed forms while the output is separated into the first half of the sequence and the second half of the sequence; that is,

$$X_k = \sum_{i=0}^{N/2-1} x_{2i} \cos\left(\frac{\pi}{N}(i)(2k+1)\right) + \sum_{i=0}^{N/2-1} x_{2i+1} \cos\left(\frac{\pi}{2N}(2i+1)(2k+1)\right) \text{ for } k=0, 1, \dots, \frac{N}{2}-1 \quad (2.38)$$

$$X_{\frac{N}{2}+k} = \sum_{i=0}^{N/2-1} x_{2i} \cos\left(\frac{\pi}{N} (i)(2(k + \frac{N}{2}) + 1)\right) + \sum_{i=0}^{N/2-1} x_{2i+1} \cos\left(\frac{\pi}{2N} (2i+1)(2(k + \frac{N}{2}) + 1)\right), \text{ for } k=0,1, \dots, \frac{N}{2}-1$$

(2.39)

Substituting

$$\cos\left(\frac{\pi}{N} (i)(2(k + \frac{N}{2}) + 1)\right) = \cos\left(\frac{\pi}{N} (i)(2(\frac{N}{2} - 1 - k) + 1)\right)$$

and

$$\cos\left(\frac{\pi}{N} (2i+1)(2(k + \frac{N}{2}) + 1)\right) = -\cos\left(\frac{\pi}{N} (2i+1)(2(\frac{N}{2} - 1 - k) + 1)\right)$$

into (2.39) yields

$$X_{\frac{N}{2}+k} = \sum_{i=0}^{N/2-1} x_{2i} \cos\left(\frac{\pi}{N} (i)(2(\frac{N}{2} - 1 - k) + 1)\right) - \sum_{i=0}^{N/2-1} x_{2i+1} \cos\left(\frac{\pi}{2N} (2i+1)(2(\frac{N}{2} - 1 - k) + 1)\right)$$

for $k=0 \dots N/2-1$

(2.40)

From (2.38) and (2.40), a type-III DCT with length N can be decomposed into one type-III DCT and one type-IV DCT with length (N/2) as illustrated in Fig.

2.4.

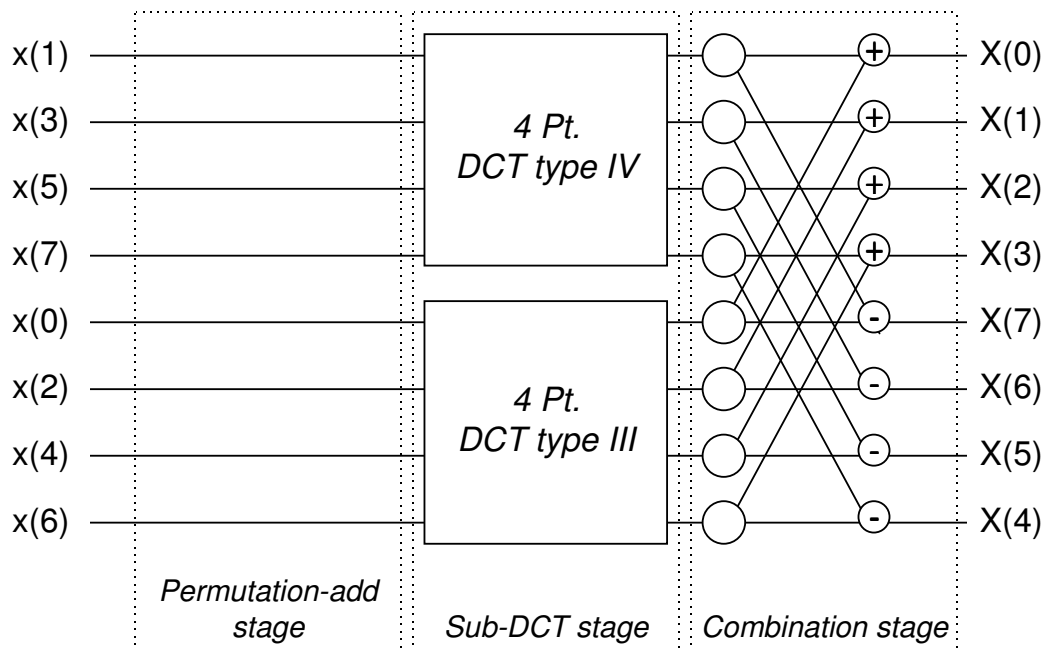


Fig. 2.4 The decomposition of one 8-point type-III DCT into one 4-point type-III DCT and one 4-point type-IV DCT.

2.3.3 Decomposition for type-IV DCT

Before proceeding with the derivation, we consider the following property.

Lemma 2: An $(N+1) \times N$ type-III DCT can be simplified into an $N \times N$ type-III DCT

$$\sum_{i=0}^N x_i \cos\left(\frac{\pi}{2N}(i)(2k+1)\right) = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(i)(2k+1)\right)$$

<Proof>: **Lemma 2** can be directly derived as follows:

$$\begin{aligned} & \sum_{i=0}^N x_i \cos\left(\frac{\pi}{2N}(i)(2k+1)\right) \\ &= \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(i)(2k+1)\right) + x_N \cos\left(\frac{\pi}{2N}(N)(2k+1)\right) \\ &= \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(i)(2k+1)\right) + x_N \cos\left(k\pi + \frac{\pi}{2}\right) \\ &= \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N}(i)(2k+1)\right) \end{aligned}$$

From (2.3), the k th coefficient of type-IV DCT for an input sequence x_i with length N is

$$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{4N}(2i+1)(2k+1)\right) \quad \text{for } k=0 \dots N-1 \quad (2.41)$$

Since $\cos A = \frac{1}{2 \cos B} (\cos(A+B) + \cos(A-B))$, (2.41) can be represented as

$$X_k = \frac{1}{2 \cos\left(\frac{\pi}{4N}(2k+1)\right)} \sum_{i=0}^{N-1} x_i (\cos\left(\frac{\pi}{2N}(2k+1)(i)\right) + \cos\left(\frac{\pi}{2N}(2k+1)(i+1)\right)) \quad (2.42)$$

Separating input sequences into even and odd terms yields

$$\begin{aligned}
X_k = & \frac{1}{2 \cos\left(\frac{\pi}{4N}(2k+1)\right)} \left\{ \sum_{i=0}^{N/2-1} x_{2i} \cos\left(\frac{\pi}{N}(2k+1)(2i)\right) + \sum_{i=0}^{N/2-1} x_{2i+1} \cos\left(\frac{\pi}{N}(2k+1)(2i+1)\right) \right. \\
& \left. + \sum_{i=0}^{N/2-1} x_{2i} \cos\left(\frac{\pi}{2N}(2k+1)(2i+1)\right) + \sum_{i=0}^{N/2-1} x_{2i+1} \cos\left(\frac{\pi}{2N}(2k+1)(2i+2)\right) \right\}
\end{aligned} \tag{2.43}$$

Set $x_{-1} = x_N = 0$, the four terms in (2.43) can be represented as

$$\begin{aligned}
X_k = & \frac{1}{2 \cos\left(\frac{\pi}{4N}(2k+1)\right)} \\
& \left\{ \sum_{i=0}^{N/2} (x_{2i} + x_{2i-1}) \cos\left(\frac{\pi}{N}(2k+1)(i)\right) + \sum_{i=0}^{N/2-1} (x_{2i} + x_{2i+1}) \cos\left(\frac{\pi}{2N}(2k+1)(2i+1)\right) \right\}
\end{aligned}$$

From **Lemma 2**,

$$\begin{aligned}
X_k = & \frac{1}{2 \cos\left(\frac{\pi}{4N}(2k+1)\right)} \\
& \left\{ \sum_{i=0}^{N/2-1} (x_{2i} + x_{2i-1}) \cos\left(\frac{\pi}{N}(2k+1)(i)\right) + \sum_{i=0}^{N/2-1} (x_{2i} + x_{2i+1}) \cos\left(\frac{\pi}{2N}(2k+1)(2i+1)\right) \right\}
\end{aligned} \tag{2.44}$$

From (2.44), a type-IV DCT with length N can be decomposed into one type-IV

DCT and one type-III DCT with length (N/2) as illustrated in Fig. 2.5.

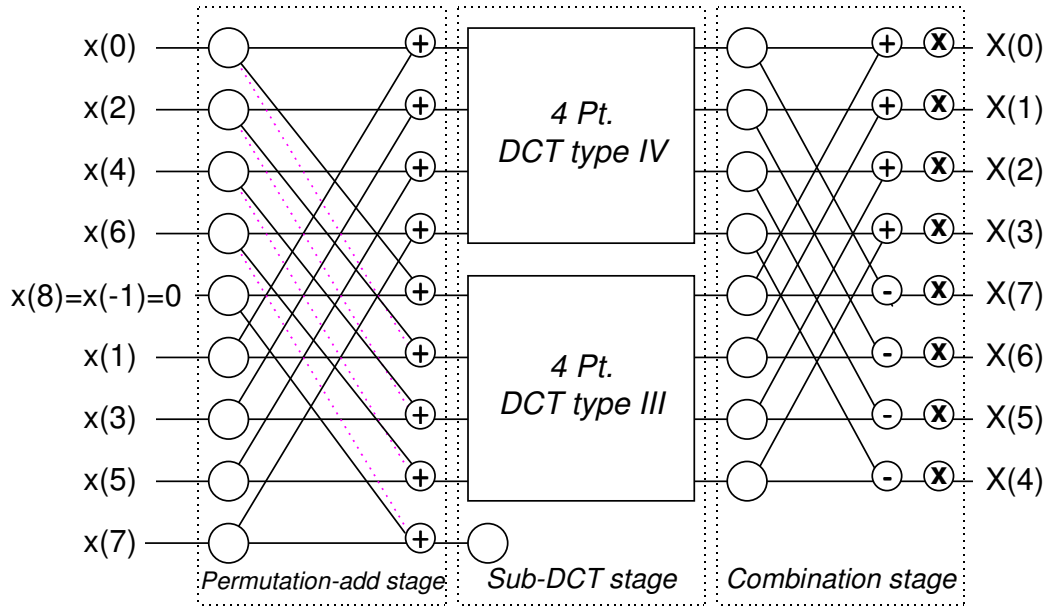


Fig. 2.5 The decomposition of one 8-point type-IV DCT into one 4-point type-III DCT and one 4-point type-IV DCT.

From Fig. 2.3-Fig. 2.5, the arithmetic complexities for all three types of the DCT are individually

$$\text{DCT-II}(N) = A(N) + \text{DCT-IV}(N/2) + \text{DCT-II}(N/2),$$

$$\text{DCT-III}(N) = A(N) + \text{DCT-IV}(N/2) + \text{DCT-III}(N/2),$$

and $\text{DCT-IV}(N) = A(N-1) + M(N) + \text{DCT-IV}(N/2) + \text{DCT-III}(N/2)$

where $\text{DCT-II}(N)$, $\text{DCT-III}(N)$, and $\text{DCT-IV}(N)$ are the arithmetic complexity of the type-II, type-III, and type-IV DCT with length N . $A(\mu)$ and $M(\kappa)$ indicate the number of real addition and multiply are μ and κ , respectively. Table 2.2 lists the arithmetic complexity of the new algorithm and the existing algorithms[2][16][37][45] for the radix-2 DCTs. The results illustrate that the fast algorithm not only unifies the computing methods for types II, III, and IV DCT but also has a complexity as low as the well-known algorithms.

2.4 Concluding Remarks

Variant forms of the modulated cosine transforms (MCTs) have been widely used in different audio standards. This section has illustrated that all these MCTs can be derived into two modules: the permutation and the discrete cosine transform. Especially the MCTs in encoders are derived as an input permutation and the DCT while the MCTs in decoder the DCT and the post permutation. The derived DCTs are either type-II, type-III, or type IV.

This chapter has proposed a new fast algorithm for the above three types of discrete cosine transform. The new algorithm has been developed with decomposition from one type of the DCT into the interleaving of type-II, type-III, or type-IV. The fast algorithm has been shown not only the low complexity but also has good features in regularity, complexity, and general applicability in all MCTs in audio coding standards. This chapter is adopted from [15], [12].

Classes	MCT transform pair	CMFBs in standards
TDAC	$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(2k+1)\right)$ $x_i = \sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{2N} \left(2i+1 + \frac{N}{2}\right)(2k+1)\right)$ <p>for $k = 0, 1, \dots, N/2 - 1$ and $i = 0, 1, \dots, N - 1$</p>	MPEG-4, MPEG-2—AAC, MPEG layer 3 2 nd Level, AC-3 Long Transform

TDAC-Variant	$X_k = \sum_{i=0}^{\dot{N}-1} x_i \cos\left(\frac{\pi}{2\dot{N}}(2i+1)(2k+1)\right)$ $x_i = \sum_{k=0}^{\dot{N}/2-1} X_k \cos\left(\frac{\pi}{2\dot{N}}(2i+1)(2k+1)\right)$ <p>for $k = 0, 1, \dots, \dot{N}/2 - 1$ and $i = 0, 1, \dots, \dot{N} - 1$</p>	AC-3 Short Transform 1
	$X_k = \sum_{i=0}^{\dot{N}-1} \dot{x}_i \cos\left(\frac{\pi}{2\dot{N}}(2i+1+\dot{N})(2k+1)\right)$ $x_i = \sum_{k=0}^{\dot{N}/2-1} X_k \cos\left(\frac{\pi}{2\dot{N}}(2i+1+\dot{N})(2k+1)\right)$ <p>for $k = 0, 1, \dots, \dot{N}/2 - 1$ and $i = 0, 1, \dots, \dot{N} - 1$</p>	AC-3 Short Transform 2
Polyphase Filter Bank	$X_k = \sum_{i=0}^{N-1} x_i \cos\left(\frac{\pi}{N}\left(i - \frac{N}{4}\right)(2k+1)\right)$ $x_i = \sum_{k=0}^{N/2-1} X_k \cos\left(\frac{\pi}{N}\left(i + \frac{N}{4}\right)(2k+1)\right)$ <p>for $k = 0, 1, \dots, N/2 - 1$ and $i = 0, 1, \dots, N - 1$</p>	MPEG layers 1, 2, MPEG layer 3 1 st Level

Table 2.1 The formulae and the classification of the CMFBs in current audio coding standards.

	Proposed DCT II		Proposed DCT III		Proposed DCT IV		[4],[9], [10] DCT II		[8] DCT III		[4] DCT IV	
<i>Op.</i>	<i>x</i>	<i>+</i>	<i>x</i>	<i>+</i>	<i>x</i>	<i>+</i>	<i>x</i>	<i>+</i>	<i>x</i>	<i>+</i>	<i>x</i>	<i>+</i>
2	1	2	1	2	3	3	1	2	1	2	3	3
4	4	9	4	9	8	12	4	9	4	9	8	12
8	12	29	12	29	20	36	12	29	12	29	20	36
16	32	81	32	81	48	96	32	81	32	81	48	96
32	80	209	80	209	112	240	80	209	80	209	112	240
64	192	513	192	513	256	588	192	513	192	513	256	588

Table 2.2 Arithmetic operations required in the fast algorithms of DCTs where *Op* stands for the arithmetic operations required for the row, where *x* denotes

multiplication operation while + addition operation. The 2, 4, 8, 16, 32, and 64 in first column denote the transform length. The entries of the row associating with the transform length illustrate the operations required for the algorithm labeled in the entry of the first row of the column.

Chapter 3

Fast Frequency Analysis for the Psychoacoustic Model

3.1 Introduction

For the perceptual audio coder as illustrated in Fig. 1.1, the frequency analyzer are required in the psychoacoustic model and the filterbank. In the psychoacoustic model, frequency information is required to model hearing model and thus a frequency analysis is required. For filterbank, frequency analysis is necessary to transform signals from time domain to frequency domain to remove the redundancy from the psychoacoustic model. A summary on frequency analysis schemes in filterbanks and psychoacoustic model are given in Table 3.1. For MPEG group, the frequency analysis on filterbank and psychoacoustic model is implemented in different approaches: Fourier transform and subband/hybrid filterbank. AC-3 coder uses the same frequency analyzer in both filterbank and psychoacoustic model. Obviously, from the viewpoint of the computation loading, the design of AC-3 coder is more efficient than the one of MPEG group due to the redundant computation of frequency analysis on psychoacoustic model and filterbank.

