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AN OVERLAPPED BLOCK MOTION COMPENSATION FOR HIGH QUALITY MOTION PICTURE CODING

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ABSTRACT

This paper describes an overlapped block motion compensation scheme, which is especially suitable for hybrid coding schemes combined with overlapped transforms, such as LOT, MDCT and Wavelet Transform.

In the proposed method, motion compensation is operated using enlarged and overlapped blocks. When motion estimation is executed, a window function is operated to the prediction error signals. The power of the obtained signal is calculated and used for motion vector evaluation. When motion compensated prediction is executed, the same window function is operated to the prediction signals. A whole prediction signal is generated by summing up all the block signals obtained.

With this new overlapped block motion compensation scheme, prediction signals without blocking edges are obtained. Mean Square Error (MSE) of the prediction error signal is reduced about 19%, and the FFT analysis results show that the coefficient power in the frequency domain is concentrated to the lower frequency part. This promises higher coding efficiency especially for the overlapped transforms, but it is also effective for the conventional block based transforms.

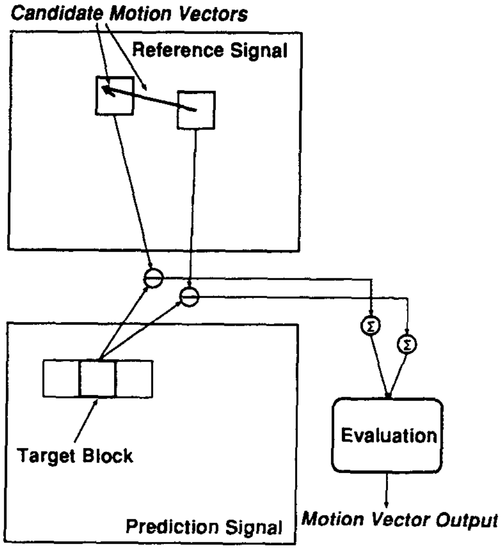

Figure 1. Conventional Motion Estimation

I. INTRODUCTION

It is well known that motion compensation (MC) is a good means for realizing high coding efficiency for motion picture coding [1]. Therefore, motion compensation is often employed in hybrid coding schemes along with orthogonal transforms. For example, the combination of Discrete Cosine Transform (DCT) [2, 3] and motion compensation is used in the CCITT / ISO standards. However, they often produce annoying blocking effects on decoded pictures at relatively low bitrate. This is because most of the orthogonal transforms are block based ones and are applied to block segmented pictures. It is a significant problem especially for a high quality picture coding such as the one for HDTV.

Therefore, much research activity is reported in new transforms, such as Lapped Orthogonal Transform (LOT) [4], Modified Discrete Cosine Transform (MDCT) [5] and Wavelet Transform [6, 7, 8, 9]. As these transforms use overlapped block segmentation, they generate little or no blocking effects on decoded pictures even at very low bitrate.

However, the conventional motion compensation, which still employs non-overlapped block segmentation, generates interfering blocking edges in the prediction error signals.

This paper describes an overlapped block motion compensation scheme, which is found in our advanced researches for high quality picture coding [9, 10, 11].

II. CONVENTIONAL MOTION COMPENSATION PROBLEMS

In a conventional motion compensation scheme, input signals are segmented into non-overlapped blocks. Figure 1 shows a conventional motion estimation scheme. Generally, motion vector estimation is operated using a block matching method, and one motion vector per block is obtained. Figure 2 shows a conventional motion compensated prediction scheme. As shown in the figure, the motion compensated prediction is carried out using the obtained motion vectors with a block by block basis. A prediction signal of a target block is obtained from the reference signal using the obtained motion vector.

Figure 3 shows an example of a prediction error signal generated by the conventional motion prediction. The conventional method often produces blocking edges in prediction signals because neighboring motion vectors are not always the same. These edges generate similar blocking edges

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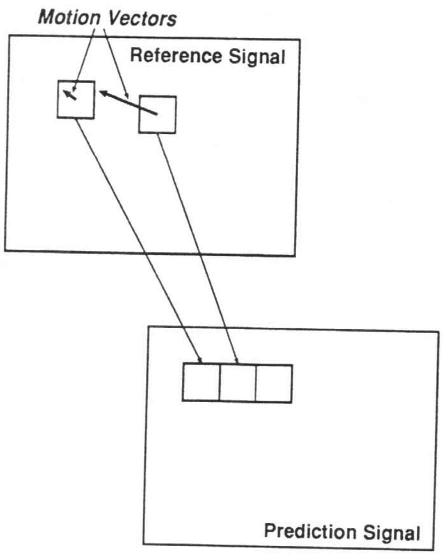

Figure 2. Conventional MC Prediction

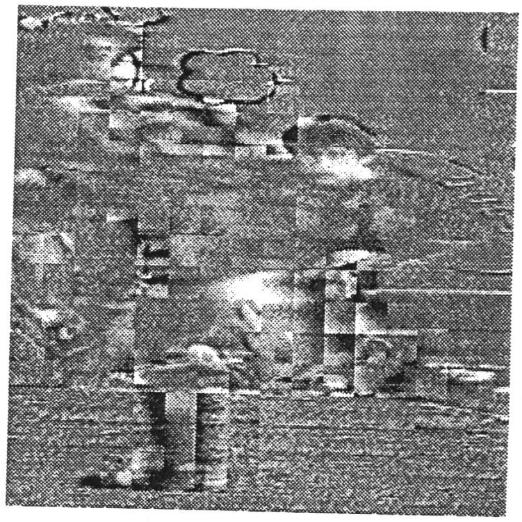

Figure 3. Prediction Error Signal with the Conventional Method

in prediction error signals as shown in this figure.

When the block based transform such as DCT is applied to the prediction error signals, these blocking edges do no harm to the coding efficiency, as far as the same block segmentation edges are used both for motion compensation and for transforms, which is the usual case.

However, when the overlapped transforms are applied to the prediction error signals, the blocking edges appear within an overlapped block and are transformed. These edges contain large signal power in the high frequency domain of transformed coefficients and coding efficiency is reduced.

