channel estimation is used. The f_dT of fading channels is 0.000723 in this simulation. All receivers in this simulation adopt pilot symbols for channel estimation and a Rake structure for the multipath fading channel. The SRM receiver does not need an additional training sequence by setting the tap weight of a desired user to 1 in the initial state.

Fig. 3 shows a comparison of BER performance among three CDMA receivers, in which the MMSE receiver models the modified MMSE receiver in [1] with a fractionally spaced filter with a tap spacing four times smaller than the chip duration. Setting E_b/N_0 to 15dB, the MMSE receiver achieves 5.8dB gain in BER over that of a conventional receiver (MF), while the SRM receiver has gain of 17.8dB. Another simulation result shows that the SRM receiver also outperforms other receivers in a lower signal-to-interference ratio ($K = 25$).

From the results, it is clear that the SRM receiver significantly outperforms all other MMSE receivers. A potential issue is that the SRM receiver requires a function-based base station structure, which needs a high-speed router to link between each functional module.

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Transformation from 512-point transform coefficients to 256-point transform coefficients for Dolby AC-3 decoder

Szu-Wei Lee and Chi-Min Liu

The Dolby AC-3 standard defines 512-point and 256-point transforms to provide high audio quality when signals rapidly change in time. This authors provide theoretical derivations and proofs for the transformation from 512-point transform coefficients to 256-point transform coefficients.

Introduction: The Dolby AC-3 is currently the HDTV audio standard and is widely used in DVD films [1]. While the AC-3 draft has been frozen, the standard defines one 512-point (referred to as the long transform) and two 256-point transforms (referred to as the first and second short transforms) to improve coding quality for signals which rapidly change in time. However, since the transforms are of differing lengths, this increases the difficulty of implementing channel operations in the frequency domain. Channel operations, such as downmixing and channel reduction, are linear operations that mix numbers of channels to another channel. The objectives of performing channel operations in the frequency domain are frequency selection for flexible computational complexity, and equalisation for sound enhancement. Since the transforms adopted in AC-3 can be 512-point or 256-point, channel operations cannot mix the channels encoded by the short transforms with those encoded by the long transform. This Letter develops a transform matrix, denoted by V, to transform the coefficients of the long transform to those of the short transforms. As a result, the channels encoded by the long transform are pre-processed by matrix V so that channel operations can mix them with those encoded by the short transforms.

Overview of AC-3 transform process: The long transform is the time domain aliasing cancellation (TDAC) filter bank proposed in [2], but the short transforms are slightly different. Fig. 1 outlines the AC-3 transform process. The transform process comprises four modules: overlapping-windowing (OW), forward modified discrete cosine transform (MDCT), backward MDCT and windowing overlapping-and-adding (WOLA).

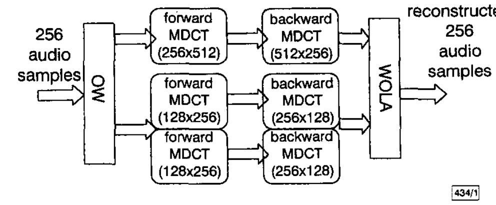

Fig. 1 *Transform process of AC-3*

For forward transforms, each audio block containing 512 overlapping samples, which overlaps the second half of the previous audio block, is multiplied by analysis window $h(n)$ to produce 512 windowing samples. The 512 windowing samples can be transformed by either one 256×512 forward MDCT (hereafter referred to as L) for the long transform, or two 128×256 forward MDCTs (referred to as S_1 and S_2) for the short transforms. As a result, the forward transforms produce 256 frequency coefficients or two sets of 128 frequency coefficients.

For backward transforms, the 256 frequency coefficients are transformed by a 512×256 backward MDCT (referred to as L^+), or the two sets of 128 frequency coefficients are transformed by two 256×128 backward MDCTs (referred to as S_1^+ and S_2^+). The backward MDCTs result in 512 windowing samples. The 512 windowing samples are multiplied by synthesis window $f(n)$ to produce 512 overlapping samples. The first half of the 512 overlapping samples are then overlapped and added to the second half of the previous 512 overlapping samples. Finally, the 256 time samples are reconstructed.