Standards	Filterbank	Frequency analysis in psychoacoustic model
MPEG-1 layer 1/2	Subband	1024 pt. Fourier transform
MPEG-1 layer 3	Hybrid	1024 pt and 256 pt. Fourier transform
MPEG-2 layer 1-3	Hybrid	1024 pt. and 256 pt. Fourier transform
MPEG-2 AAC	Subband/hybrid	Fourier transform
Dolby AC-3	Transform	Transform

Table 3.1 Audio standards and frequency analysis in psychoacoustic model.

Hybrid filterbank mentioned in [33] would be one solution for efficiently computing the frequency analysis for MPEG groups while maintaining the same frequency resolution required in the psychoacoustic model. This chapter applies the hybrid filterbank to the psychoacoustic model to reduce the computing complexity and improve the quality.

3.2 Hybrid Filterbank for Psychoacoustic Model in MPEG

The ISO/MPEG layer 1/2 audio compression is receiving a wide range of applications. In the encoding process of MPEG, the psychoacoustic model exploits audio irrelevancy that is the key role to achieve high compression ratio without losing audio quality. However, the Fourier transform (FT) which has been used by the two psychoacoustic models suggested in standard draft requires high computational complexity, and hence leads to high hardware and software cost for real-time applications. This section presents a new design named the hybrid filterbank to replace the FT. The hybrid filterbank can be integrated with the psychoacoustic model and provides a much lower

complexity than the FT. Also, this section shows that the hybrid filter is more suitable for the stereo coding and hence can provide a better quality for the intensity stereo coding, which is the key technology for the MPEG-1 to achieve near transparent quality lower than 96x2 kbits for two stereo channels.

Like most perceptual audio coders [28][40][33], MPEG audio encoder can be considered from four parts: the time-frequency mapper, the psychoacoustic model, quantization and frame packing as shown in Fig. 3.1. The psychoacoustic model exploits audio irrelevancy that is usually defined in frequency domain. The time-frequency mapper maps the time-domain signals into a frequency representation to reduce the data redundancy and provides the ease with the integration with the psychoacoustic model. The quantization quantizes the audio signals from time-frequency mapper based on the information from the psychoacoustic model. The frame packing packs the quantized signals with some synchronous information like sampling frequency for identified by MPEG decoders.

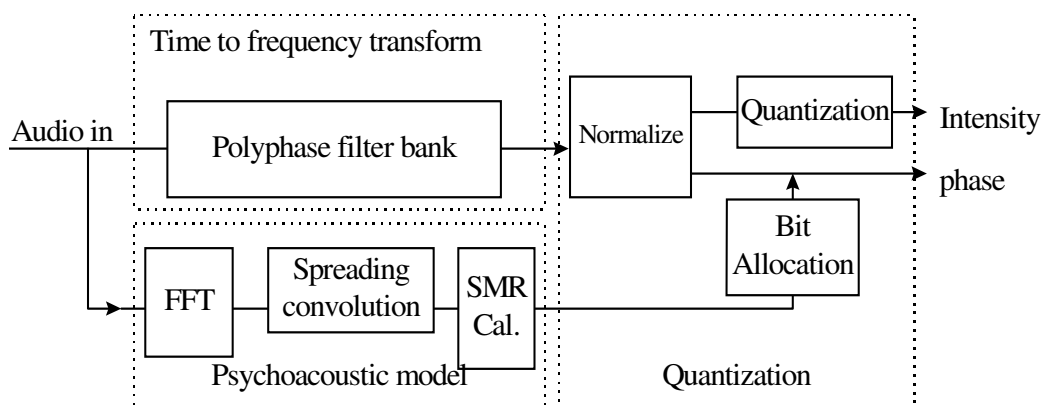


Fig. 3.1 The Structure of the FFT-based MPEG Encoder

In the encoding process of MPEG, the 1024-point Fourier transform (FT) has been used by psychoacoustic models to analyze the frequency components in the

1152 samples of one frame. If the conventional real-data fast FT (FFT) [19] has been adopted for implementing the FT, the complexity has an order of $(4*256*\log(512))$. Such a complexity leads to high implementation cost for real-time applications.

This section presents a new design named the hybrid filterbank to replace the FT. The hybrid filterbank can be integrated with the psychoacoustic models and provides a much lower complexity than the FT. Also, this section shows that the hybrid filter is more suitable for the stereo coding and hence can provide a better quality for the intensity stereo coding, which is the key technology for the MPEG-1 to achieve near transparent quality lower than 96x2 kbits for two stereo channels.

This rest of this section is organized as follows: Section 3.2.1 illustrates the design of hybrid filterbanks. The hybrid filterbank has problems in the phase shift and the aliasing components arising from the decimation in the 1st level filterbank. Section 3.2.2 provides the method to solve the two problems. Sections 3.2.3, 3.2.4, and 3.2.5 consider the complexity and the integration of the hybrid filterbanks with the psychoacoustic models in MPEG. Section 3.2.6 evaluates the design through spectrum analysis, subjective measure, and objective measure to show the feasibility of the hybrid filterbank.

3.2.1 Filter response in hybrid filterbanks

The motivation of the hybrid filterbanks can be considered from the two frequency analyzers in the time-frequency mapper and the psychoacoustic model. The MPEG has adopted a 32-band polyphase filterbank that can provide

a frequency resolution $\pi/32$ with sidelobe attenuation 96 dB while the FT with Hann window a resolution $\pi/512$ with attenuation 32 dB. The approach of the hybrid filterbank is to cascade another filterbank, named the second (2nd) level filterbank, to the output of the original polyphase filterbank, named the first (1st) level filterbank, to achieve a high frequency resolution. The block diagram of the hybrid filterbank is shown in Fig. 3.2.

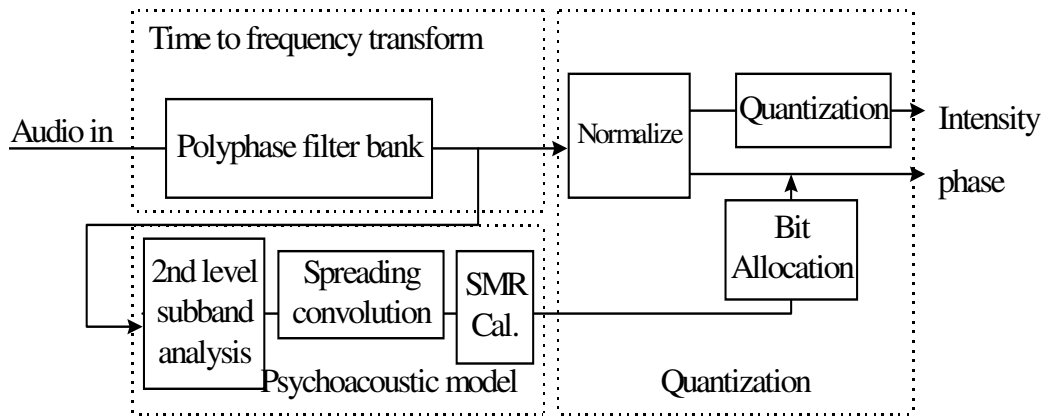


Fig. 3.2 Structure of MPEG encoder based on the hybrid filterbanks

Fig. 3.3 shows the detailed structure of the hybrid filterbank. The structure adopts a 16-band filterbank based on the time domain aliasing cancellation (TDAC) filterbank [30] for each band of the 1st level filter bank to achieve a frequency resolution as high as the FT. The input-output relation of the TDAC filterbank is

$$X_i(k) = \sum_{n=0}^{N-1} h(n)x_i(n) \cos\left[\frac{\pi}{2N}(2n+1+\frac{N}{2})(2k+1)\right] \text{ for } 0 \leq k \leq \frac{N}{2}-1$$

(3.1)

where $x_i(n)$ is the n th output of the band i from the 1st level polyphase filterbank, $X_i(k)$ is the corresponding output of the 2nd level filterbank and $h(n)$ is the window function deciding the band selectivity in the 2nd level filterbank. To

achieve a frequency resolution $\pi/512$ the same as the FT, the value of N is set to 32. Also, to have a frequency selectivity the same as the FT, we select the window function

$$h(n) = \sin\left(\frac{\pi}{N}\left(n + \frac{1}{2}\right)\right) \quad \text{for } n = 0, \dots, N-1 \quad (3.2)$$

which has a sidelobe attenuation 24 dB as shown in Fig. 3.4. The function has the property

$$h(n)^2 + h\left(n + \frac{N}{2}\right)^2 = 1 \quad \text{for } 0 \leq n \leq \frac{N}{2} - 1 \quad (3.3)$$

which is a necessary condition leading to the perfect reconstruction filterbanks [38]. Substituting (3.2) into (3.1) yields

$$X_i(k) = \sum_{n=0}^{N-1} \sin\left(\frac{\pi}{N}\left(n + \frac{1}{2}\right)\right) x_i(n) \cos\left(\frac{\pi}{2N}\left(2n+1 + \frac{N}{2}\right)(2k+1)\right) \quad \text{for } k = 0 \text{ to } \frac{N}{2} - 1 \quad (3.4)$$

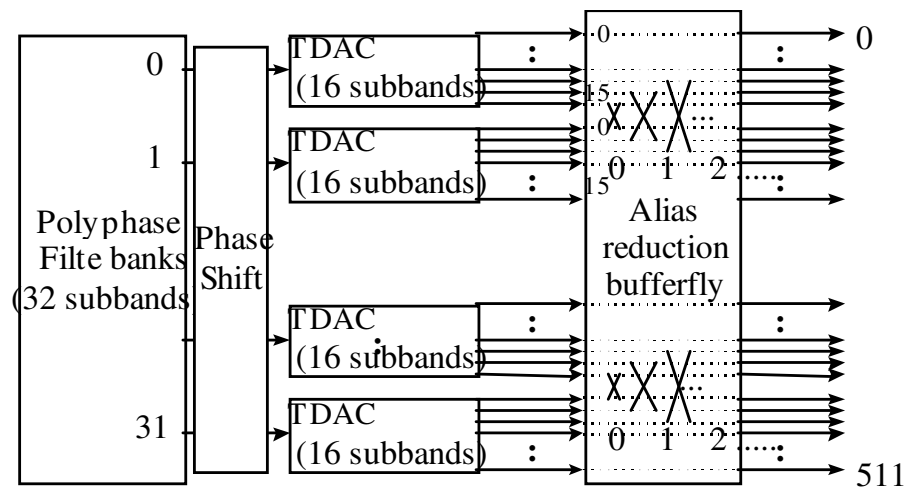


Fig. 3.3 Detailed structure of the hybrid filterbank

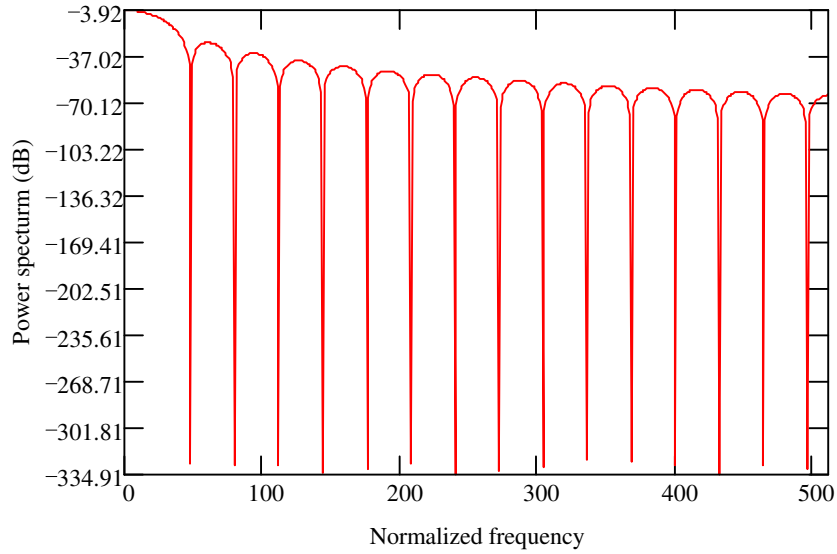


Fig. 3.4 Power spectrum of the 2nd level filterbank

3.2.2 Phase shifter & alias reduction

As mentioned in [1] and [32], the hybrid filterbank has problems in the phase shift and the aliasing components arising from the 1st level filterbank. We follow the similar concept in [1] and [32] to design a phase shifter and an alias reduction butterfly to solve these two problems.

Due to the decimation operation implied in the 1st level filterbank, the 1st filterbank has a phase shift π in the odd-indexed subbands. The phase shift causes a reversed spectrum for the subband. If further spectral analysis is needed to achieve higher frequency resolution, this shift should be corrected. This phase shift can be corrected by multiplying $(-1)^n$ to the subband signal in the odd-indexed subbands; that is

$$\begin{aligned}
X_i(k) = & \begin{cases} \sum_{n=0}^{N-1} (-1)^n \sin\left(\frac{\pi}{N}\left(n + \frac{1}{2}\right)\right) x_i(n) \cos\left(\frac{\pi}{2N}\left(2n + 1 + \frac{N}{2}\right)(2k + 1)\right) & \text{for even } i \\ \sum_{n=0}^{N-1} \sin\left(\frac{\pi}{N}\left(n + \frac{1}{2}\right)\right) x_i(n) \cos\left(\frac{\pi}{2N}\left(2n + 1 + \frac{N}{2}\right)(2k + 1)\right) & \text{for odd } i \end{cases} \\
& \text{for } k = 0 \text{ to } \frac{N}{2} - 1
\end{aligned} \tag{3.5}$$

where odd/even stands for odd/even indexed subband of 1st level filterbank. The phase shifter can be combined into window function to avoid computation burden.

It has been well known that the decimation operation leads to aliasing and there are decimation in the hybrid filterbanks. The aliasing effects indicate a many-to-one merging between the input frequency and output frequency of filterbanks, and hence lead to the difficulty distinguishing the “many” frequency components from the “one” frequency component. The merged frequencies and the corresponding merging weights are decided by the filter bandwidth and the magnitude response of the filter in filterbanks. For the filterbank designed in last section, since that the sidelobe attenuation is around 24 dB, the aliasing term of the frequency in a filter band can be reasonably approximated by the frequency components from the nearest neighboring band. For the hybrid filterbank design in Fig. 3.3, aliasing arises from both the 1st filterbanks and the 2nd filterbanks. The aliasing terms in the 1st level filterbank lead to the merging of frequencies with distance as far as $\pi/32$ while that in the 2nd level filterbank $\pi/512$. Since that the psychoacoustic models in MPEG needs a frequency resolution $\pi/512$, the aliasing terms from the 1st level filterbank should be suitably corrected to

increase the frequency resolution.

Fig. 3.5 shows the frequency responses for the two neighboring filters in the 1st level filterbank before decimation. The lattice lines in Fig. 3.5 show the resolution boundary for the 2nd level filter bands. The cross lines in Fig. 3.5 shows the merged bands from the decimation in the 1st level filterbank.

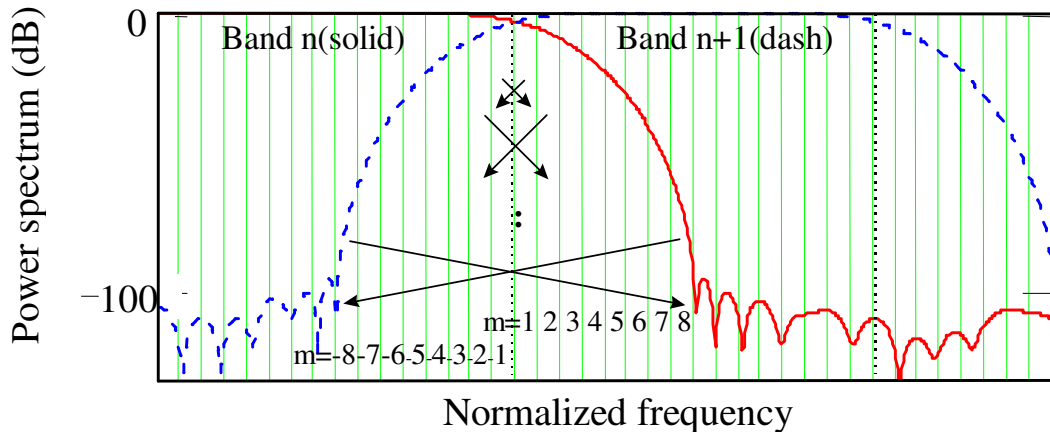


Fig. 3.5 Alias in neighboring subbands

Edler [1] has designed the butterfly structure in Fig. 3.6 to ease the aliasing errors in hybrid filterbanks. The hybrid structure in Fig. 3.3 has included the butterfly structure to compensate the aliasing terms. The butterfly operation is

$$\begin{aligned}
 u_i &= d_m (r_i - c_m r_j) \quad \text{with } i = 16 \cdot k - 1 - m \\
 u_j &= d_m (r_j + c_m r_i) \quad \text{with } j = 16 \cdot k + m \\
 &\text{with } d_m = 1 / \sqrt{1 + c_m^2}, \quad -N/2 \leq m \leq -1
 \end{aligned}$$

(3.6)

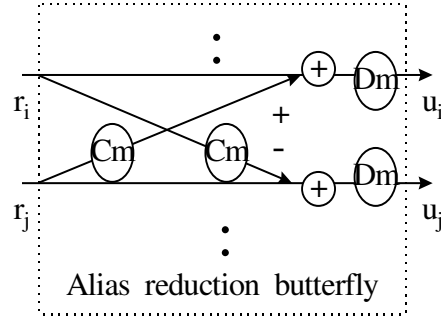


Fig. 3.6 Structure of alias reduction butterfly

For the bands other than those labeled as $m=-1$ and 1 , the weighting factors are calculated using the ratio between the filter response energy in the signal band and that of the aliasing band:

$$c_m = \sqrt{\frac{\text{Energy of alias band}}{\text{Energy of signal band } m}} = \sqrt{\frac{\int_{\text{aliasing band}} |H(\omega)|^2 d\omega}{\int_{\text{signalband}} |H(\omega)|^2 \omega}} \quad (3.7)$$

where $H(\)$ is the frequency response of one filter in the 1st filterbank.