Therefore, effective reduction of the blocking edges in the prediction error signals is required to realize the higher performance of the overlapped transforms and to realize higher coding efficiency.

III. OVERLAPPED BLOCK MOTION COMPENSATION

An enlarged and overlapped block segmentation is employed in the proposed motion compensation scheme to reduce the blocking effects of prediction signals. A window function is also employed both for motion estimation and for motion compensated prediction to enable the overlapped block segmentation.

A. Overlapped Block Motion Estimation

First, an overlapped block motion estimation is carried out. Figure 4 shows the overlapped block motion estimation scheme. The input signals are segmented into overlapped blocks. Using a candidate motion vector v_i, a prediction signal of a target block $P_{v_i}(x,y)$ is obtained with the enlarged block size, where (x,y) is the position within a block and the block size is $n \times n$. A prediction error signal $E_{v_i}(x,y)$ is obtained by taking differences between the prediction signal $P_{v_i}(x,y)$ and the input original signal $S(x,y)$.

$$E_{v_i}(x,y) = P_{v_i}(x,y) - S(x,y) \tag{1}$$

A window function $W(x,y)$ is operated to the prediction error signals. The window function is designed to decay towards block boundaries. In this case, a window function of a sine shape is employed.

$$E_{v_i W}(x,y) = E_{v_i}(x,y) \times W(x,y) \tag{2}$$

The candidate motion vectors are evaluated using the window operated prediction error signals $E_{v_i W}(x,y)$. In this case, mean absolute error (MAE) is used for calculating the signal power. In this way, one motion vector per block, which generates the least power, is determined.

$$MAE = \frac{1}{n^2}\sum_{x=1}^{n}\sum_{y=1}^{n}|E_{v_i W}(x,y)| \tag{3}$$

B. Overlapped Block Motion Compensated Prediction

Next, an overlapped block motion compensated prediction is carried out. Figure 5 shows the overlapped block motion compensated prediction scheme. The overlapped block segmentation of the input signals, which is the same segmentation as the overlapped block motion estimation, is employed. A prediction signal of a target block $P_v(x,y)$ is obtained from a reference signal using the estimated motion vector v with the enlarged block size. The window function

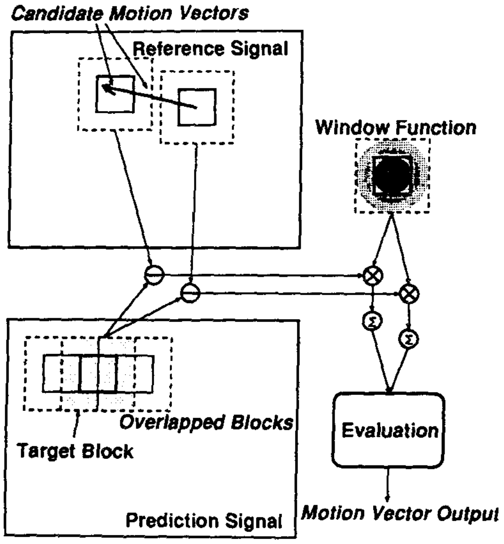

Figure 4. Overlapped Block Motion Estimation

$W(x,y)$, which is the same as the one for the motion estimation, is applied to the block prediction signal and the final prediction signal $P_vW(x,y)$ is obtained.

$$P_vW(x,y) = P_v(x,y) \times W(x,y) \tag{4}$$

The whole prediction signal is generated by summing up these window operated prediction signals of neighboring blocks. Prediction error signals are obtained by taking differences between the prediction signal and the original signal.

With the overlapped block motion estimation and the overlapped block motion compensated prediction, prediction signals without blocking edges are obtained.

IV. COMPUTER SIMULATIONS

The proposed overlapped block motion compensation scheme was evaluated by computer simulations. Both conventional and proposed methods were evaluated for motion estimation and motion compensated prediction. A simple prediction without motion compensation was also simulated for a reference.

The luminance signal of an HDTV sequence was used as a source signal of the simulations. Interframe prediction was employed for the prediction method. The block size of 16×16 was employed for the conventional motion compensation, and the block size of 32×32 was employed for the overlapped block motion compensation. The maximum area size of the both motion estimations was ±15×±15.

Figure 6 shows the prediction error signal obtained with the overlapped block motion compensation. It is observed that the blocking edges shown in Figure 3 are largely reduced. Figure 7 and 8 show FFT analysis results of the prediction error signals obtained by the conventional method and by the proposed one, respectively. It is also seen that the high frequency coefficients, which represent blocking edges in the spatial domain, are reduced.

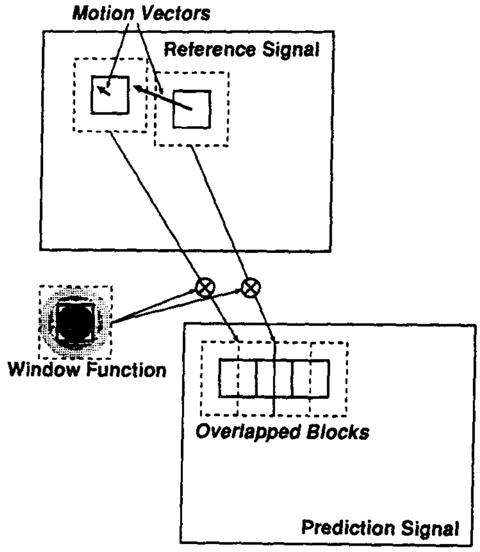

Figure 5. Overlapped Block MC Prediction

Table 1. Prediction Error Signal Power per Pel (MSE)

		Motion Estimation	
		Conventional	Overlapped
Motion Compensated Prediction	none	754.4	
	Conventional	256.8	259.0
	Overlapped	215.8	210.1

Table 1 shows the prediction error signal power per pel (Mean Square Error). With the overlapped block motion estimation and prediction, the power of the prediction error signal is reduced about 19% from 256.8 to 210.1. From this fact, it can be also said that the overlapped block motion compensation scheme is effective not only for the overlapped transforms but also for the non-overlapped block based transforms such as DCT.