Notation: L^+ is a 512×256 matrix and each entry nk of L^+ is defined by

$$\langle L^+\rangle_{nk} = \cos\Big(\frac{\pi}{1024}(2n+1)(2k+1) + \frac{\pi}{4}(2k+1)\Big)$$

We further decompose L^+ into two square matrices, i.e.

$$L^+ = \begin{bmatrix} L_1^+ \\ L_2^+ \end{bmatrix}$$

where L_1^+ and L_2^+ are 256×256 matrices. S_1^+ and S_2^+ are 256×128 matrices. Each entry nk of S_1^+ is defined by

$$\langle S_1^+\rangle_{nk} = \cos\Big(\frac{\pi}{512}(2n+1)(2k+1)\Big)$$

and each entry nk of S_2^+ is defined by

$$\langle S_2^+\rangle_{nk} = \cos\Big(\frac{\pi}{512}(2n+1)(2k+1) + \frac{\pi}{2}(2k+1)\Big)$$

S_1 and S_2 are transpose matrices of S_1+ and S_2^+, respectively, with dimensions 128×256. H and F are 512×512 diagonal matrices with entries $\langle H\rangle_{nn} = h(n)$ and $\langle F\rangle_{nn} = f(n)$. Also, partitioning matrices H and F into two parts yields

$$H = \begin{bmatrix} H_1 & 0 \\ 0 & H_2 \end{bmatrix} \quad \text{and} \quad F = \begin{bmatrix} F_1 & 0 \\ 0 & F_2 \end{bmatrix}$$

Y_f is a 256×1 vector denoting the frequency coefficients of the long transform. $Y1_f$ and $Y2_f$ are 128×1 vectors, where $Y1f$ denotes the frequency coefficients of the first short transform, $Y2_f$ denotes the coefficients of the second short transform, and subscript f is the block number. X_f and $\tilde{X}_f$ represent the original and reconstructed time domain signals, respectively.

Derivation: Since the transforms adopted in the AC-3 satisfy the perfect reconstruction constraint, the original time domain signal X_f can be obtained from the synthesis signal $\tilde{X}_f$, i.e.

$$X_f = \tilde{X}_f = \begin{bmatrix} \tilde{X}1_f \\ \tilde{X}2_f \end{bmatrix}$$

and

$$\tilde{X}1_f = \begin{cases} F_1 \cdot L_1^+ \cdot Y_f + F_2 \cdot S_2^+ \cdot Y2_{(f-1)} \\ \quad \text{if previous block adopts short transforms} \\ F_1 \cdot L_1^+ \cdot Y_f + F_2 \cdot L_2^+ \cdot Y_{(f-1)} \\ \quad \text{if previous block adopts long transform} \end{cases} \tag{1}$$

$$\tilde{X}2_f = \begin{cases} F_2 \cdot L_2^+ \cdot Y_f + F_1 \cdot S_1^+ \cdot Y1_{(f+1)} \\ \quad \text{if next block adopts short transforms} \\ F_2 \cdot L_2^+ \cdot Y_f + F_1 \cdot L_1^+ \cdot Y_{(f+1)} \\ \quad \text{if next block adopts long transform} \end{cases} \tag{2}$$

Theorem (i): The following matrices, Z_1, Z_2, Z_3 and Z_4, are zero matrices, where Z_1, Z_2, Z_3 and Z_4 are defined by $Z_1 = S_1 \cdot H_1 \cdot F_2 \cdot S_2^+$, $Z_2 = S_1 \cdot H_1 \cdot F_2 \cdot L_2^+$, $Z_3 = S_2 \cdot H_2 \cdot F_1 \cdot S_1^+$, and $Z_4 = S_2 \cdot H_2 \cdot F_1 \cdot L_1^+$.

Proof of theorem (i): Let $z1(k1,k2)$ be the entry of the matrix Z_1, i.e.

$$z1(k1,k2) = \sum_{n=0}^{255} \cos\Big(\frac{\pi}{1024}(4k_1+2)(2n+1)\Big) h(n) f(256+n) \times \cos\Big(\frac{\pi}{1024}(4k_2+2)(2n+1+256)\Big) \tag{3}$$

Since the analysis window $h(n)$ and synthesis window $f(n)$ are equal and have the symmetry property, eqn. 3 can be rewritten as

$$z1(k1,k2) = \sum_{n=0}^{255} \Big[h(n)h(256-1-n)\cos\Big(\frac{\pi}{1024}(4k_1+2)(2n+1)\Big) \times \cos\Big(\frac{\pi}{1024}(4k_2+2)(2n+1+256)\Big)\Big] \tag{4}$$

We now partition the summation term in eqn. 4 into two groups. Each has 128 terms.