However, the compensation should be modified for the bands labeled as $m=-1$ and 1 . As described above, there are aliasing from the 2nd level filterbank. For example, the band labeled as $m=2$ have aliasing terms from the band labeled as $m=1$ and $m=3$. However, the aliasing terms for $m=-1$ and $m=1$ are only from the band $m=-2$ and $m=2$, respectively. To take the special effect into the butterfly, the weighting factors for $m=-1, 1$ are calculated as

$$c_1 \text{ or } c_{-1} = \sqrt{\frac{\text{Energy of alias band } (1-r)}{\text{Energy of signal band } m}} \quad (3.8)$$

where γ is the ratio between the filter response energy of the signal and the aliasing terms in the 2nd level filterbank. Table 3.2 summarizes the values of the

weighting factors.

m	c_m	d_m
-1	-0.56859	0.86930
-2	-0.49539	0.89607
-3	-0.28182	0.96251
-4	-0.14189	0.99008
-5	-0.05942	0.99824
-6	-0.01952	0.99981
-7	-0.00429	0.99824
-8	-0.00049	1.00000

Table 3.2 Eight weighting factors of alias reduction butterfly.

3.2.3 Complexity analysis

The substitution of the hybrid structure for the FT in the psychoacoustic models of MPEG provides two advantages in complexity. First, since that the two frequency analyzers in Fig. 3.1 can be merged into the hybrid structure in Fig. 3.2, the complexity can be reduced. The second advantage in complexity is from the flexible tuning of frequency resolution in hybrid structure for the different perceptual resolution. If the perceptual resolution (which is the bandwidth of the critical band) is considered in Fig. 3.7, only 12 TDAC filterbanks with alias reduction butterfly structures are required for low frequency range.

Table 3.3 shows the complexity of the hybrid structure compared with the FFT. The 1024-point real-data FFT requires $256 \cdot \log(512)$ complex multiplications and $512 \cdot \log(512)$ complex additions with Hann window of 512 multiplications, while 32 2^{nd} level TDAC filterbanks with the 6 aliasing cancellation butterfly structures require only an order of $32(16 \cdot \log 32 + 32$

+6*2*2) when the fast algorithm of the TDAC filterbank [42] is applied. Further reduction from the perceptual resolution can reduce the complexity as indicated in row 4 of Table 3.3.

Algorithms of frequency mapping in psychoacoustic model	# of multiplications per 1152 samples	# of additions per 1152 samples
1024 pt. FFT (real FFT) + Hann window	$4*256*\log(512)+512=9728$	$2*256*\log(512)+2*512*\log(512)=9216$
32 (32 pt. TDAC filterbank + window)	$32*16*\log(32)+32*32=3584$	$32*32*\log(32)=5120$
32 (32 pt. TDAC filterbank + window + Alias cancellation)	$3584+32*6*2*2=4352$	$5120+32*6*2=5504$
12 (TDAC + window + Alias cancellation) + critical bands	$12/32*(4352)=1632$	$12/32*(5504)=2064$

Table 3.3 Complexity comparison between FFT and hybrid filterbank.

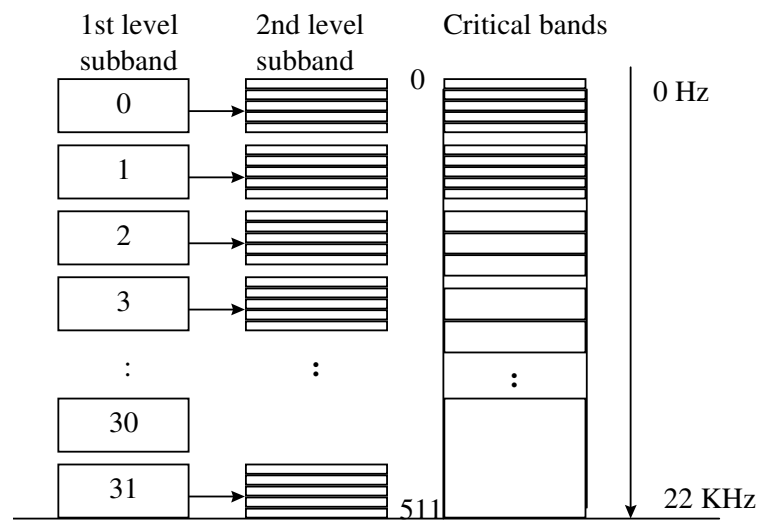


Fig. 3.7 Hybrid filterbank resolution vs. critical band

3.2.4 Cooperating with the intensity mode

The other advantage of the substitution of the hybrid structure for the FT in the psychoacoustic models of MPEG is on the stereo encoding. As mentioned in

section 5.3 or [6], the intensity stereo coding is the key technology for layer 2 in MPEG-1 to achieve a near transparent quality at a bit rate as low as 96x2 kbits for the two stereo channels. However, the original FT analysis has problems in maintaining a consistent frequency analysis with the stereo signals. When the high frequency parts of the two stereo channels are combined into one channel in intensity stereo coding or the coupling scheme shown in Fig. 3.8, the original FT analysis result is not representative for the frequency analysis of the combined channels.

One way to overcome this inconsistent problem is to recalculate the FT analysis and the psychoacoustic model for the two channels somehow based on the combined channels. This recalculation leads to heavy computing load. On the other hand, when these stereo coding schemes are applied, the hybrid structure can be easily tuned to a consistent analysis. Modification of the frequency analysis and the corresponding psychoacoustic model can be performed only on part of the frequency range for the combined channels through the hybrid structure. The hybrid filterbank cooperating with the intensity stereo coding scheme is shown in Fig. 3.9.

3.2.5 Tonality measure

The determination of the tonality of a spectrum line or a band is important in the psychoacoustic model to calculate the sensitivity of the human on the lines or bands. The psychoacoustic model 2 indicated in MPEG draft considers the tonality through a simple prediction calculated in polar coordinates in the complex plane [33]. The tonality detection above is originally designed based on

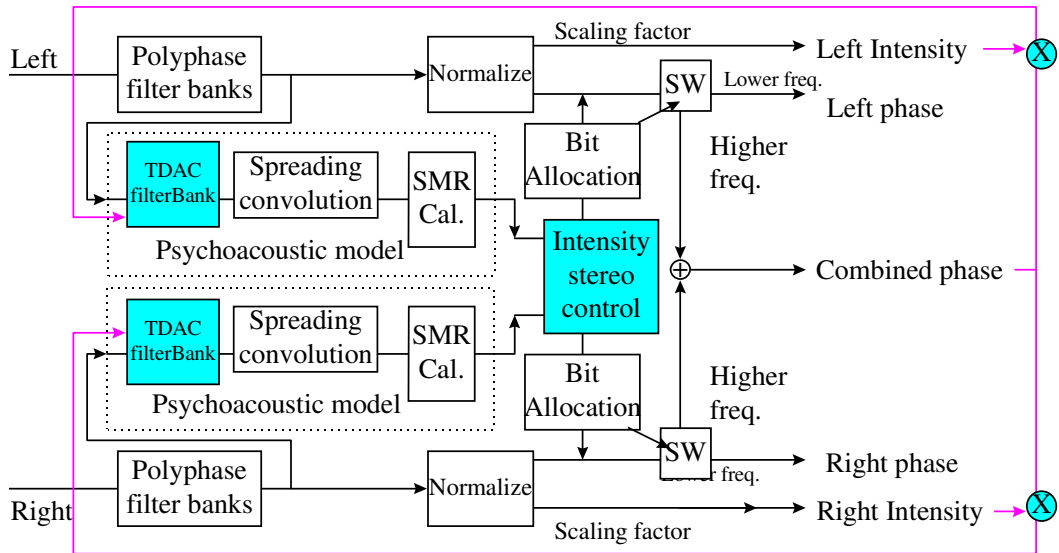


Fig. 3.9 Intensity stereo coding through the hybrid-based psychoacoustic model

3.2.6 Effects of the hybrid filterbank and quality measurement

The effects of the hybrid filterbank and the corresponding modification can be illustrated by comparing the spectrum from the FT and that from the hybrid filterbank. The spectrum analysis for signals with five components at frequencies 400Hz, 800Hz, 1600Hz, 3200Hz and 6400Hz are shown in Fig. 3.10 through the FT (dotted line), the hybrid filterbank without alias reduction (dashed line with 100dB shifting up) and the hybrid filterbank with alias reduction (solid line with 200dB shifting up). The location of each frequency of the hybrid filterbank are almost the same as the one of FT and the alias component of the hybrid filterbank with alias reduction can effectively reduce the aliasing terms.

Several audio segments have been adopted to measure the signal-to-masking ratio [6] from the FT and the various hybrid filterbank. Two of the results are shown in Fig. 3.11 and Fig. 3.12 where the FT is denoted by the solid line, the hybrid filterbank with alias reduction by dotted line, and the

hybrid filterbank with only 12 bands in the 2nd level by dashed line. As the two figures above, Fig. 3.13 shows the signal-to-masking ratio with a 12K Hz high frequency tone to test the performance of hybrid filterbank psychoacoustic model under the pure high frequency tone. The results show that the hybrid filterbank with low complexity can provide a result similar to the FT. Also, informal listening tests show that the audio segments coded by the psychoacoustic model of the FT and the hybrid filterbank are almost imperceptible.

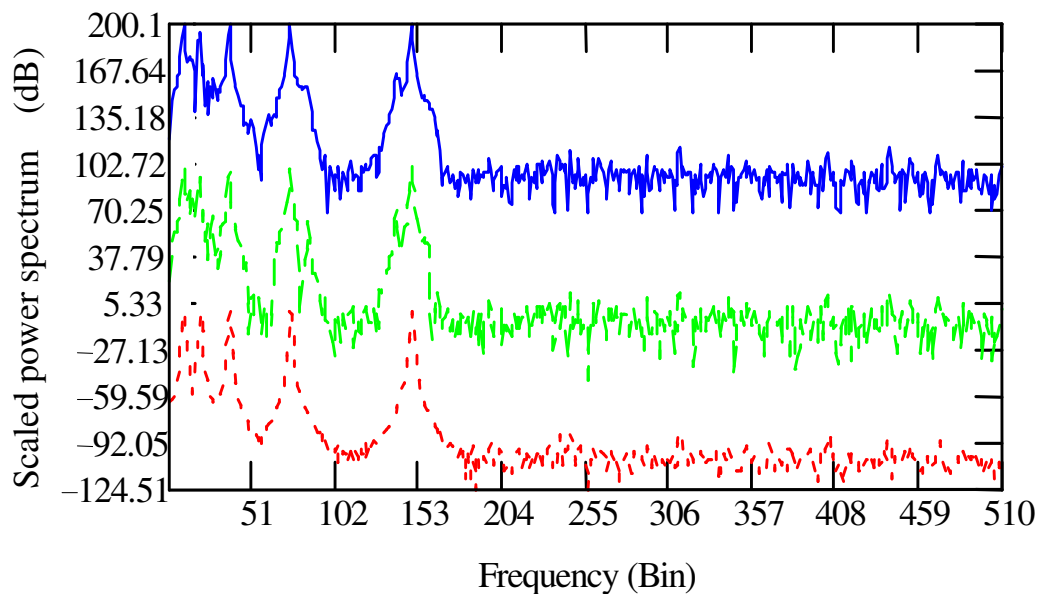


Fig. 3.10 Signal with frequency located at 400Hz, 800Hz, 1600Hz, 3200Hz and 6400Hz analyzed by 1024 pt. FT (dotted line), the hybrid filterbank (dashed line) and the hybrid filterbank with alias reduction butterfly (solid line)

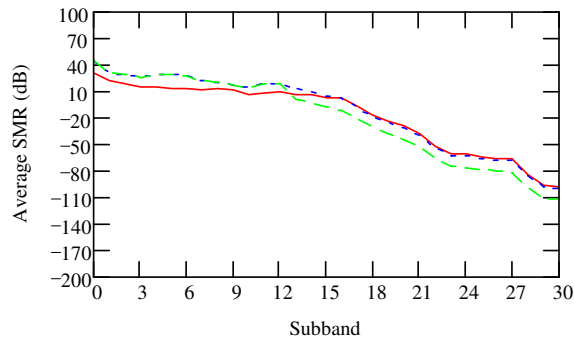


Fig. 3.11 Average signal-to-masking ratio of each subband for female vocal sound.

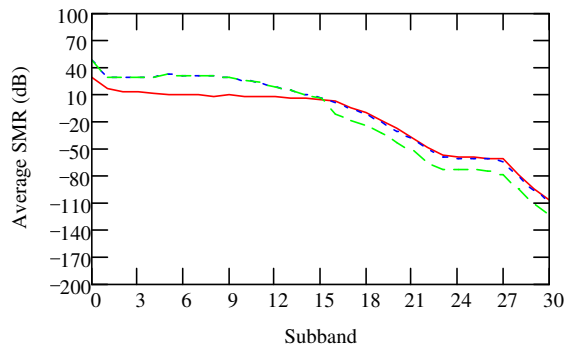


Fig. 3.12 Average signal-to-masking ratio of each subband for classical symphony orchestra.

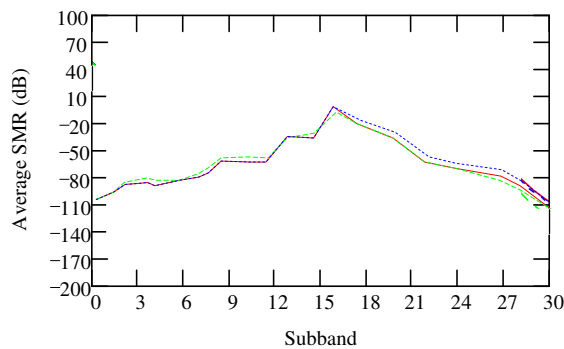


Fig. 3.13 Average signal-to-masking ratio of each subband for high frequency tone at 12 Hz.

3.3 Concluding Remarks

This section has presented a new design named hybrid filterbanks to replace the FT adopted in the psychoacoustic model suggested in the draft on the MPEG layer 1/2 audio coding. This section has given the means to solve the phase shift and aliasing problems in the hybrid structure. The hybrid filterbank can be well integrated with the psychoacoustic model and provide a much lower complexity than the FT. We have also shown that the hybrid filterbank can cooperate with intensity stereo coding scheme to obtain higher audio quality. Due to the flexibility of the hybrid filterbank, a consistent psychoacoustic model with the intensity stereo coding channel can be obtained with little computation increasing. The hybrid filterbank is tested through spectrum analysis, subjective measure, and objective measure to show the feasibility.

Chapter 4

Fast Bit Allocation Method

Subband and transform coder generate frequency domain decomposition of audio signals. When considering with the knowledge of human hearing, this approach offers the possibility to encode the subband components in a way that minimize the audibility of quantization noise. The quantization noise can be minimized, when subband components are to quantize in different quantizer resolution. The quantizer resolution is increase when more bits are assigned to this transform component. The total bit number for the subband components is fixed by the design of bit rate of audio coder. A bit allocation algorithm dynamically distributes the fixed bit pool over the subband component to make the audible noise minimized.