V. CONCLUSION

An overlapped block motion compensation scheme is proposed. The proposed scheme is especially suitable for hybrid coding schemes combined with overlapped transforms, such as LOT, MDCT and Wavelet Transform. Computer simulations have been carried out and it has been certified that the overlapped motion compensation reduces blocking edges of prediction error signals and also reduces the power of prediction error signal about 19%. This new motion compensation scheme promises higher coding efficiency not only for the overlapped transforms but also for the conventional block based transforms.

ACKNOWLEDGMENTS

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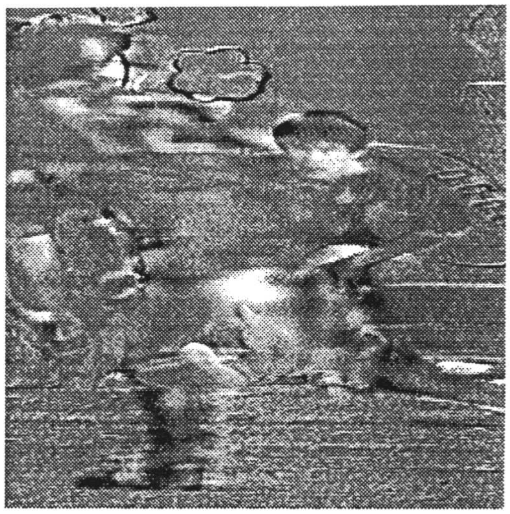

Figure 6. Prediction Error Signal with the New Method

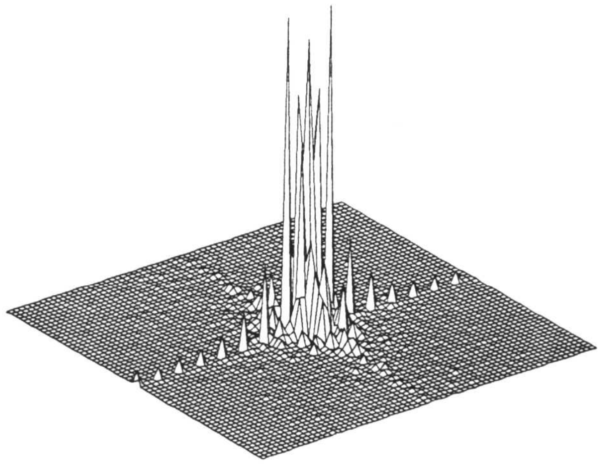

Figure 7. FFT Result of Prediction Error Signal (Conventional MC)

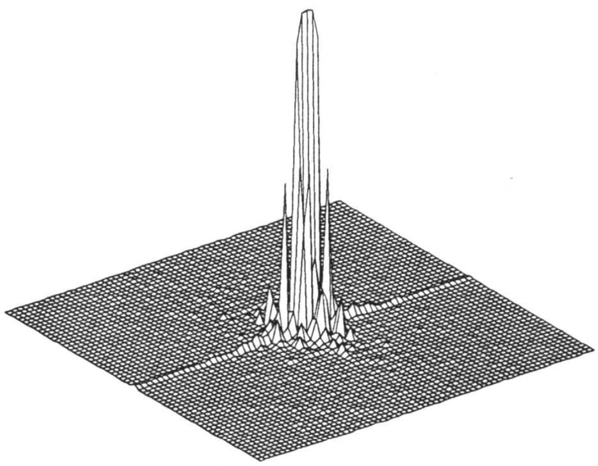

Figure 8. FFT Result of Prediction Error Signal (Overlapped Block MC)

A HYBRID DCT CODEC: PERFORMANCE AT LOW BIT-RATES AND IN THE PRESENCE OF TRANSMISSION ERRORS

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Abstract - Two papers were presented in the International Symposium on Circuits and Systems of 1990 and 1991: they describe the codecs developed and implemented in the framework of the European Project Eureka EU 256. The main goal of the project is the transmission of TV and HDTV signals for contribution applications. However, the codecs have demonstrated their inherent flexibility in terms of operative bit-rate and applications: the studies on hybrid DCT (Discrete Cosine Transform) are continuing to cope with other applications, e.g., SNG (Satellite News Gathering) and secondary distribution. Further optimizations are required to operate at lower transmission rates and a particular attention must be given to the performance in the presence of severe transmission errors. This paper up-dates the information on the developments and standardization activities and outlines the future developments.

I. INTRODUCTION

Compression techniques are applied to allow the digital transmission and distribution of TV and HDTV signals. The implementation of such techniques is becoming feasible. The standardization process of a system for transmission of TV signal suitable for contribution purposes has been terminated. The first codec based on this standard will be on the market in the course of this year.

Meanwhile, various proposals have been presented, and in some cases implemented, for the distribution of TV and HDTV signals to the viewer. Most proposals are based on algorithms similar to the system adopted for the contribution purposes.

The characteristics required for a system for distribution purposes are: high compression ratio and good subjective quality, also when the channel conditions are not perfect (i.e., high error rate due to noise, reflections, distortions), commonality across delivery and recording media and across various levels of TV system performance. Finally, the receiver should be implementable with a limited number of LSI, to reduce the costs and allow a fast grown of the new service, when it is introduced.

II. STANDARDIZATION ACTIVITIES

As reported in the previous paper [1], presented at the ISCAS '91, the work on standardization of bit-rate reduction system for the transmission of television signals is carried out in Task Group CMTT/2. CCIR Rec. 723 specifies the parameters for the transmission of 4:2:2 signals for contribution purposes at the Third Hierarchical Level of CCITT Rec. G.702, i.e., 34 and 45 Mbit/s. This standard coincides with that specified by the ETSI (European Telecommunications Standards Institute).

The verification of the performance of the system, implemented in an actual codec, is planned for the Spring '92, and in order to adopt the Recommendation in May '92. The EBU (European Broadcasting Union) is planning to adopt this system, in the Eurovision Network, for the satellite transmission at 34 Mbit/s of 4:2:2 and composite TV signals.