$$\begin{aligned} z1(k1,k2) = &\sum_{n=0}^{127} \Big\{h(n)h(256-1-n)\cos\Big(\frac{\pi}{1024}(4k_1+2)(2n+1)\Big) \times \cos\Big(\frac{\pi}{1024}(4k_2+2)(2n+1+256)\Big)\Big\} \\ &+ \sum_{n=128}^{255} \Big\{h(n)h(256-1-n)\cos\Big(\frac{\pi}{1024}(4k_1+2)(2n+1)\Big) \times \cos\Big(\frac{\pi}{1024}(4k_2+2)(2n+1+256)\Big)\Big\} \end{aligned} \tag{5}$$

Replacing the index n in the second group by $(255-n)$, we obtain

$$\begin{aligned} z1(k1,k2) = &\sum_{n=0}^{127} \Big\{h(n)h(256-1-n)\cos\Big(\frac{\pi}{1024}(4k_1+2)(2n+1)\Big) \times \cos\Big(\frac{\pi}{1024}(4k_2+2)(2n+1+256)\Big)\Big\} \\ &+ \sum_{n=0}^{127} \Big\{h(n)h(256-1-n)\cos\Big(\frac{\pi}{1024}(4k_1+2)(512-1-2n)\Big) \times \cos\Big(\frac{\pi}{1024}(4k_2+2)(768-1-2n)\Big)\Big\} \end{aligned} \tag{6}$$

Since the cosine terms satisfy the following properties

$$\cos\Big(\frac{\pi}{1024}(4k_1+2)(2n+1)\Big) = -\cos\Big(\frac{\pi}{1024}(4k_1+2)(512-1-2n)\Big)$$

$$\cos\Big(\frac{\pi}{1024}(4k_1+2)(2n+1+256)\Big) = \cos\Big(\frac{\pi}{1024}(4k_2+2)(768-1-2n)\Big) \tag{7}$$

Z_1 is a zero matrix. For Z_2, Z_3 and Z_4 the proofs can be carried out similarly.

Theorem (ii): The short transform coefficients can be obtained if the long transform coefficients are known. Specifically, the transform matrix from the coefficients of the long transform to those of the short transforms is described as follows

$$\begin{bmatrix} Y1_f \\ Y2_f \end{bmatrix} = V \cdot Y_f \quad \text{where } V = \begin{bmatrix} V_1 \\ V_2 \end{bmatrix} = \begin{bmatrix} S_1 \cdot H_1 \cdot F_1 \cdot L_1^+ \\ S_2 \cdot H_2 \cdot F_2 \cdot L_2^+ \end{bmatrix} \tag{8}$$

Proof of theorem (ii): The short transform coefficients can be expressed as

$$Y_f = [S_1 \quad S_2] \begin{bmatrix} H_1 & 0 \\ 0 & H_2 \end{bmatrix} X_f \tag{9}$$

Substituting eqns. 1 and 2 into eqn. 9 and applying *theorem (i)* yields

$$\begin{bmatrix} Y1_f \\ Y2_f \end{bmatrix} = \begin{bmatrix} S_1 \cdot H_1 \cdot F_1 \cdot L_1^+ \\ S_2 \cdot H_2 \cdot F_2 \cdot L_2^+ \end{bmatrix} Y_f \tag{10}$$

Conclusion: In this Letter, we have established the transform matrix that converts the coefficients of long transforms into those of the short transforms. This matrix makes possible the implementation of channel operations in the frequency domain. This work has given the theorems and proofs that support derivation of this matrix.

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Ultra-low nonlinearity low-loss pure silica core fibre for long-haul WDM transmission

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A new design of pure silica core fibre with an effective area > 110μm² and attenuation of 0.171dB/km at 1.55μm has been successfully developed. It exhibits ultra-low nonlinearity and excellent bending loss performance.

Introduction: Recent investigation has shown that a combination of dispersion-unshifted singlemode fibres and dispersion-compensating fibres is one of the most promising transmission line configurations for long-haul WDM (wavelength division multiplexing) transmission systems [1 – 3]. The former fibre, which has relatively large chromatic dispersion at 1.55μm, is effective for reducing fibre nonlinearity effects, such as four-wave mixing, while the latter compensates for the accumulated dispersion. Since dispersion-compensating fibres tend to have slightly higher attenuation than standard fibres, use of low-loss pure silica core fibres (PSCFs) with an attenuation of 0.17–0.18dB/km at 1.55μm instead of