The bit allocation is aimed to assign suitable parameters to the encoder to achieve the best audio quality under the restricted bit number. Hence control over the quality and the bit numbers are two fundamental requirements for the bit allocation. The complexity of the task depends on the difficulties to have the quality and bit control. For MPEG Layers 1 and 2, both the quality and the bit requirement are controlled by a uniform quantizer. Hence the bit allocation is just to apportion the total number of bits available for the quantization of the subband signals to minimize the audibility of the quantization noise.

For MPEG Layer 3 and MPEG-2 AAC, control over the quality and the bit rate is difficult. This is mainly due to the fact that they both use a non-uniform quantizer whose quantization noise is varied with respect to the input values. In other words, it fails to control the quality by assigning quantizer parameters according to the perceptually allowable noise. In addition, the bit-rate control issue can be examined from the variable length coding used in MPEG Layer 3 and MPEG-2 AAC. The variable length coding assigns variable bit-length to different values, which means that the bits consumed should be obtained from the quantization results, and cannot be from the quantizer parameters alone. Thus, the bit allocation is one of the main tasks leading to the high complexity of the encoder. This chapter presents a new bit allocation method to ease the complexity in section 4.1. We examine the issues through MPEG Layer 3.

For Dolby AC-3, it is also difficult to determine the bit allocation. As mentioned above, AC-3 adapts its range according to the specified exponent strategy. There are 3072 possible strategies for the six blocks in a frame. These strategies affect the temporal resolution and the spectral resolution of the quantization ranges. These encoded exponents also affect the analysis result of the psychoacoustic model, which is a special feature of the hybrid coding in Dolby AC-3. The exponents and the resultant psychoacoustic results determine the quantization results. Hence the intimate relation among the exponents, the psychoacoustic models, and the quantization has led to high complexity in bit allocation. This issues and the solution on the bit allocation in Dolby AC-3 has been analyzed in section 4.2.

In this chapter, new bit allocation algorithms are proposed on the basis of

MPEG and AC-3 coder standard. The bit allocation algorithms will yield to close-form bit allocation equations to minimize the audible noise under a fixed bit rate constraint. The close-form bit allocation equations allow single step bit allocation. Thus, comparing to the iterative bit allocation design of MPEG and AC-3, computing complexity is much lower.

4.1 Introduction

From [44], the perceptual optimal solution for subband bit allocation is the quantized noise for each subband should be a ratio to masking threshold. That is, Noise-to-Mask ratio (NMR) in dB will be a constant for each subband. As shown in Fig. 4.1, the noise energy curve for different bit rate, $\text{Noise}_k\#1$ and $\text{Noise}_k\#2$ are parallel to masking threshold curve Masking_k in dB.

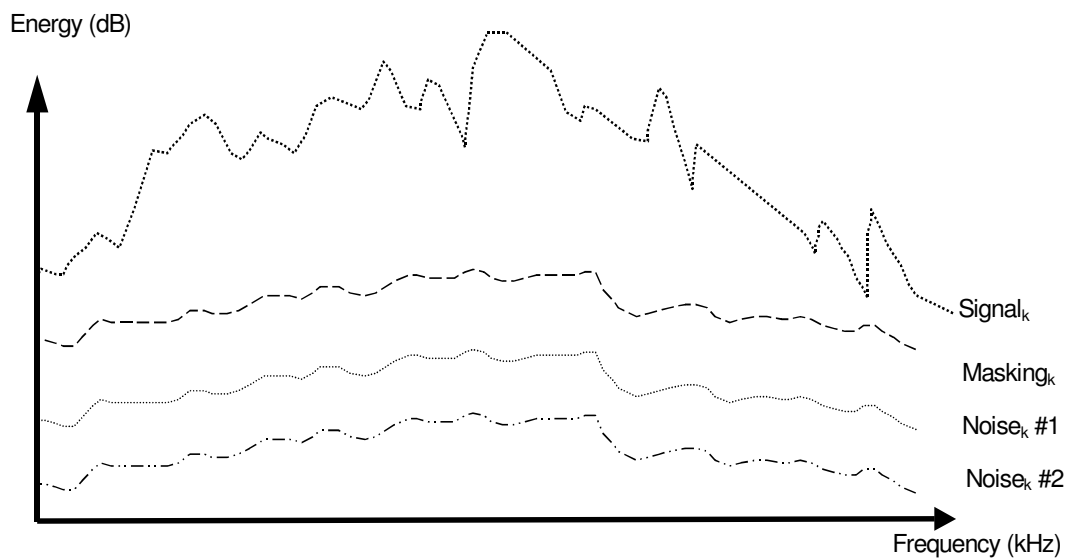


Fig. 4.1 The relation of optimal noise shaping for different bit rate for Noise 1

and Noise 2 with Signal_k and Masking_k.

4.2 Fast Bit Allocation Method in MPEG Layer 3

Before developing the fast bit allocation algorithm, the fast noise estimator is required. The noise estimator will calculate the required bits or step size when given noise required for each subband. In this section, two schemes for the noise estimator are enumerated: analysis-by-synthesis and predictive scheme. First, the straightforward scheme is analysis-by-synthesis (ABS), that is, to calculate iteratively the noise for all step size and choose the step size with nearest calculated noise. Fig. 4.2 shows the relation between quantizer, de-quantizer, and noise estimator. The input signals XR_{sfb} are quantized by the quantizer according to step size Δ_{sfb} . The quantized coefficients are reconstructed by de-quantizer to \tilde{XR}_{sfb} . The noise in the subband can be estimated by the difference of input signal XR_{sfb} and reconstructed signal \tilde{XR}_{sfb} . That is

$e_{sfb} = \sum_{i \in sfb} xr_i - \tilde{xr}_i$. In ABS scheme, to calculate required step size requires a

heavy complexity due to iterative process. Second, predictive scheme for the noise estimation, shortly noise predictor, is to obtain step size by a close-form equation for the relation of step size and noise. The noise predictor formulae have two advantages over the ABS noise estimator: (1) speed up the noise estimation process since noise is estimated without the analysis-by-synthesis noise estimation (2) the noise predictor formulae provide more flexibility to

predict the noise for different step sizes without iteratively calculation of noise for each step size. The noise predictor is faster than the ABS scheme but it also causes prediction error and ABS one not. Table 4.1 shows the noise estimation and bit allocation scheme among the design of MPEG groups and AC-3. In MPEG-1/2 layer 1, 2 and AC-3, uniform quantizers and noise predictors of the quantizers, 6 dB per bit, are used. For MPEG layer 3 and AAC, the non-uniform quantizer and ABS noise estimator are used.

For MPEG layer 1/2 and AC-3, the noise predictor of the uniform quantizer is reviewed from [36]. For MPEG layer 3 and AAC, the ABS noise estimator is used in current standards due to the non-uniform quantizer and Huffman coding. Section 4.2.1 presents a close-form formula for the noise predictor for the non-uniform quantizer of MPEG layer 3.

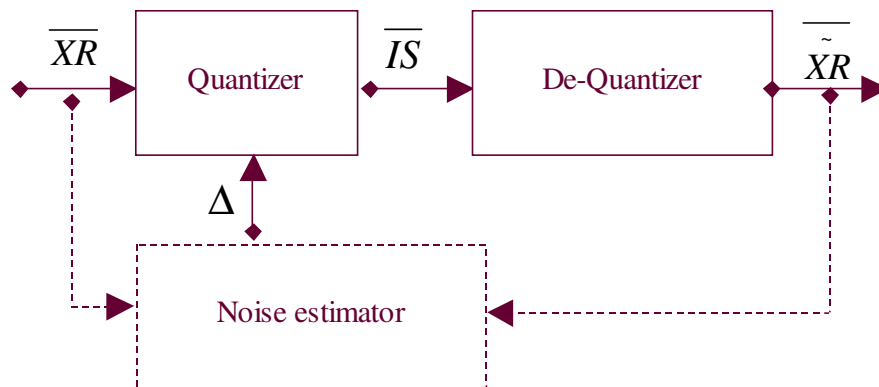


Fig. 4.2 Relation of noise estimator and quantizer in ABS scheme.

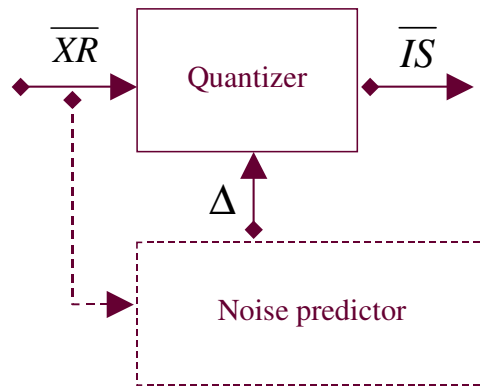


Fig. 4.3 Relation of noise estimator and quantizer in predictor scheme.

Algorithms	Quantizer	Noise estimation scheme	Bit allocation scheme	References
MPEG-1/2 layer 1/2	Uniform	Predictive	Iterative	[35][18]
MPEG-1/2 layer 3	Non-uniform, Huffman coding	ABS	Iterative	[4][31]
MPEG-2 AAC	Non-uniform, Huffman coding	ABS	Iterative	[31]
Dolby AC-3	Uniform	Predictive	Predictive	[21][3][8][9]

Table 4.1 Noise estimation and bit allocation scheme in audio standards

4.2.1 Noise predictor for non-uniform quantizer

For the non-uniform quantizer, it is more complex for the derivation of the noise predictor than the uniform. MPEG layer 3 quantizer is taken as an example. For MPEG AAC quantizer, similar process is applicable. From MPEG layer 3 standard [24], the non-uniform quantizer is done via a power-law function. In this way, larger values are coded with less accuracy, and noise shaping is already built into the quantization process. The quantized values are coded by Huffman coding. To adapt the coding process to different local statistics of the signals, the optimum Huffman table is selected from a number of choices. The

Huffman coding works on pairs and, in quadruples by different frequency location. To get even better adaptation to signal statistics, different Huffman code tables can be selected for different parts of the spectrum.

In the following paragraphs, we will formulate a closed-form equation of the noise predictor for the non-uniform quantizer. Since the variable bit length of the Huffman coding, the Huffman coding process is ignored in the noise prediction process. Thus, from MPEG layer 3 standard [24], the simplified formula for the non-uniform quantizer of layer 3 is given as follows:

$$is_i = \text{int} \left(\frac{xr_i^{\frac{3}{4}}}{\Delta_{sfb}} \right), \text{ where step size } \Delta_{sfb} = 2^{\frac{3}{4}(\text{gain}_{gr} - \text{scale}_{sfb})}. \quad (4.1)$$

By mapping (4.1) and Fig. 4.4, the non-uniform quantizer of MPEG layer 3 is realized by a compressor, a scalar, and a uniform quantizer where the compressor compressing the input signals xr_i by the exponential function of ratio 3/4; thereafter, the scalar scaling by step size Δ_{sfb} for each subband sfb ; the uniform quantizer is realized by a nearest integer function $\text{int}(\cdot)$. Thus, the quantized signals is_i are integer and quantization error ε_i will be in the range of 0 to 1. That is, quantization error of integer quantizer ε_i will be under the condition $|\varepsilon_i| < 1$.

Before further discussing of noise predictor of the non-uniform quantizer (4.1), the steps of simplifying the non-uniform quantizer from the MPEG layer 3 standards to (4.1) are introduced. From MPEG standard [24], the formula of the non-uniform quantizer can be expressed as

$$is_i = \text{int} \left(xr_i 2^{scale_{sfb} - gain_{gr} - 0.0946} \right)^{\frac{3}{4}}, \quad (4.2)$$

where scale factor $scale_{sfb} = 1/2(1 + scalefac_scale)(scalefac_{sfb} + preflag_{gr} \cdot pretab_{sfb})$ for each band sfb; $scalefac_scale$ is 0 or 1, $scalefac_{sfb}$ is in the range of 0~15, and the pre-amplified flag $preflag_{gr} \cdot pretab_{sfb}$; global gain $gain_{gr} = 1/2(global_gain_{gr} - 210)$ for each granule of MPEG layer 3 frame. By ignoring 0.0946, the step size can be obtained by

$$\begin{aligned} is_i &= \text{int} \left(xr_i 2^{scale_{sfb} - gain_{gr}} \right)^{\frac{3}{4}} \\ &= \text{int} \left(xr_i^{\frac{3}{4}} 2^{\frac{3}{4}(scale_{sfb} - gain_{gr})} \right) \\ &= \text{int} \left(\frac{xr_i^{\frac{3}{4}}}{\Delta_{sfb}} \right) \end{aligned}$$

where step size $\Delta_{sfb} = 2^{\frac{3}{4}(gain_{gr} - scale_{sfb})}$ (4.3)

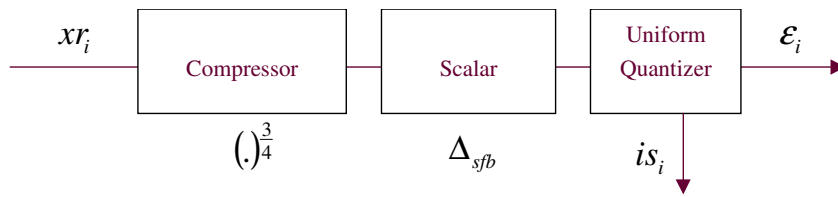


Fig. 4.4 Non-uniform quantizer in MPEG layer 3, where step size as (4.3),

$$\Delta_{sfb} = 2^{\frac{3}{4}(gain_{gr} - scale_{sfb})}$$

Now, we will derivate the noise prediction formulae for the quantizer.

From Fig. 4.4, we can have the input signal xr_i and reconstructed signal \tilde{xr}_i in the

following two formulae.

$$xr_i = \left((is_i + \varepsilon_i) \Delta_{sfb} \right)^{\frac{4}{3}} \quad (4.4)$$

and

$$\tilde{x}r_i = \left(is_i \Delta_{sfb} \right)^{\frac{4}{3}}. \quad (4.5)$$

The quantization error of the non-uniform quantizer e_i equal to the difference of input signal xr_i and $\tilde{x}r_i$. We will have

$$e_i = xr_i - \tilde{x}r_i = \left((is_i + \varepsilon_i) \Delta_{sfb} \right)^{\frac{4}{3}} - \left(is_i \Delta_{sfb} \right)^{\frac{4}{3}} = \left(1 + is_i^{-1} \varepsilon_i \right)^{\frac{4}{3}} is_i^{\frac{4}{3}} \Delta_{sfb}^{\frac{4}{3}} - \left(is_i \Delta_{sfb} \right)^{\frac{4}{3}} \quad (4.6)$$

From (4.6), by the definition of the function $f(\varepsilon_i) = \left(1 + is_i^{-1} \varepsilon_i \right)^{\frac{4}{3}}$, we can have the quantization error in the form of

$$e_i = f(\varepsilon) is_i^{\frac{4}{3}} \Delta_{sfb}^{\frac{4}{3}} - \left(is_i \Delta_{sfb} \right)^{\frac{4}{3}}. \quad (4.7)$$

By Tylor expansion, we can have the first order approximation of $f(\varepsilon) \approx 1 + f'(\varepsilon)\varepsilon$.

$$f'(\varepsilon_i) = \frac{4}{3} \left(1 + is_i^{-1} \varepsilon_i \right)^{\frac{1}{3}} (is_i^{-1}) \approx \frac{4}{3} is_i^{-1} \quad (4.8)$$

We can have

$$f(\varepsilon_i) \approx 1 + f'(\varepsilon)\varepsilon = 1 + \frac{4}{3} is_i^{-1} \varepsilon_i \quad (4.9)$$

From (4.7), (4.8) and (4.9), the quantization error will be

$$e_i = f(\varepsilon_i)is_i^{\frac{4}{3}}\Delta_{sfb}^{\frac{4}{3}} - (is_i\Delta_{sfb})^{\frac{4}{3}} \approx \frac{4}{3}is_i^{\frac{1}{3}}\varepsilon_i\Delta_{sfb}^{\frac{4}{3}} \quad (4.10)$$

From (4.8) and assume quantized signals is_i and quantized error of the uniform quantizer ε_i are independent, we can have the expectation of quantization error of the non-uniform quantizer e_i as the follows:

$$E[e_{sfb}^2] \approx \frac{16}{9}\Delta_{sfb}^{\frac{8}{3}}E[IS_{sfb}^{\frac{2}{3}}\mathcal{E}^2] \approx \frac{16}{9}\Delta_{sfb}^{\frac{8}{3}}E[IS_{sfb}^{\frac{2}{3}}]E[\mathcal{E}_{sfb}^2] \quad (4.11)$$

According to

[36], the quantization error variance of uniform quantizer can be formulated as

$\delta^2 = \Delta^2 / 12$; that is $\delta_{\varepsilon}^2 = E[\mathcal{E}_{sfb}^2] = \frac{1}{12}$, so the formula (4.11) becomes

$$E[e_{sfb}^2] \approx \frac{4}{27}\Delta_{sfb}^{\frac{8}{3}}E[IS_{sfb}^{\frac{2}{3}}] \quad (4.12)$$

By (4.5) and (4.12), the quantization error of the non-uniform quantizer is

$$E[e_{sfb}^2] \approx \frac{4}{27}\Delta_{sfb}^{\frac{8}{3}}E\left[\left(\frac{XR_{sfb}^{\frac{3}{4}}}{\Delta_{sfb}}\right)^{\frac{2}{3}}\right] = \frac{4}{27}\Delta_{sfb}^2E[XR_{sfb}^{\frac{1}{2}}] \quad (4.13)$$

From (4.13), the signal-to-noise ratio can be expressed as

$$SNR(dB) = 10\log_{10}\left(E[XR_{sfb}^2] / \frac{4}{27}\Delta_{sfb}^2E[XR_{sfb}^{\frac{1}{2}}]\right) \quad (4.14)$$

From (4.13), the noise predictor of the non-uniform quantizer depends not only on the step size Δ_{sfb} but also inputs signals XR_{sfb} .