During the meeting of March '91, TG CMTT/2 appointed a Special Rapporteur, and a group of experts (known as SRG), to define the functional requirements for distribution and to co-ordinate the collaboration among the members of CMTT/2 and other groups of experts working in this field. The target is the definition of a draft Recommendation by the end of the present study period (1990-94) for TV and by the end of 1996 for HDTV distribution. Other standardization groups actively operating in the field of distribution of digital TV to the final users are the ISO/IEC Working Group JTC1/SC2/WG11-MPEG and the group of experts of Study Group XV of CCITT. MPEG should define, by the end of 1992, a generic standard for coded representation of audio-visual information at bit-rates up to about 10 Mbit/s. SG XV is working in the field of video compression for ATM (Asynchronous Transfer Mode) networks, mainly in the area of conversational services.

III. SECONDARY DISTRIBUTION AND COMPATIBILITY

The main goal of the SRG is the definition of a compatible system for secondary distribution to code Conventional definition TV (CTV), Enhanced Definition TV (EDTV) and High Definition TV

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(HDTV) [2]. To achieve this goal, the first step is the identification of the formats to handle.

CCIR Rec. 709 specifies as objective for the system to be adopted in studios the progressive scanning, i.e., 1:1 interlace ratio. For current implementations, an interlace ratio of 2:1, or an equivalent sample rate reduction process, may be used.

It is likely that a future service of digital TV will present a few new improvements, easy to be appreciated by the viewers.

The quality associated with the HDTV should be outstandingly higher than that provided by present analogue standards: 16 by 9 aspect ratio and a definition four times better (1920 pels per line and twice the number of lines per field).

However we can also hypothesize the adoption of an EDTV format, sufficient for most of the uses. The EDTV could be characterized by a 16 by 9 aspect ratio, half the horizontal and vertical resolution of the HDTV, possibly a 1:1 interlace ratio. The perceived quality would be significantly higher than that of present analogue services. The definition would match that offered by current cameras and displays and the production costs for the programs would be lower than those for HDTV.

In such a scenario, CTV, characterized by a 4 by 3 aspect ratio, will be less appealing for new productions. Therefore, for simplicity, we can limit our considerations to EDTV and HDTV.

A possible compatible scheme for digital distribution of EDTV/HDTV signals is depicted in Figure 1.

A pyramidal encoding, based on the use of hybrid Discrete Cosine Transform (DCT), can be a candidate to fulfill the requirements of compatibility, efficiency, good subjective quality and is sufficiently easy to be implemented with the present technology.

IV. THE EUREKA 256 ALGORITHM

Eureka 256 has devised and implemented the coding scheme for contribution-quality applications, to operate at a compression factor not higher than 6, i.e., 2.5 bit/pel. However the scheme has demonstrated to provide a quality sufficient for distribution purposes up to compression factor of about 15, i.e., about 1 bit/pel.

The future digital system for EDTV and HDTV could then be similar to the systems developed for contribution purposes, e.g., the EU 256 system.

This system is extensively described in the previous contribution to ISCAS and in various papers and articles, e.g., [3]. The EU 256 codec is based on the algorithm adopted by ETSI and in CCIR Rec. 723.

Intra-field, inter-field and motion compensated (MC) inter-frame DCT exploits the spatial and temporal redundancy. Entropy encoding, VLC (Variable Length Coding) and RLC (Run Length Coding), exploits the non-uniform distribution of DCT coefficients.

V. THE CODECS AND THEIR PERFORMANCE

Two codecs are available, produced by Alcatel-Telettra. One codec is intended for the transmission of CTV, component and composite: it is based on the ETSI specifications.

The other codec is for the transmission of HDTV signals. The new version, equipped with motion evaluation and compensation, has been demonstrated for the first time at the Telecom in Geneva, October 1991. It can operate at various transmission rates, from 140 Mbit/s, to provide contribution quality, down to 45 Mbit/s, with a quality that we can consider sufficient for distribution purposes.

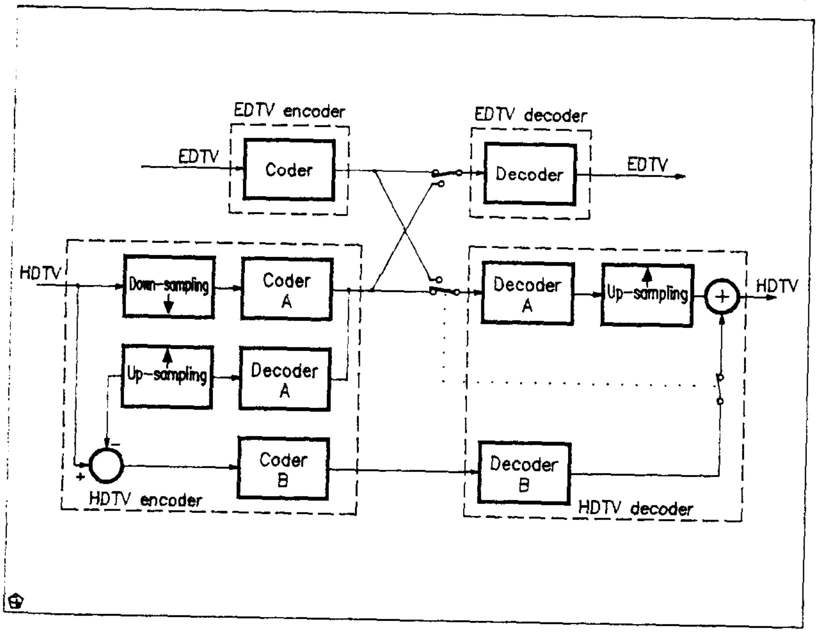

Figure 1: An input HDTV signal is down-sampled (horizontally and vertically, e.g., by a factor of 1/2), and a component compatible with the EDTV format is obtained. This component is coded by Coder A and the residual component (difference between input HDTV and locally-decoded output of Coder A) is coded by Coder B. At the HDTV decoder, upsampling is carried out for the output of Decoder A and the resulting information, combined with the output of Decoder B, reconstructs the whole HDTV picture. We achieve: downward compatibility when coded data from Coder A of HDTV is decoded by the EDTV Decoder; upward compatibility when coded data from the EDTV Coder is decoded by Decoder A of the HDTV receiver.

A further, and significant, improvement will be achieved when the motion evaluation is computed with ½ pel accuracy (Figure 2). Presently the accuracy is limited to 1 pel.

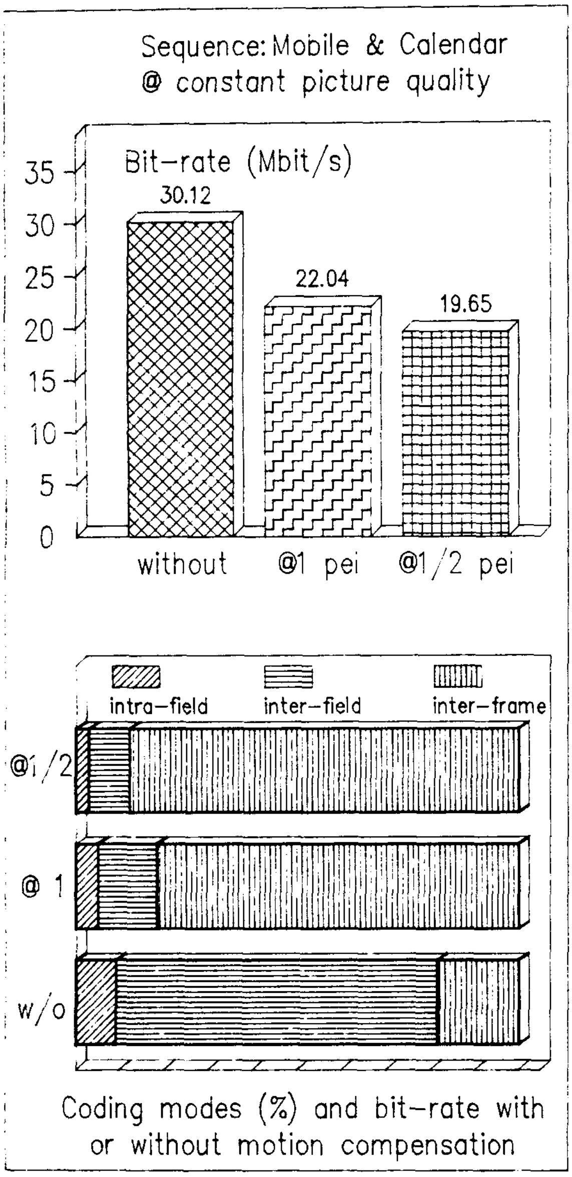

Figure 2: Results of computer simulations on the sequence "Mobile and Calendar" to evaluate the gain obtained by using the motion compensation. The transmission factor (i.e., the picture quality) is maintained constant and the resulting bit-rate is computed when no motion compensation is applied, when the motion is evaluated at 1 pel of resolution or at ½ pel. The percentages of coding modes in the three cases are also shown. A better prediction, due to the motion compensation, increases the use of the inter-frame mode and provides a gain in terms of bit-rate up to 35%, depending on the picture content.

VI. OPTIMIZATION AT LOW BIT-RATES

The requirements for distribution applications are different from those for contribution applications. We must take into account this fact in defining the new system.

The encoder complexity can be much higher than that of the decoder. In the decoder, we can accept a more complex circuitry for the signal processing, due to the increasing integration factor. On the other hand, we should keep limited in the decoder the quantity of memory, already highly integrated.

It has been demonstrated that the adoption of a-posteriori choices at the encoder provides a significant improvement in the performance [4].

A better prediction can reduce the amount of transmitted information. The addition of the motion compensation for the interfield prediction is considered promising.

The use of the EU 256 codec at low bit-rates has confirmed that the source material characterized by a high level of noise is very critical. Therefore the introduction of a denoiser is beneficial for distribution purposes, even though it can limit the perfect reproduction of higher frequencies.

Adaptive quantization, non-uniform quantizer with threshold and reconstruction values optimized for low bit-rates can provide further improvements.

An efficient coding of the DC coefficients and an optimized reassignment of the VLC words (B-code words) according to the statistics of the quantized DCT coefficients and run-length of zeros at low bit-rates is very important. Computer simulations indicate that the VLC reassignment increases the coding efficiency of about 9, 15, 25 % at a bit-rate of 1, 0.7, 0.5 bit/pel, respectively.

Moreover, we should consider the possible use of a progressive, or pseudo-progressive, format on the coding efficiency, delivered quality and compatibility with other applications.

VII. RESILIENCE IN THE PRESENCE OF ERRORS

The techniques to protect the data from channel errors are forward error correction (FEC) and concealment. In the EU 256 system the data are organized as octects protected by a Reed Solomon code RS(255,239) with interleaving factor two. This allows the correction of up to 8 symbols of 8 bits on a block of 2040 bits and error bursts up to 128 bits on a block of 4080 bits. The redundancy is less than 7%.

FEC is very effective in reducing the number of errors, even in the presence of bursts, however, when it fails the numbers of residual errors become overwhelming. The failure characteristic, quality versus BER, of such a digital transmission system is very steep. Practically there is a threshold: for BER lower than 10^{-5}, the quality is unimpaired, for BER higher than 10^{-3}, the service is no more available [5].

To provide a smoother failure characteristic, a procedure apt to detect residual errors and to apply a concealment technique is very attractive.

To this purpose, it is convenient to evaluate the sensitivity of individual bits to random errors. In fact, the effect of errors on the picture quality can range from catastrophic to imperceptible.

In the video framing implemented by EU 256, we can identify two classes of data: those coded with fixed length words (FLC) and those coded with variable length words (VLC) [1].

The FLC data include:

- a palette of motion vectors applied on the whole field;
- the coding mode (intra-field, inter-field and motion compensated inter-frame) applied on single macro-blocks, i.e., two luminance 8 by 8 blocks and the corresponding two chrominance blocks;
- the address to select the motion vector to apply to each macro-block;
- the transmission factor (relevant to the quantization precision applied on the entire stripe, i.e., a set of 8 video lines, both luminance and chrominance).