4.2.2 Fast bit allocation for non-uniform quantizer

From section 4.2.2, the noise predictor formulae of uniform and non-uniform quantizer are given. Fast bit allocation can be developed by the formulae. The noise predictor formula for uniform quantizer is widely adopted in current design of audio standard. As shown in Table 4.1, all audio standards using uniform quantizer, MPEG 1/2 layer 1, 2 and AC-3 use the noise predictor formula to speed up the noise estimation process. Several papers [8][9][35][18] propose fast algorithms on the uniform quantizer bit allocation on the basis on the noise predictor formula.

From MPEG standard [24] and related papers [31], the original design of the bit allocation is as following. A global gain that determines the quantization step size and scalefactors that determine the noise-shaping factors for each scalefactor band are applied before actual quantization. The process to find the optimum gain and scalefactors for a given block, bit-rate and output from the perceptual model is usually done by two nested iteration loops in an analysis-by-synthesis way. (1) Inner iteration loop (rate loop): If the number of bits resulting from the coding operation exceeds the number of bits available to code a given block of data, this can be corrected by adjusting the global gain to result in a larger quantization step size, leading to smaller quantized values. This operation is repeated with different quantization step sizes until the resulting bit demand for Huffman coding is small enough. (2) Outer iteration loop (noise control loop): To shape the quantization noise according to the masking threshold, scalefactors are applied to each scalefactor band. If the quantization noise in a given band is found to exceed the masking threshold as supplied by

the perceptual model, the scalefactor for this band is adjusted to reduce the quantization noise. Since achieving a smaller quantization noise requires a larger number of quantization steps and thus a higher bit-rate, the rate adjustment loop has to be repeated every time new scalefactors are used. The two nested loops ensure the demand of bit rate and noise shaping for each subband by iteratively using analysis-by-synthesis noise estimator. A new fast bit allocation algorithm based on the noise predictor formula presented in 4.2.1 is proposed. The new bit allocation also meets the demand of bit rate and noise shaping for each subband by single step prediction.

From

[44], the perceptual optimal solution for subband bit allocation is the quantized noise for each subband should be a ratio to masking threshold Thr_{sfb}^2 . That is the expected noise will be

$$E[e_{sfb}^2] = c \cdot Thr_{sfb}^2 \tag{4.15}$$

where c is a constant varied with bit rate. According to (4.13), substituting (4.15) into (4.13), we can obtain

$$E[e_{sfb}^2] = c \cdot Thr_{sfb}^2 \approx \frac{4}{27} \Delta_{sfb}^2 E[XR_{sfb}^{1/2}],$$

or in the form of

$$\Delta_{sfb}^2 \approx \frac{27}{4} c \cdot Thr_{sfb}^2 / E[XR_{sfb}^{1/2}] \tag{4.16}$$

According to (4.1) for the step size, we can have

$$\Delta_{sfb}^2 = 2^{\frac{3}{2}(gain_{gr} - scale_{sfb})} \approx \frac{27}{4} c \cdot Thr_{sfb}^2 / E[XR_{sfb}^{1/2}] \quad (4.17)$$

From (4.17), the difference of global gain and scalefactor is approximate to

$$gain_{gr} - scale_{sfb} \approx \frac{2}{3} \log_2 \frac{27}{4} c \cdot Thr_{sfb}^2 / E[XR_{sfb}^{1/2}],$$

or in the form of

$$gain_{gr} - scale_{sfb} = \frac{2}{3} (\log_2^{\frac{27}{4}} + \log_2^c + \log_2^{Thr_{sfb}^2} - \log_2^{E[XR_{sfb}^{1/2}]}) \quad (4.18)$$

Since scalefactor $scale_{sfb}$ is in the range of 0~31. To obtain scalefactor for each subband, let $gain'_{gr} = \text{Max}_{sfb} \{ gain_{gr} - scale_{sfb} \}$. The scalefactor for each subband will be

$$scale'_{sfb} = gain'_{gr} - scale_{sfb}. \quad (4.19)$$

Reordering the formula (4.18) and substituting the resulting scalefactor from (4.19) yields

$$gain_{gr} - scale'_{sfb} = \frac{2}{3} (\log_2^{\frac{27}{4}} + \log_2^c) + \frac{2}{3} (\log_2^{Thr_{sfb}^2} - \log_2^{E[XR_{sfb}^{1/2}]}). \quad (4.20)$$

From (4.20), the global gain $gain_{gr}$ varies with the bit rate related constant c and scalefactor $scale_{sfb}$ varies for each subband according to the masking threshold Thr_{sfb}^2 and input signals $E[XR_{sfb}^{1/2}]$.

The experiment results are given for the fast allocation. In Fig. 4.5, the Noise curve for the original MPEG bit allocation and proposed algorithm are

compared with the masking curve. The result shows that the new proposed algorithm will cause the noise curve more parallel to the noise curve provided by original MPEG. Table 4.2 show the performance of the new proposed algorithm. The speedup for the bit allocation is almost ten speed of the original one.

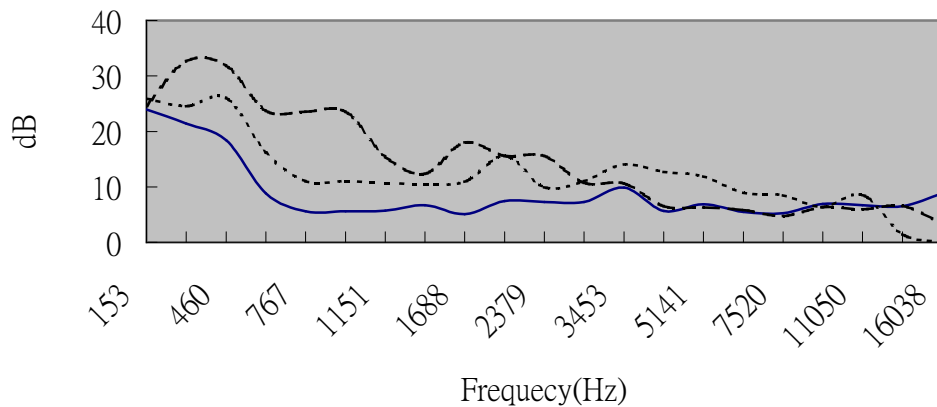


Fig. 4.5 Signal-to-masking ratio (SMR) and signal-to-noise ratio (SNR) curve. Solid line is the SMR value; long slash line is the SNR value for original bit allocation; short slash line is the SNR value for new bit allocation algorithm under 128 kbit/s.

Testing material	MPEG-1	Proposed algorithm
9_1	37.66194	5.86177
9_2	34.81273	4.199569
9_3	40.94803	5.726937
butter1	40.03659	7.596248
coco	34.66328	3.814438

dance1	35.2703	7.615006
flute	33.86285	9.401907
harp	35.83702	5.383456
hat1	52.23678	11.73032
heart1	24.4059	3.601169
man	35.17282	3.923124
memory	44.34348	5.908672
mist	33.20018	6.535363
music	39.83675	8.446721
point1	30.96771	6.527983
summer	41.64852	3.866236
tsai	34.75923	5.072571
winter	37.41175	4.886839
Woman1	38.71363	7.504144

Table 4.2 Average iteration number for different testing material for the proposed and MPEG bit allocation algorithm

4.3 Fast Bit Allocation Method in AC-3

4.3.1 Addressed issues

The Dolby AC-3 [27] is currently the audio standards for the United States Grand Alliance HDTV system audio coding standard and widely adopted for DVD films. The Dolby AC-3 encoding process can be illustrated in Fig. 4.6.

The audio sequences are transformed into a domain referred to as spectral domain. Each spectral line in the spectral domain is represented as floating point consisting of exponent and mantissa. The exponents are encoded by suitable coding strategy and fed into psychoacoustic model. The psychoacoustic model calculates the perceptual resolution according to the encoded exponents and the proper perceptual parameters. Finally, the information of the perceptual resolution and the available bits are used to decide the appropriate quantization manner to quantize the mantissa of the spectral lines under restricted bits. The bit allocation process is to determine the suitable exponent coding strategies, the proper perceptual parameters, and the appropriate quantization manners in the encoding process with restricted bit number.

Consider the exponent coding process in Fig. 4.6. The difficulties of the exponent coding are on the efficient search for the large number of strategies and the criterion deciding the best strategies. In AC-3, it provides four exponent coding strategies for each audio block referred to as D15, D25, D45 and REUSE. Except for the first audio block, the remaining audio blocks can use the REUSE coding strategy. Hence, there are $3 \times 4 \times 4 \times 4 \times 4 = 3072$ possible strategies for the six blocks in a frame. The search space is large and there needs an efficient search method. Furthermore, even an exhaustive search is executed there needs a criterion for selecting the strategies. Since that there is no analytic relation between the final audio quality and these exponent strategies, an optimum solution is to follow an analysis-by-synthesis method. That is, all the candidate strategies for exponent coding are tried and hence provide the necessary information for the remaining encoding process. Then, the optimal coding

strategy is selected from the associated coded or synthesis audio having the best quality. However, the complexity for the process is again too high to be practical. In this section, we propose a selection criterion and an efficient search method for exponent strategies.

Consider next the psychoacoustic model in Fig. 4.6. The psychoacoustic model calculates the perceptual resolution according to the encoded exponents and perceptual parameters. The difficulty of the process is the way to adapt the perceptual parameters to the current audio content. The AC-3 standard draft suggests that the perceptual parameters are fixed to simplify the complexity of bit allocation process. However, for low bit rate system such as that below 64 Kbit/s for a channel, these parameters are quite critical for audio quality. This chapter presents the method to adapt the parameters to the audio contents.

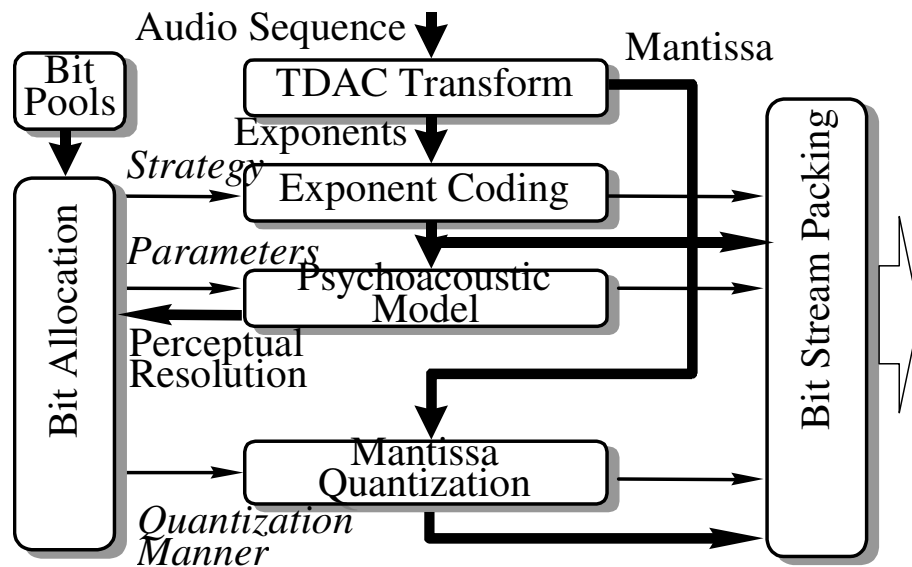


Fig. 4.6 Encoding process for AC-3.

The third difficulty is on the mantissa quantization. The major problem arising from the mantissa quantization process is on the efficient search for the

value of quantization parameter provided by the AC-3 to fit the available bits. In AC-3, the mantissa quantization process is to quantize the mantissa of each spectral line according to the perceptual resolution and the values of quantization parameter. There are 1024 selections for the parameter in AC-3 and a vehicle searching for the optimal value fitting the restricted bits is needed. The problem is that there is no direct relation among the values of the parameter, the perceptual resolution, and the available bits. That is, there is no way finding the suitable quantization value directly from the perceptual resolution and the available bits. This section proposes the efficient algorithm for searching the optimal value of the quantization parameter in AC-3.

The rest of this section is organized as follows: Section 4.3.2 illustrates the efficient searching algorithm and selection criteria for exponent coding process. Section 4.3.3 provides the method to adapt the perceptual parameters to current audio content and also gives the efficient searching algorithm for the quantization parameter. Section 4.3.4 shows the experiment results. Section 4.3.5 gives a brief conclusion.

4.3.2 Exponent coding method

In AC-3, each spectral line is represented by an exponent and a mantissa. All the exponents are coded by the exponent coding process. The coding strategies available in AC-3 are referred to as D15, D25, D45 and REUSE. The coding strategy D15 provides the finest frequency resolution and hence requires a large number of bits. On the contrary, the strategy D45 gives the coarsest frequency resolution and hence consumes a less number of bits. Especially, the

strategy REUSE indicates that the exponents of current block are the same as the previous block and hence there is no bit requirement for the exponent of current audio block.

As described in last section, two difficulties on the exponent coding are the large combinational space of the exponent coding strategies and the selection criterion. This section proposes a selection criterion and the associated efficient search method for the exponent strategies. The block diagram of the exponent coding process is illustrated in Fig. 4.7. The process consists of three steps. First, the available bits of the exponents are determined from the current bit rate. A ratio of 20% of the overall bit rate has been adopted to select the exponent strategies. The ratio has been determined through immense experiments. On the ratio, the second step is to list all the exponent strategies that consume a bit number less than the available bits. For music sequence adopting a fixed frame rate, the candidates are fixed and will not vary with frames. Finally, all the candidates are used to encode the exponents. On all the associated encoded exponents, the strategy that minimizes the error criterion is selected. The error criterion is listed as follows:

$$E = \sum_{k=0}^5 \sum_{b=0}^{255} [\exp_o(k,b) - \exp(k,b)] \quad (4.21)$$

where $\exp_o(k,b)$ is the original exponent of block k and spectral bin b before encoding, and $\exp(k,b)$ is the corresponding exponent encoded by a candidate strategy. In a frame defined by AC-3, there are six blocks and 256 spectral bins in a block. The criterion is reasonable in the sense that the formula indicates the error between the coded and the original exponents. The overall process can find

the best fitted exponent strategy under the bit rate constraint.

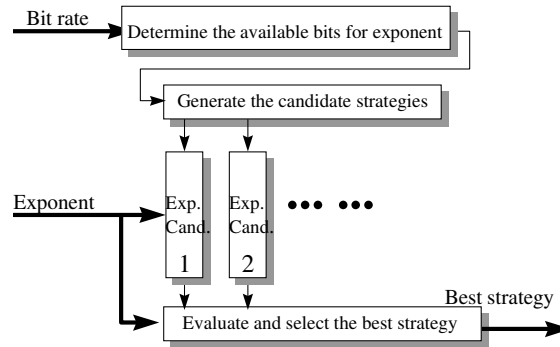


Fig. 4.7 Block diagram of exponent coding process.

4.3.3 Perceptual parameters

In audio coding, the psychoacoustic model gives the information on the perceptual resolution of audio signals. The perceptual resolution is the key information to compress an audio sequence without losing audio quality. The perceptual resolution is calculated from the masking effects of signals. Masking effects demonstrate the perceptual resolution of spectral lines when various types of audio contents exist. Especially, two types of masking effects are considered in audio compression. The first type is the masking effects from the existing of narrow band noise. The other is the masking effects from tonal signals. The two types of masking result in different masking effects and hence different perceptual resolution. This section presents a method to detect the two types of masking effects from the audio exponents. The parameters in the psychoacoustic model of AC-3 are determined according to the detection results.