From theoretical investigations and computer simulations, the FLC data, particularly the palette and the coding modes, result very sensitive to errors. A strong protection is recommended for such information, corresponding to about 0.05 bit/pel. A more powerful FEC or double transmission would greatly improve the resilience against errors.

The VLC data include the DCT coefficient values, the run-length of zeros and the symbols identifying the end of each block (EOB). We should design the VLC to maximize the efficiency and to minimize the error propagation. The B-code, adopted by EU 256 and in CCIR Rec. 723, satisfies this requirement.

Two VLC words are assigned to the EOB and the sequence of the EOBs inside a video frame is generated following an appropriate pattern. A deletion or a creation of an EOB word is easily detected.

When the blocks, or macro-blocks, affected by residual errors are identified an easily implementable concealment strategy is applicable. In fact, the erroneous information can be replaced with the corresponding block in the previous field or frame. This information is available in the frame memory of the receiver, since the hybrid-DCT require a prediction information.

If the correct information on the coding mode and the motion vector is available, the concealment can exploit the temporal correlation at the utmost.

The sensitivity to errors is minimum for data relevant to the high-frequency DCT coefficients: therefore detection of errors is not necessary for them. Practically only the most significant bits of the low-frequency DCT coefficients should be protected against residual errors. When errors are detected, the concealment can be applied.

VIII. CONCLUSIONS

Coding systems based on hybrid DCT with motion compensation are very effective. Good subjective quality is attainable with less than 1 bit/pel. Integrated circuits are now on the market for DCT and motion compensation; 16 Mbit RAM will be available soon to facilitate the use of large frame memories. The implementation of EDTV and HDTV decoders for the consumer market will be possible in the near future.

The digital video quality provided to the viewer could be similar to that available in studios, for most of the picture material.

We should make efforts to define the organization of data, the error correcting and concealment techniques, the modulation schemes to preserve the digital quality, even in the presence of severe impairment on the transmission channel.

The involvement of the RAI Research Center in the studies and the standardization activities is aimed to these goals.

ACKNOWLEDGMENTS

The system for video compression here described has been developed in the framework of the European Project Eureka EU 256. Partners of the project are RAI - Radiotelevisione Italiana (I), Retevision (E), Universidad Politécnica de Madrid (E), Telettra S.p.A. (I) and Telettra S.A. (E). The studies on the performance of the system and its optimization at high transmission error rates are carried out by RAI in the framework of the European Space Agency contract ESA 9100/90/NL/PM.

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OPTIMAL BUFFER-CONSTRAINED SOURCE QUANTIZATION AND FAST APPROXIMATIONS

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ABSTRACT

We formalize the description of the buffer-constrained quantization problem. For a given set of admissible quantizers to code a discrete non-stationary signal sequence in a buffer-constrained environment, and any global distortion minimization criterion which is additive over the individual elements of the sequence, we formulate the optimal solution as well as slightly suboptimal but much faster approximations. As a first step, we define the problem as one of constrained, discrete optimization and establish its equivalence to some of the problems studied in the field of integer programming. Dynamic programming using the Viterbi algorithm is shown to provide a way of computing the optimal solution. Finally, we provide a heuristic algorithm based on Lagrangian optimization using an operational rate-distortion framework that, with much-reduced computing complexity, approaches the optimally achievable SNR.

1. INTRODUCTION

The application of variable bit rate (VBR) techniques to non-stationary sources, such as video sequences, in a constant bit rate (CBR) transmission environment, requires the definition of buffer control policies. Recently, the need to address the problem of buffer-constrained quantization in the context of image and video coding has risen sharply to the point where buffer-control algorithms are being proposed for the MPEG video standard [1]. Many applications like CD-ROM storage of images and video sequences, windows applications for workstations, buffer-limited JPEG [2] coding, and MPEG buffer control strategies are *non real-time* finite-buffer constrained coding applications where computationally expensive methods are not taboo if the complexity-performance tradeoff is worthwhile, specially if the one-time coding complexity can reduce transmission cost, e.g. limiting the amount of buffer memory needed by the user.

This provides the motivation to investigate optimal quantization strategies for the coding of signal sequences in a finite buffer environment, and to quantify the performance tradeoffs involving key design parameters like buffer size or buffer occupancy "granularity". The possession of an optimal solution can also be an invaluable benchmark for assessing the performance of real-time constrained and practical coders as well as for quantifying the suboptimality of fast heuristics. In asynchronous network applications (Asynchronous Transfer Mode or ATM networks), the idea of "self-policing" by the user, to guarantee conformance with the negotiated transmission parameters while ensuring an optimal grade of quality delivered, can also be very appealing, as the user has more "control" over the quality of service he can expect from the network. Besides, negating the value of buffer-control algorithms by making the encoder resort to a large enough buffer size to absorb all source bitrate variations may not only be unacceptable because of end-to-end delay restrictions, but also economically unwise even when delay is not an issue, as there may exist "smarter" shorter-buffer solutions that yield the same performance.

We formalize the generalized problem of buffer-constrained independent quantization of a sequence and describe how, given a set of quantizers, a finite buffer, and any additive cost measure over the sequence elements, an optimal solution can be found [3]. We show how this problem, one of discrete optimization with constraints, can be construed as a deterministic dynamic programming problem with the Viterbi algorithm used to compute the optimal solution. After drawing parallels between this buffer-constrained quantization problem and the less complex budget-constrained unbuffered quantization problem [4], we present a recursive Lagrange-multiplier based algorithm that provides a fast nearly-optimal solution with much reduced complexity. For simplicity, we use the mean-squared-error (MSE) distortion criterion in our simulations, though any additive criterion is admissible in general. The source sequence elements (8x8 pixels image blocks in our simulation) are quantized with the JPEG coder [2] using a finite number of quantization scales as the admissible quantization set.

2. PROBLEM DEFINITION

2.1. A First Formulation

Let us consider a sequence made of blocks, representing discrete analog samples or sets of samples (depending on the application), that are to be coded independently, possibly after some unitary transformation. For a given finite set of quantizers, the problem consists in choosing, for each block,

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that quantizer within the admissible set, that will minimize the *global* cost of coding the sequence, where the cost is additive over the independent individual blocks.