The psychoacoustic model in AC-3 calculates the masking threshold from the following three steps: First, the encoded exponents are transformed into power spectral density (PSD) through

$$\text{psd}(k,b) = 3072 - \exp(k,b) * 128 \quad (4.22)$$

Then, the bins of the PSD are combined into bands according to the perceptual bandwidth. At low frequencies, the band size is 1, and at high frequencies the band size is 16. Third, the masking threshold of a band is computed by summing the masking effects from other bands. The masking effect of a band from the signals in other band is illustrated through the spreading function in Fig. 4.8. For a signal existing at band i with energy E , the spreading function indicates the resultant masking threshold of the bands above band i . The spreading function is approximated by two curves: a fast decaying curve and a slow decaying curve. The fast gain is the signal-to-mask ratio, that is the ratio between the energy of the masking sound and the masking threshold in band i . The gain can be chosen according to the audio contents. In AC-3 standard draft, the value is fixed and selected as -30dB. However, in [26] the Voluminous experiments demonstrate that the corresponding parameter is selected from -10dB to -20dB for tonal signal and -5dB to -10dB for narrow band noise. This section shows the method to determine the values of the fast gain.

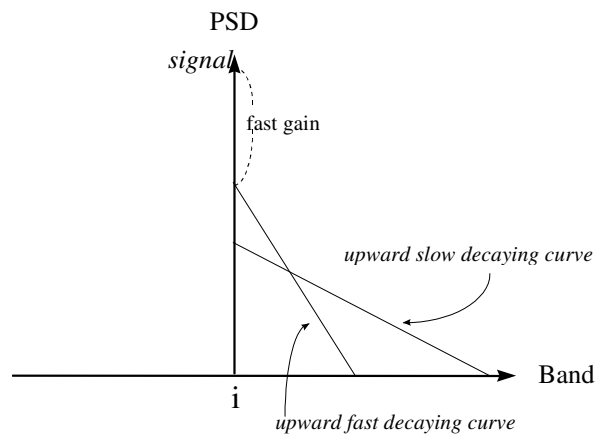


Fig. 4.8 Modeling spreading function.

Due to the limit on AC-3, the fast gain is transmitted once per audio block rather than for each spectral bin. Hence, a simple method for the parameter selection is that the parameters are adopted according to the information of audio block rather than single spectral line. That is, if the audio block is tone-like, the conservative value -30dB is retained. On the contrary, if the audio block is noise-like, the value -10dB is selected. However, the difficulty is the tonality measure for an audio block.

Two properties of the tonal signals are the spectral peaks and the spectral similarity between blocks. Since that the exponent strategies decided in (4.21) has considered both the spectral and temporal similarity, the tonality can be selected directly through the exponent strategies. Since that the tonal signal has higher spectral peak than other frequency components near it, if the audio block is tone-like, it implies that the exponents of the block have to be encoded through the highest spectral resolution strategy, that is the D15 mode. In addition, since that the tonal signal can be determined from the likeness of a spectrum band through several audio blocks, those blocks using REUSE are also tone-like. Furthermore, if the exponent strategy is D45, the audio block is considered to be a noise-like block.

Now the information of the exponent coding process is used to decide the psychoacoustic parameters. As mentioned above, the perceptual parameters are transmitted once per audio block rather than per spectral bin. Hence, the conservative value of the fast gain is retained. If the result of the exponent coding process gives that the block is in the D15 mode and the following blocks are in the REUSE mode, the block is tone-like and the fast gain is selected as

-24dB. If the exponent strategy is D45, the associated block is noise-like and the fast gain is selected as -12dB. For the D25 mode, the average value -18dB is adopted.

Consider the flowchart of mantissa quantization shown in Fig. 4.9. The mantissa quantization retrieves the masking threshold $Mask_{bin}$ from the psychoacoustic model. The masking curve is added with the parameters $SNROFFSET$ to produce the noise curve. The signal to noise ratio can then be obtained for each spectral bin. The bit number of the mantissa can then be determined from the ratio of the signals and noise. In the flowchart, the problem is on the selection of the optimal value of $SNROFFSET$. There are 1024 selections for $SNROFFSET$ and a vehicle searching for the optimal value to fit the available bits is required. This section considers the efficient searching algorithm for the values of $SNROFFSET$.

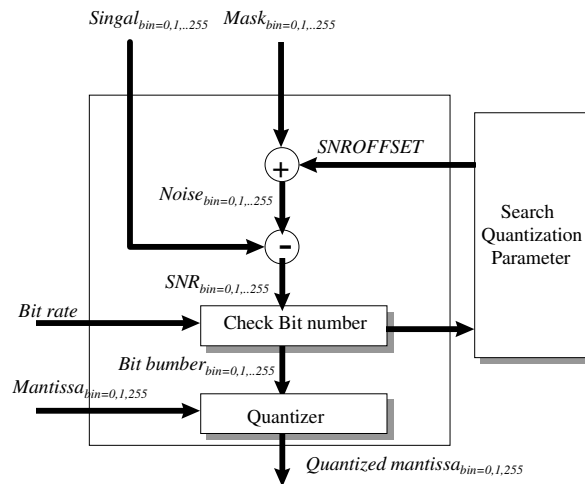


Fig. 4.9 Flowchart of mantissa quantization.

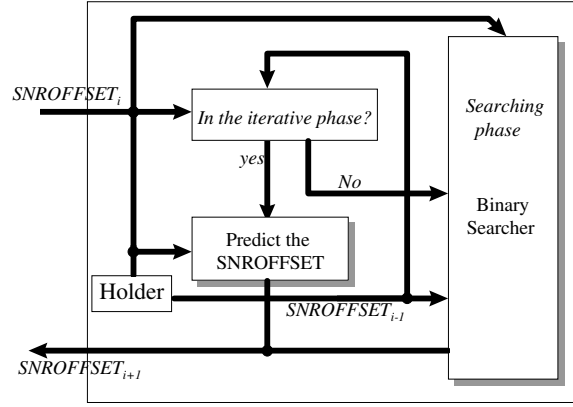


Fig. 4.10 Block diagram of the quantization parameter search.

Since that there are 1024 selections of $SNROFFSET$, therefore, at least ten iterations are needed to find the optimal quantization parameter if the binary searching algorithm is performed. To further reduce the complexity, we propose a new searching algorithm. Our experiments demonstrated that the new algorithm is more efficient than the binary searching algorithm.

The proposed searching algorithm consists of two phases: (1) iterative phase and (2) searching phase. The block diagram of the quantization parameter search is shown in Fig. 4.10. Initially, the proposed searching algorithm is in the iteration phase. In this phase, the quantization parameter, $SNROFFSET$, is predicted in each iteration. The predictive equation is given as follows:

$$SNROFFSET_i = SNROFFSET_{i-1} + \frac{R_{i-1} - R_{av}}{nBIN_{i-1}} \times \mu \quad (4.23)$$

where $SNROFFSET_i$ is the quantization parameter at iteration i , $nBIN_i$ is the number of spectral lines with positive bit number and R_i is the allocated bit number in the i -th iteration. R_{av} is the current available bit number and μ is step size. In our experiments, we choose the step size μ as 128.

In AC-3, the psychoacoustic model is performed on the PSD domain

[3]. The PSD is derived by the encoded exponent expressed in (4.22). Hence, the PSD-decibel has the following relation:

$$128 \text{ units PSD} = 6 \text{ dB} \tag{4.24}$$

From

[36], since that additional one bit resolution increases the signal-to-noise ratio by 6dB for uniform quantizers, the signal-to-noise ratio is increased by 128 units PSD. Therefore, the step size μ is chosen as 128. In the low bit rate system, the symmetric quantizers are often used. In the condition, the step size μ has to be decreased to avoid over-prediction.

The iteration terminates when the following two conditions are met: (a) $R_i \leq R_{av}$, $R_{i-1} > R_{av}$ or (b) $R_{i-1} \leq R_{av}$, $R_i > R_{av}$. The search phase then searches the optimal value from the range between $SNROFFSET_i$ and $SNROFFSET_{i-1}$ by the binary search algorithm. Since that the optimal quantization parameter is bounded by $SNROFFSET_i$ and $SNROFFSET_{i-1}$ which is the sub-region of 0 to 1024, the binary searching algorithm takes less than ten iterations to find the optimal value of $SNROFFSET$.

source	butter	tsai	dance	flute	heart1	memory	second	march	Russian	Chinese
count	5.18	5.81	4.94	4.91	6.02	5.78	6.09	5.40	5.86	4.25

Table 4.3 Average iteration counts per frame.

4.3.4 Experiment results

This section considers the efficiency of the encoding algorithm. In the

following experiments, each audio channel is encoded at the bit rate of 64 Kbit/s with sampling frequency of 44.1 KHz. The bit number of the exponents is 435 in one frame. The exponents coding strategies that consume less than 20% frame bit rate are listed in Table 4.4. The three audio sequences illustrated in Fig. 4.11 can provide a typical example for the experiments. The decided exponent coding strategy also decides the tonality of the block. Fig. 4.11 illustrates three examples of the tonality decision. The decisions are quite consistent with audio contents.

For the experiments on searching the values of the *SNROFFSET*, a total of ten 20 sec stereo audio songs including vocal, symphony, piano and so on are taken as the materials. Table 4.3 lists the average iteration numbers per frame of mantissa quantization for above materials. The iteration numbers demonstrate that the proposed method provides an iteration number much lower than ten that is the iteration counts of binary searches for 1024 values.

(1)	[D15,REUSE,REUSE,REUSE,REUSE,REUSE]
(2)	[D25,REUSE, REUSE,D25,REUSE,REUSE]
(3)	[D25,REUSE,REUSE,D45, REUSE,D45]
(4)	[D25,REUSE,D45,REUSE,D45, REUSE]
(5)	[D45,D45,REUSE,D45, REUSE,D45]

Table 4.4 Candidates of exponent coding strategies.

4.3.5 Remarks

In AC-3 encoder, the bit allocation is quite computation intensive and there is no article analyzing the problem. This section has analyzed the problem and

presented efficient methods of the bit allocation through three aspects: (1) the exponent coding, (2) the psychoacoustic model, and (3) the mantissa quantization. For the exponent coding, the problem is on the selection criterion and the efficient search method for the exponent strategies. For the psychoacoustic models, the difficulty is on the selection of the perceptual parameters adapting to audio contents. For the mantissa quantization, the issue is on the efficient search methods for the optimal value of the quantization parameter. On the three aspects, this section has presented methods to achieve efficient bit allocation.

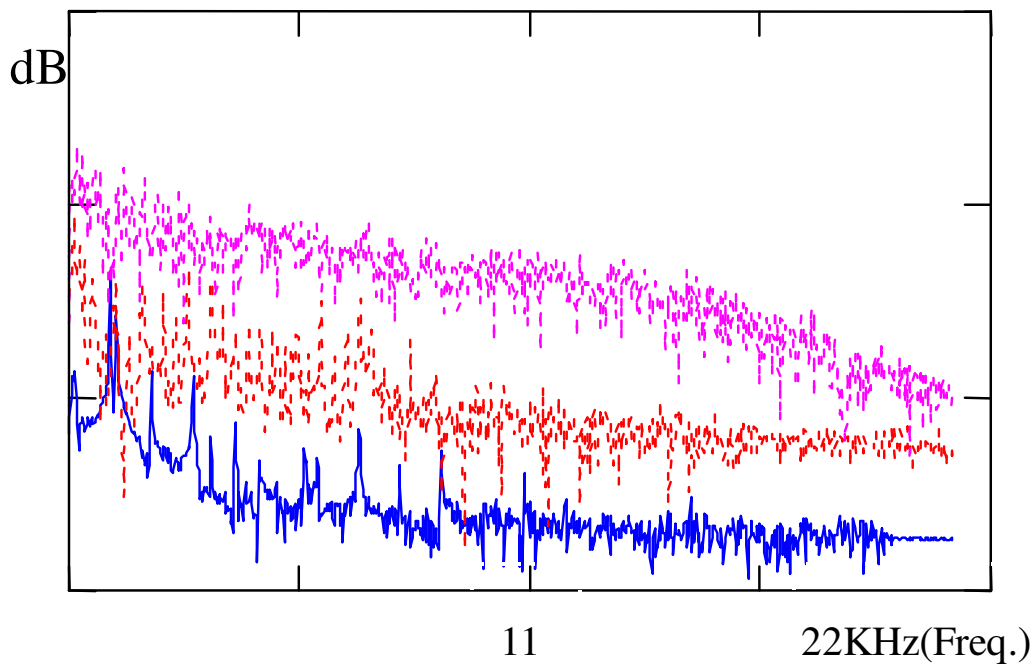


Fig. 4.11 Frequency responses of three typical audio sequences, where the lowest curve is encoded by D15, the middle curve by D25 and the highest curve by D45.

4.4 Concluding Remarks

In this chapter, fast algorithms for bit allocation is addressed. The fast algorithm for bit allocation is based on the fast noise estimator. The fast noise estimator, not using the ABS noise estimator to iteratively calculate the noise of each step sizes, provides a close-form equation for the relation of bits/step size and quantization noise. With the noise prediction formulae of uniform or non-uniform quantizer, several speedup algorithms are proposed in different papers. In this dissertation, the non-uniform quantizer of MPEG layer 3 is taken as an example. A single step bit allocation ensuring the criteria of maximal perceptual coding gain for this quantizer is proposed and it is also applicable to MPEG AAC non-uniform quantizer.

Chapter 5

KL Transform for Intensity/Coupling Coding

5.1 Introduction

When the two channels of stereo signals are coded, the stereo irrelevancy for the two channels expresses that the ability of the human auditory system to resolve the exact location of audio sources decreases with frequency. As stated in [17] and [29], the localization of the stereophonic image for the frequencies above 2 kHz is determined by the signal envelope instead of the signal fine structures. Following the stereo irrelevancy, the audio standards have developed the coupling or intensity schemes to efficiently remove the irrelevancy. Table 5.1 gives a summary on the coupling/intensity schemes in these audio coding standards.

In this chapter, KL (Karhunen-Loève) transform is introduced to design and analyze the intensity/coupling schemes. When integrating the KL transform into intensity coding/coupling schemes of MPEG and AC-3, two issues arise. The first issue lies on KL transform for intensity/coupling scheme might not perceptually optimal even if it is optimal in numerical sense. Second, due to the

constraints of different audio coders, KL coupling scheme might not tightly integrate with stereo matrix design of different coders. For example, in MPEG, during the summation process, when the signals in the left and right channels do not have the same signal sign, the signals from the two channels will be mutually canceled and it is hard to reconstruct the canceled information.

Algorithm	Stereo matrix	Coupling schemes mechanism	References
MPEG-1/2 layer 1/2	Intensity stereo	1. Scalefactors for L, R and one summation term are transmitted.	[24][23][29] [17][6]
MPEG-1/2 layer 3	Intensity/Mid-side stereo	1. Scalefactors for L, R and one summation term are transmitted.	[24][23][29] [17][6]
MPEG-2 AAC	Intensity/Mid-side stereo	1. Scalefactors for L, R and one summation term are transmitted.	[31][25]
Dolby AC-3	Coupling/Re-matrix	1. Scalefactors for L, R and one summation term are transmitted. 2. Phase flag is available. 3. Dithering scheme.	[43][27][11]

Table 5.1 A summary of stereo matrix mechanism among audio standards.

5.2 KL Transform for AC-3

When applying the Dolby AC-3 coder for the stereo music compression, the coupling scheme that combines the two channels stereo audio signals in high frequency into one channel is the key technology for the Dolby AC-3 to achieve the bit rates lower than 96x2 kbits/sec while preserving high stereo audio quality. This section proposes four coupling methods for the AC-3 encoder. These four methods vary with the complexity and performance. These four methods are

compared through both subjective and objective tests. These four coupling methods are also combined with the dithering scheme and examined through subjective and objective tests. The result shows that the dithering scheme can effectively ease the coupling artifacts and enhance the audio quality.

5.2.1 Addressed issues

The coupling scheme, which applies the low perceptual sensitivity of the stereo signals in high frequency to audio compression, is the key technology to achieve near transparent quality at the bit rates below 96x2 kbits/sec. The principle of the coupling scheme is derived from the stereo irrelevancy from the auditory systems. The stereo irrelevancy expresses that the ability of the human auditory system to resolve the exact location of audio sources decreases with frequency. As stated in [17] and [29], the localization of the stereophonic image for the frequencies above 2 kHz is determined by the signal envelope instead of the signal fine structures. Following the stereo irrelevancy, the AC-3 coder has developed the coupling scheme to achieve efficient compression. However, the standard draft [27] illustrates the decoupling process for the decoder and leaves unmentioned the coupling process for the encoder. This section proposes and compares four coupling methods for the coupling process of the AC-3 encoder.

Fig. 5.1 illustrates the block diagram for the coupling process in the Dolby AC-3. The audio sequences in stereo signal pairs are individually transformed into spectral lines and grouped into vectors referred to as the coupling bands. Fig. 5.1 shows the coupling process for one band corresponding to the same frequency range in a stereo signal pair. The bands from the left and the right

channels are coupled through the coupling block in Fig. 5.1. The coupling process produces four outputs: the coupling vector or band C_{band} , the two coordinate values (s_L , s_R) and a phase flag p . The coupling band C_{band} is quantized and packed into the AC-3 bit stream. In this manner, the bands from the left and the right channels have been reduced into one band to achieve data reduction. The decoder multiplies the left coordinate (or the right coordinate with negative if the phase flag is on) with the coupling band to reconstruct the left band (or right band). For the coupling process, the design criterion for the encoder is to provide appropriately the four coupling information such that the stereo signal bands can be reconstructed with good listening quality.

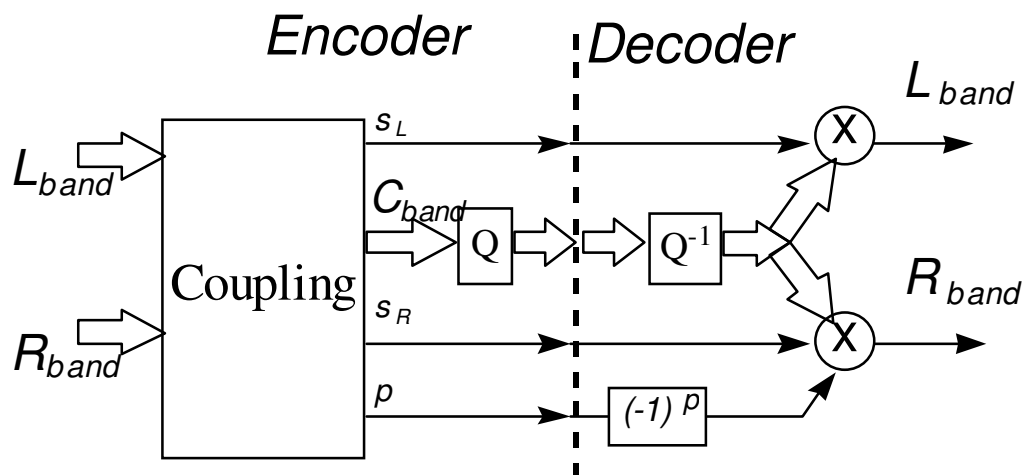


Fig. 5.1 Block diagram of the coupling process in a coupling band of the Dolby AC-3 codec.

As mentioned above, the sensitivity of the stereophonic image for the frequencies above 2 kHz is determined by the signal envelope instead of the signal fine structures. The coupling scheme in AC-3 keeps the audio contents

through the coupling band C_{band} , and preserves the envelope through the two coordinates (s_L, s_R) . Since the two bands have been reduced to one coupling band, it is impossible to reconstruct without loss the original two bands from the single band. Hence the design objective of the coupling is to keep envelope of the two bands through the coupling coordinates and minimizes the loss of the audio content through the coupling band. The coupling scheme is similar to the intensity coding in MPEG-1/2 audio coding. We have applied the Karhuner-Loeve transform to the intensity scheme to achieve the above objective in section 5.3 and also in [6]. The AC-3 has a higher potential to achieve a better performance than the intensity stereo in MPEG because of the two additional options: the phase flag and the dithering scheme. On these potential, this section proposes four coupling methods for the AC-3. Section 5.2.3 gives the subjective and objective comparison for these four methods.

5.2.2 Four proposed coupling methods

We developed four methods for the coupling scheme. These four methods differ in the complexity and the associated fidelity concepts as illustrated in Fig. 5.2-Fig. 5.5. Considering the SUM algorithm in Fig. 5.2, the coupling vector C_{band} is evaluated by summing the band signals R_{band} and L_{band} in the left and the right channels. For energy preservation, the two coordinate values (s_L, s_R) are calculated from the square root of the energy ratio for the R_{band} and C_{band} , and the ratio L_{band} and C_{band} . The phase flag P is fixed to be 0 in this method. The detailed algorithm of the SUM algorithm is illustrated as follows:

Encoding process for the SUM algorithm

1. The phase flag evaluation process

$$p_{\text{band}}=0.$$

2. The summation process

$$C_{\text{band}}=L_{\text{band}}+R_{\text{band}}.$$

3. The coordinates evaluation process

$$s_L=\text{Energy}(L_{\text{band}})^{0.5}/\text{Energy}(C_{\text{band}})^{0.5}$$

$$s_R=\text{Energy}(R_{\text{band}})^{0.5}/\text{Energy}(C_{\text{band}})^{0.5}$$

$$\text{where } \text{Energy}(S_{\text{band}}) = \sum_{\text{bin in band}} S_{\text{bin}}^2 .$$

(5.1)

For the NORM_SUM algorithm in Fig. 5.3, the coupling vector C_{band} is calculated by summing the energy-normalized signals $R_{\text{band}}/\text{Energy}(R_{\text{band}})^{0.5}$, $L_{\text{band}}/\text{Energy}(L_{\text{band}})^{0.5}$. The two coordinate values (s_L , s_R) and the phase flag p are decided in the same way as the SUM algorithm. The NORM_SUM algorithm indicates that the larger value of L or R will not dominate during the summation process as the SUM algorithm. The detailed algorithm of the NORM_SUM algorithm is illustrated as follows:

Encoding process for the NORM_SUM algorithm

1. The phase flag evaluation process

$$p_{\text{band}}=0.$$

2. The summation process

$$C_{\text{band}}=$$

$$L_{\text{band}}/\text{Energy}(L_{\text{band}})^{0.5}+R_{\text{band}}/\text{Energy}(R_{\text{band}})^{0.5}$$

3. The coordinates evaluation process

$$s_L = \text{Energy}(L_{\text{band}})^{0.5} / \text{Energy}(C_{\text{band}})^{0.5}$$

$$s_R = \text{Energy}(R_{\text{band}})^{0.5} / \text{Energy}(C_{\text{band}})^{0.5}$$

where $\text{Energy}(S_{\text{band}})$ is defined in (5.1).

The KLT_MSE algorithm in Fig. 5.4 directly applies the Karhuner-Loeve (KL) transform to the coupling process in AC-3. The KL transform and the inverse KLT for $N=2$ can be viewed as the rotation matrix

$$\begin{bmatrix} I \\ E \end{bmatrix} = \begin{bmatrix} \cos \alpha & \sin \alpha \\ -\sin \alpha & \cos \alpha \end{bmatrix} \begin{bmatrix} L \\ R \end{bmatrix}; \begin{bmatrix} L \\ R \end{bmatrix} = \begin{bmatrix} \cos \alpha & -\sin \alpha \\ \sin \alpha & \cos \alpha \end{bmatrix} \begin{bmatrix} I \\ E \end{bmatrix} \quad (5.2)$$

where L and R are signals of the left and right channels, and I and E are transformed intensity and error channel. The rotation angle for the KL transform can be evaluated from

$$\tan(2\alpha) = \frac{2c_{lr}}{c_{ll} - c_{rr}}; -\frac{\pi}{2} \leq \alpha < \frac{\pi}{2} \quad (5.3)$$

where C_{ll} and C_{rr} are the autocorrelation coefficients of the left and the right channels. C_{lr} is the cross-correlation coefficient of the left and the right channels. In least mean square error sense between decoded signals and input signals, the error channel is ignored and the KLT matrix becomes

$$\begin{bmatrix} I \\ 0 \end{bmatrix} = \begin{bmatrix} \cos \alpha & \sin \alpha \\ -\sin \alpha & \cos \alpha \end{bmatrix} \begin{bmatrix} L \\ R \end{bmatrix}; \begin{bmatrix} L \\ R \end{bmatrix} = \begin{bmatrix} \cos \alpha & -\sin \alpha \\ \sin \alpha & \cos \alpha \end{bmatrix} \begin{bmatrix} I \\ 0 \end{bmatrix}. \quad (5.4)$$

From (5.4), the coordinates of left and right channels for the KLT_MSE algorithm are $\cos \alpha$, $\sin \alpha$ and the coupling vector can be obtained by

$L_{band} \cos \alpha + R_{band} \sin \alpha$. In order to embed into the AC-3, the coordinates in AC-3 allow only positive values. Thus, by the phase modifier flag p , the coordinates of left and right channels and the coupling vector are changed to $\cos \alpha$, $\sin \alpha(-1)^p$ and $L_{band} \cos \alpha + R_{band} \sin \alpha(-1)^p$. From above, the KLT_MSE algorithm ensures the least mean square error of the original coupling vector and decoded coupling vector even the signals of the left and the right channels are negatively correlated. The detailed KLT_MSE algorithm is demonstrated as follows:

Encoding process for the KLT_MSE algorithm

1. The rotation angle evaluation process

The rotation the angle α defined in (2).

2. The phase flag evaluation process

$$p = \begin{cases} 1 & \text{if } \sin(\alpha) < 0 \\ 0 & \text{otherwise} \end{cases}$$

3. The summation process

$$C_{band} = L_{band} \cos \alpha + R_{band} \sin \alpha(-1)^p.$$

4. The coordinates evaluation process

$$s_L = \cos \alpha$$

$$s_R = \sin \alpha(-1)^p.$$

For the KLT_ENG algorithm in Fig. 5.5, a compromise between the SUM and KLT_MSE algorithm is considered. The two coordinate values (s_L , s_R) are decided from the square root of the energy ratio for the R_{band} and C_{band} , and the energy ratio for L_{band} and C_{band} . The detailed algorithm of the KLT_ENG

algorithm is shown as follows:

Encoding process for the KLT_ENG algorithm

1. The rotation angle evaluation process

The rotation angle α is defined in (2).

2. The phase flag evaluation process

$$p = \begin{cases} 1 & \text{if } \sin(\alpha) < 0 \\ 0 & \text{otherwise} \end{cases}$$

3. The summation process

$$C_{band} = L_{band} \cos \alpha + R_{band} \sin \alpha (-1)^p.$$

4. The coordinates evaluation process

$$s_L = \text{Energy}(L_{band})^{0.5} / \text{Energy}(C_{band})^{0.5}$$

$$s_R = \text{Energy}(R_{band})^{0.5} / \text{Energy}(C_{band})^{0.5}$$

where $\text{Energy}(S_{band})$ is defined in (5.1).

Among them, the methods in Fig. 5.4 and Fig. 5.5 are developed based on the KL transform. The KLT can minimize the square-errors during the coupling of two bands into one band. However, the KLT also leads to higher complexity than the other two methods.

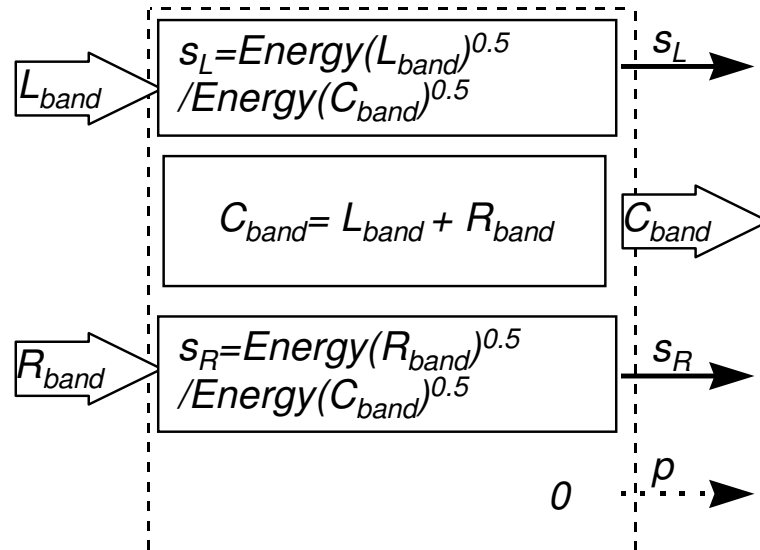


Fig. 5.2 The SUM algorithm for the coupling process.

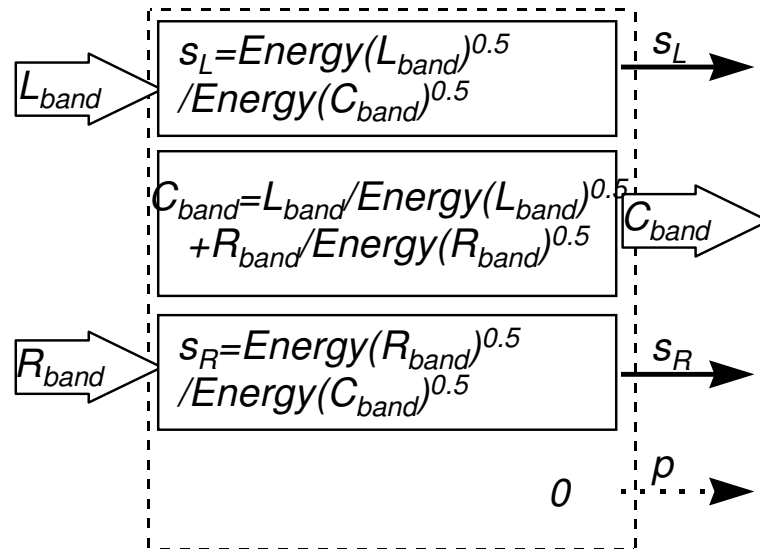


Fig. 5.3 The NORM_SUM algorithm for the coupling process.

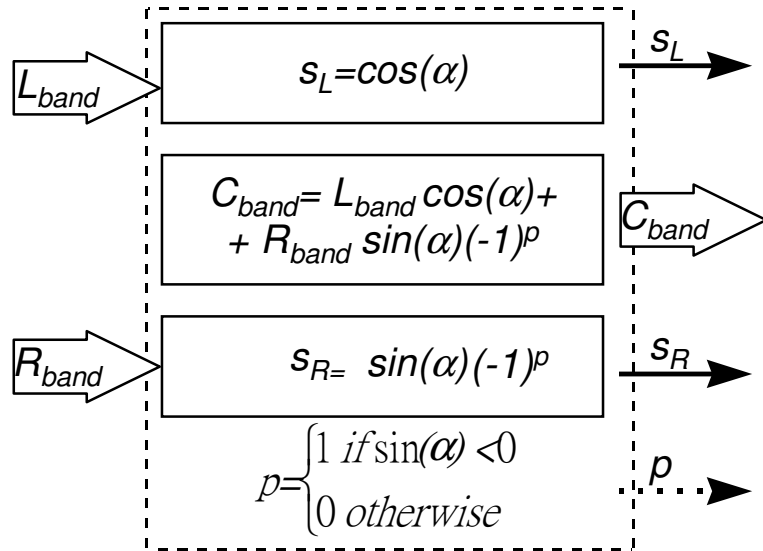


Fig. 5.4 The KLT_MSE algorithm for the coupling process.

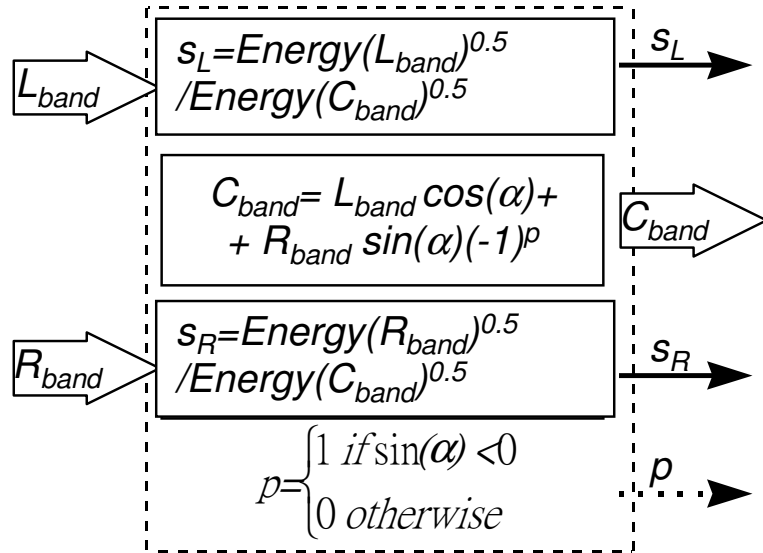


Fig. 5.5 The KLT_ENG algorithm for the coupling process.