Our system consists of three basic elements: the encoder, the decoder (each including a buffer, See Figure 1) and the transmission channel. In the general case, although transmission need not be synchronous (e.g. video transmission over ATM networks), it can be seen that, since the encoder and the decoder are usually attached to synchronous devices, a constant delay restriction exists between the input to the coder and the output of the decoder. See Figure 2. As a consequence, given the constant delay through the system and the finite channel rate, we conclude that the buffer size will be finite.

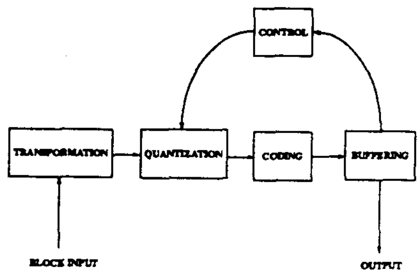

Figure 1: Block diagram of the encoder

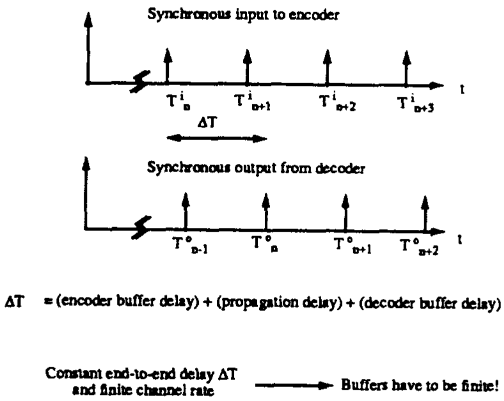

Figure 2: Synchronous devices at encoder and decoder in a system with limited channel rate imply constant end-to-end delay and therefore finite buffering.

We can now formulate the problem of optimal buffer-constrained quantization:

Formulation 1 *Given a set of quantizers, a finite buffer, an average throughput rate, and a sequence of blocks to be coded independently, select the optimal sequence of quantizers corresponding to each block such that the total cost measure is minimized and buffer overflow is avoided.*

2.2. An Integer Programming Formulation

A fundamental facet of this problem is that the set of available quantizers is finite in size, which discretizes the set of admissible solutions and makes it natural to look at integer programming (IP) formulations [5, 6] to help solve it.

Consider the allocation for N blocks and suppose there are M quantizers available to code each block. To denote the use of a given quantizer, the set of variables to be optimized is defined as: x_{ij}, which is 1 if quantizer j is used for block i and is 0 otherwise, for $i = 1, \ldots, N$ and $j = 1, \ldots, M$.

Let d_{ij} and b_{ij} be, respectively, the distortion and the number of bits produced by the coding of block i with quantizer j.

Formulation 2 *(0-1 Integer Programming)*
Given B_{MAX} (the buffer size) and r (the channel constant bit rate), find values for $x_{ij} \in \{0,1\}$ to minimize:

$$D_{TOT} = \sum_{i=1}^{N} \sum_{j=1}^{M} d_{ij} x_{ij}$$

subject to:

$$\sum_{j=1}^{M} x_{ij} = 1, \quad \forall i = 1, \ldots, N \qquad (1)$$

and

$$\sum_{i=1}^{k} \sum_{j=1}^{M} b_{ij} x_{ij} - (k-1) \cdot r \leq B_{MAX}, \quad \forall k = 1, \ldots, N \quad (2)$$

Constraint (1) requires that only one quantizer is used for each block, while constraint (2) is the overflow restriction. Note that underflow will be avoided under the condition that distortion has to be minimized.

3. DYNAMIC PROGRAMMING SOLUTION USING THE VITERBI ALGORITHM

3.1. The Viterbi Algorithm

This problem can be solved using dynamic programming (DP) and, in particular, the Viterbi algorithm [7, 8], can be employed. The basic idea consists of starting at the initial buffer state and growing a path for every admissible quantizer (that does not cause buffer overflow), resulting in a trellis diagram whose states are the buffer-occupancy levels. See Figure 3. Each trellis path, corresponding to a quantizer choice, has a cost associated with it corresponding to the distortion incurred by the quantizer while the quantizer's coding bitrate dictates the destination state of the path. For an additive cost function, the well-known Viterbi algorithm provides the optimum choice of quantizers to code the sequence. This technique establishes a rule to prune out the suboptimal paths in the trellis: if a node can be reached by more than one path, only the minimum cost path will be kept.

3.2. The Optimal Solution

By investigating the optimal solution, one can study the characteristics of the optimal system configuration, and use it, as motivated earlier, as a benchmark for evaluating existing buffer-control practices and heuristics. For example,

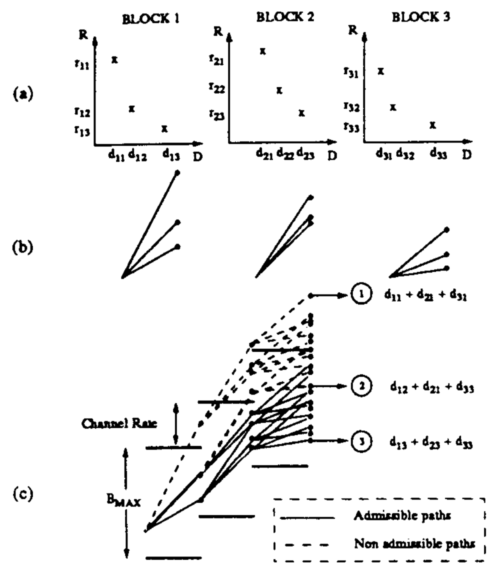

Figure 3: The problem seen from the Viterbi algortihm point of view: (a) The rate-distortion characteristics of the the blocks for the available quantizers. (b) Equivalent representation. Each of the branches corresponds to the choice of a specific quantizer and has attached a cost. The length of the branch is proportional to the rate. (c) All possible paths for the three blocks considered. Path 2 cannot be used because of overflow. 1 and 3 are, respectively, the maximum and minimum distortion paths.