5.2.3 Experiments on the coupling methods

The performances of the four coupling methods are compared through objective tests and subjective tests. A total of nine 20 sec stereo audio songs including vocal, symphony, piano and so on are taken as the materials for testing. The detailed descriptions of the test materials are listed in Table 5.2. The objective measure is verified by the segmental noise-to-masking ratio (NMR)

value defined by averaging the NMR values in each coupling band in each frame as

$$NMR_{seg} = \frac{1}{F} \sum_f \left(\frac{1}{B} \sum_b SMR_{f,b} - SNR_{f,b} \right)$$

where the SMR stands for the signal-to-masking ratio in dB, the SNR for the signal-to-noise ratio in dB, F for the total audio frames, f for the frame number, B for the total coupling bands, and b for the coupling band number. Negative values of the NMR_{seg} indicate that the noise of the coded signal is inaudible, and larger negative values of NMR_{seg} indicate the noise may be more inaudible. The coupling scheme is performed in the range of 3.14 KHz to 12.45 KHz. The coupling methods are performed under high bit rate and the exponents are transmitted with D15 mode for six times in a frame. Table 5.3 illustrates the testing results. The results indicate that the KLT_MSE and KLT_ENG algorithm can have better NMR_{seg} values than the SUM and NORM_SUM algorithm. The SUM and NORM_SUM algorithms cause coupling artifacts and poor NMR_{seg} values due to signal cancellation when L_{band} and R_{band} are negatively correlated. We further consider the encoding for the bit rate at 128 kbits/s and the exponents strategy D15 is transmitted once per frame. The test results are summarized in Table 5.4 that indicates the order of the performance being the KLT_MSE, KLT_ENG, SUM and NORM_SUM algorithm.

In the subjective test under the critical bit rate at 128 kbits/s, the same test materials in Table 5.2 are evaluated. The results of the listening test show the order of the quality performance of the four coupling methods is the KLT_ENG, SUM, NORM_SUM, and KLT_MSE algorithm. Although the excellent performance of the objective tests, the KLT_MSE algorithm gives poor

subjective performance due to some ringing noise. The noise may be due to the discontinuous coordinates across different bands in the KLT_MSE algorithm. To sum up, the KLT_ENG algorithm gives high performances on both objective and subjective tests because it takes the advantages from the KLT_MSE algorithm on the signal preservation and the SUM algorithm on the energy preservation.

5.2.4 Dithering on the coupling bands

In AC-3, dithering scheme is to add white noise to the coded bands in the decoding process. For low bit rate audio coding, quantization leads to the noises that are correlated with signals. Such a correlation is very sensitive for the human hearing systems. Especially, the coupling scheme can also lead to the artifacts as mentioned in last section. Dithering can reduce the artifacts from either the quantization or the coupling process. The four coupling methods presented in last section are examined through subjective tests when the dithering in AC-3 is applied. In our subjective listening test for the SUM and NORM_SUM algorithm, the dithering can significantly reduce the coupling noise. As a result, the quality from the KLT_ENG, SUM, and NORM_SUM algorithm become indistinguishable when the dithering is applied.

5.2.5 Remarks

In this section, four coupling methods for the AC-3 encoder have been introduced. These four methods vary with the complexity and performance. Both subjective and objective tests have been conducted and demonstrated the

performance of the KLT_ENG algorithm is better than other algorithms. We have also demonstrated that the dithering scheme gives great improvement on the quality of the coupling methods. With the dithering scheme, the performance of the four coupling methods is similar and the algorithm with low complexity will be more essential.

Test song	Description
Symphony	The Choral symphony (Choral part)
Piano	Pure and clear piano
Violin	Violin playing from low to high frequency
Flute	Clear flute sound
Woman	Pure woman vocal song
Pipe	Pure pipe sound
Man	Man vocal song; country music song
Violoncello	Violoncello sound in low frequency
Drum	Pure pipe sound & sudden and loud drum

Table 5.2 Testing audio segments and their descriptions.

Algorithms	SUM		NORM_SUM		KLT_MSE		KLT_ENG	
	left	right	left	right	left	right	left	right
D15 6 times								
Symphony	-2.19	-2.75	-2.61	-0.95	-5.07	-7.17	-3.82	-6.18
Piano	-6.99	1.21	-5.72	1.29	-10.1	-6.01	-9.22	-4.72
Violin	5.90	7.81	5.74	10.2	1.42	-1.67	2.72	-0.66
Flute	-4.23	2.89	0.74	2.31	-10.1	-1.36	-9.49	0.02
Woman	1.17	8.35	1.26	9.23	0.45	1.17	1.36	1.96
Pipe	-12.4	-11.2	-12.1	-10.8	-12.9	-15.5	-12.4	-15.0
man	-2.91	16.5	-2.75	16.5	-3.19	-3.94	-2.97	-3.72
Violoncello	-8.61	-9.99	-9.56	-8.20	-8.23	-12.7	-7.35	-12.2
Drum	5.88	5.27	6.87	6.42	4.33	3.56	5.59	4.80

Table 5.3 NMR_{seg} values for the four proposed coupling methods under high bit

rate with D15 mode 6 times per frame.

Algorithms	SUM		NORM_SUM		KLT_MSE		KLT_ENG	
	left	right	left	right	left	right	left	right
D15 1 times								
Symphony	40.4	39.8	40.0	40.3	39.8	38.8	41.0	39.5
Piano	34.8	37.8	34.8	37.9	33.8	33.5	34.3	34.2
Violin	37.5	39.8	37.1	40.3	35.2	34.7	36.2	34.1
Flute	33.3	33.7	34.7	33.4	32.7	31.5	32.8	32.0
Woman	36.5	37.6	36.0	37.8	36.4	35.3	37.2	35.8
Pipe	33.3	34.4	33.3	34.4	33.0	33.1	33.4	33.5
man	35.9	44.0	35.4	44.0	36.0	36.0	36.2	36.6
Violoncello	36.2	36.3	35.3	36.3	36.6	36.1	37.1	36.2
Drum	36.9	37.3	37.4	37.3	36.3	36.5	37.0	37.6

Table 5.4 NMR_{seg} values for the four proposed coupling methods under the bit rate of 128 kbits/sec with D15 mode once per frame.

5.3 KL Transform for MPEG Intensity Coding [6]

The coupling scheme in MPEG is called intensity stereo coding. Several addressed problems of the original MPEG-1 intensity stereo coding and modification can be found in [17], [29]. In [39], the idea of KL (Karhunen-Loève) transform has been considered to analyze the data redundancy between the stereo channels. Also, the authors have suggested the applying of the transform to intensity coding. As mentioned in Section 5.2, this section propose two methods to implement the KL transform in the MPEG-1 layers 1 and 2 [6].

Consider the block diagram in Fig. 5.6, two problems arising from the

process. The first problem is on the consistency between the scalefactors in the encoder and the decoder. As shown in Fig. 5.6, the signals from the left and right channel are summed together and jointly scaled by a scalefactor K_J , while the decoders utilize the scalefactors K_R and K_L to rescale the decoded samples. There is no direct relation between the K_J and the pair (K_R, K_L) . Hence, the decoder and the encoder do not have consistent scalefactors. The second problem concerns with the signal cancellation in the summation process. During the summation process, when the signals in the left and the right channels do not have the same signal sign, the signals from the two channels will be mutually canceled and it is hard to reconstruct the canceled information. The researches in [17], [29] try to ease these problems by modifying the transmitted scalefactors. Such an approach can ease the problem of the consistency of scalefactors, but cannot provide help on the signal cancellation problem. This section presents an approach to modify both the scalefactor calculation and the summation manner to ease the above two problems.

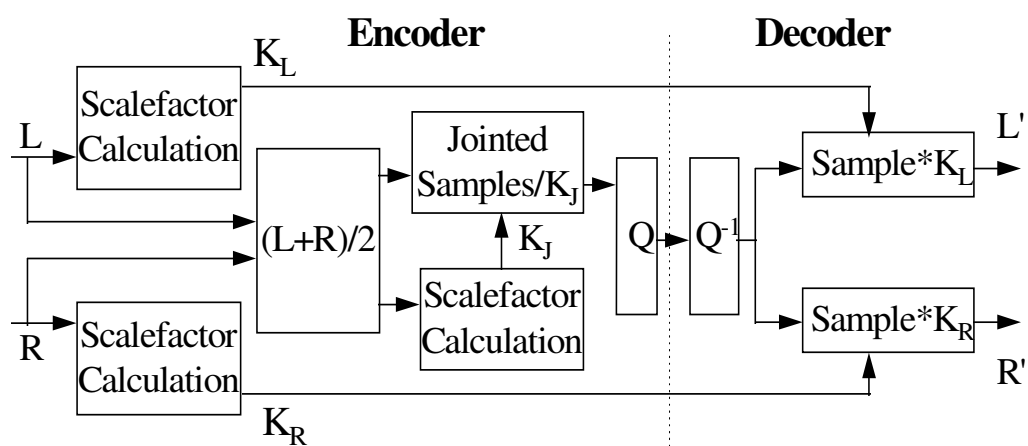


Fig. 5.6 Intensity stereo coding of MPEG-1 (SUM) in a high frequency band (adopted from [6]).

In the first method, when the angle α is positive, we perform our intensity stereo coding algorithm as shown in Fig. 5.7; when the angle is negative, we perform the original MPEG-1 intensity stereo coding. In this way, the method can be totally compatible to the MPEG-1 standard in the sense that the same decoder as MPEG-1 can be used to decode the bitstreams encoded by the method. However, the presented method has sacrificed parts of the potential of the KL transform. This method is denoted as KL_MSE compatible coding method.

In the second method, similar to phase flag in AC-3, we transmit the joint scalefactor K_J and the angle α to approximate the KL transform indicated in Fig. 5.7. The joint scalefactor is quantized as six bits based on the look-up table designed for the scalefactors in MPEG-1. The rotation angle α is also quantized as six bits. The table shows the 32 positive quantized angles that are used to quantize the legal angles ranging from 0 to $\pi/2$. The negative angles have the same values but negative signs. This method can approximate the KL transform under the same bit rate as MPEG-1, but a slight modification on the decoder is required to decode the bitstreams encoded by the method. This method is denoted as KL_MSE non-compatible coding method.

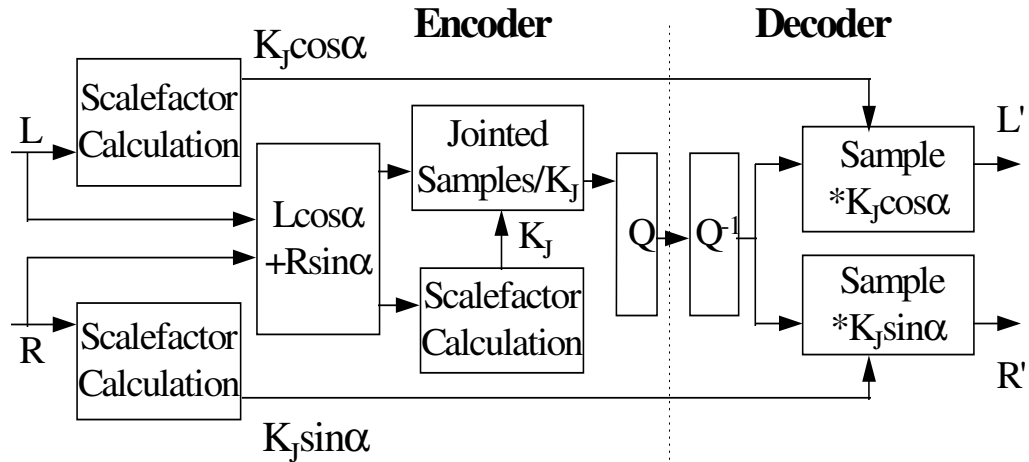


Fig. 5.7 KL_MSE intensity coding in a high frequency band (adopted from [6]).

Methods Test	Original MPEG (SUM)	KLT_MSE Compatible	KLT_MSE Non-compatible
1. Carmen	-0.5985	-0.2276	0.6296
	-1.3170	-1.2783	-0.7510
2. Songs	-7.3448	-6.5771	-5.6519
	-7.3165	-6.4914	-5.6685
3. Huqin	-1.2521	-0.8507	-0.7297
	-1.2330	-0.5945	-0.7566
4. Drum	-5.1192	-4.5126	-3.9989
	-5.2201	-4.6360	-4.8985
5. Violin	-3.0766	-2.9204	-1.5388
	-2.1584	-1.7142	-0.3412
6. Orchestra	-6.3791	-5.7489	-4.1817
	-6.7642	-6.3137	-5.0613
7. Guitar	-4.4968	-4.0042	-3.8239
	-3.6040	-2.8042	-2.7585

Table 5.5 MNR (dB) values in layer 2. In each box, the upper value is for the left channel, the lower value is for the right channel (adopted from [6]).

From [6], the MNR results of implementation in MPEG-1 layer 2 are shown in Table 5.5, respectively. All the test results show that the two KL_MSE

intensity coding methods can have a lower MNRs than the original MPEG intensity coding method. Among the two KL intensity coding methods, the KL_MSE non-compatible coding method can have a better performance than the compatible one.

5.4 Concluding Remarks

KL transform is introduced to obtain the optimal solution for the coupling process in numerical sense. When integrating the KL transform into coupling schemes of MPEG and AC-3, two issues arise. The first issue lies on KL coupling scheme might not perceptually optimal even if it is optimal in numerical sense. Second, due to the constraints of different audio coders, KL coupling scheme might not tightly integrate with stereo matrix design of different coders. For example, in MPEG, during the summation process, when the signals in the left and right channels do not have the same signal sign, the signals from the two channels will be mutually canceled and it is hard to reconstruct the canceled information.

Chapter 6

Conclusions and Future Works

6.1 Concluding Remarks

This dissertation has studied the design of audio standards: MPEG-1/2 and AC-3. We have proposed the fast algorithms for the filterbank, the psychoacoustic model, and the bit allocation. Also, this dissertation has designed the new intensity/coupling schemes.

On the filterbank, a unified fast algorithm of filterbank for variant form and variant size has been presented. On the psychoacoustic model, a hybrid filterbank has been proposed to replace to original frequency analyzer, Fourier transform. On the bit allocation, we first present the efficient bit allocation method for MPEG layer 3 with non-uniform quantization and variable length coding and then present criteria for the bit allocation for Dolby AC-3 and propose efficient bit allocation algorithm according to the criteria. On the intensity/coupling, this dissertation applied KL transform to design the parameters for MPEG and AC-3 to have a better encoding quality.

6.2 Future Works

This dissertation studies the design issues and experiments based on the MPEG-1 and AC-3. However, the design concepts are never restricted to the two standards. The applying of the design concepts under the constraints of the protocols by new standards such as MPEG AAC and MPEG4 is the direct extension of the dissertation. In Chapter 2: unified algorithm for fast filterbank computing, this dissertation proposes fast algorithms that unify the variant form and variant size of cosine modulated filter banks. The size of the cosine modulated filter banks is limited to a number of power of 2 due to the recursion of the fast algorithm. In fact, in MPEG-1 and MPEG-4, there are exceptions for this constraint. More researches for this issue can be studied. In Chapter 4: fast bit allocation method, this dissertation proposes an efficient bit allocation algorithm for mono channel of MPEG layer 3. More researches can be studied on the efficient bit allocation algorithms of variable bit rate for each frame and efficient algorithms for stereo channels. MPEG allows variable bit rate for each frame. This gives more flexibility to ensure perceptual quality according to the information from psychoacoustic model. When a frame deserves more bits according to psychoacoustic model, more bit rate will be given iteratively or predictive until ensuring quality. When a frame deserves fewer bits, fewer bits will be given iteratively or predictive. For stereo channels, MPEG layer 3 allows bit numbers can be shared in variant ratio for left/right or middle/side channel by the mechanism of bit reservoir and joint stereo coding. For the more and more complexity of the variable bit rate and bit share ratio for stereo channel, the

proposed algorithm mentioned in section 4.2 provides more potential for efficient bit allocation.

New mechanisms such as gain control, temporal noise shaping, prediction, and transform domain interleaved vector quantization give more potential for quality improvement, but these modules also lead to new design issues on combining with the design modules discussed in this dissertation. The combined consideration with these modules is another issue deserving further study.

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Curriculum Vita

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