Figure 4 shows how, for different rates, increases in buffer size beyond a certain level produce no significant quality improvement.

As DP methods like the Viterbi algorithm are enumerative techniques with complexity that is exponential in the number of states in the trellis, it is useful to study the merits of reducing computational complexity at the cost of incurring acceptable suboptimality. Our results show that both reducing the number of nodes at each level (for example, by "quantizing" the number of buffer states - see Figure 5) and limiting the memory of the problem (i.e. confining the decision making, in releasing a branch in the path, to a finite number of consecutive sequence blocks) can reduce the complexity while not significantly decreasing the performance.

4. HEURISTIC METHODS TO APPROACH THE OPTIMAL SOLUTION

4.1. Rationale for the Heuristic Approach

To develop fast heuristics for our problem, we turn to rate-distortion theory by noting that our allocation problem *without the buffer constraint* reduces to the classical budget-constrained allocation problem cited in the literature [4], for which a fast Lagrange multiplier based solution exists.

Further, since our problem has an essentially limited memory (due to the finite size of the buffer), we can, in practice, reduce the horizon of our optimization problem to just a finite number of sequence blocks, i.e. we can decouple

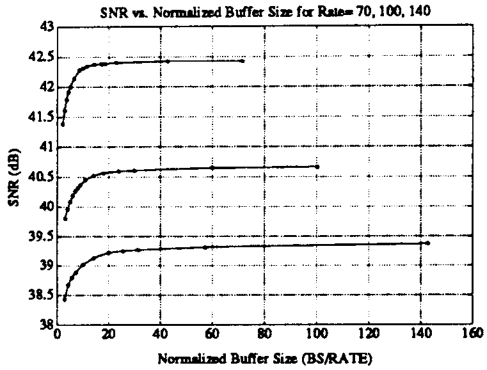

Figure 4: Optimally attainable SNR vs. normalized buffer size. Each point corresponds to the optimal quantization for the sequence at a given buffer size. Note that "rate" refers to the average number of bits used to code each source block.

the future beyond a certain point from the decision on the current block, and this without significantly reducing the coding quality (see the top curve in Figure 7).

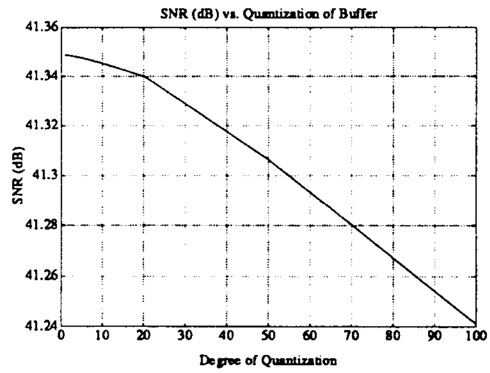

Figure 5: Effect of buffer quantization. The degree of quantization refers to the buffer state reduction factor, i.e. the number of consecutive physical buffer occupancy states that are "collapsed" into one virtual state, or node, in the trellis diagram.

4.2. Recursive Lagrangian Optimization

We combine these two ideas to formulate the following algorithm:

Algorithm 1

(step 1) At every stage k, use Lagrangian optimization [4], with budget constraint $n \cdot r + B(k) - B_{MAX}/2$, to obtain the best non-buffer-constrained allocation for the following n blocks, where $B(k)$ is the buffer occupancy level at the k-th stage as determined by the recursive algorithm.

(step 2) Use the quantizer choice found by the previous step for block k and release it to the buffer, and repeat the first step for stage $k+1$.

This is equivalent to performing a sliding window optimization so that the quantizer choice for the k-th block depends only on the rate-distortion characteristics of the following n blocks and on the buffer occupancy level at the k-th stage. Thus by exploiting the finite memory exhibited in practice by the problem (see results in Figures 4 and 7) and the fact that we can perform Lagrangian optimization very inexpensively due to the convexity of the resulting "unconstrained" case, we can approach the optimal solution. Our simulations verify that this heuristic yields solutions very close to the optimal one, as obtained using the more "brute force" Viterbi Algorithm. Experimental results have shown that, typically, using Algorithm 1 results in less than 10% of the blocks being coded with a non-optimal choice of quantizer (as computed with the Viterbi algorithm, under the same conditions).

4.3. Heuristic Improvement

A computationally efficient heuristic that sacrifices little quality is the following: use the algorithm described above, except undertake optimization over n blocks only when the algorithm results in paths whose buffer occupancy levels at any stage violate certain empirical thresholds. Call the heuristic percentage the fraction of the size of the buffer that is used as the threshold. Thus a 10% heuristic would mean that the algorithm would be recomputed when it results in any path whose buffer occupancy level is below 10% or above 90% of the total buffer size (note that a 50% heuristic would be equivalent to Algorithm 1). This algorithm can be formulated as:

Algorithm 2

(step 1) At every stage k, if the buffer occupancy is within the defined thresholds use the allocation previously computed for this block. Otherwise, use (step 1) of Algorithm 1 to compute the allocation for the following n blocks,

(step 2) Use the quantizer choice found by the previous step for block k and release it to the buffer, and repeat the first step for stage $k+1$.

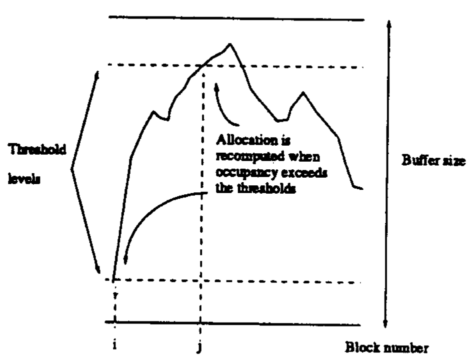

Figure 6: Algorithm 2: the allocation, using Algorithm 1 for n consecutive blocks is recomputed only when the buffer occupancy exceeds the thresholds.

In Figure 7, the SNR of both the Viterbi solution with limited memory (top curve) and the heuristic (10%) approximation are compared. For a sufficiently large number

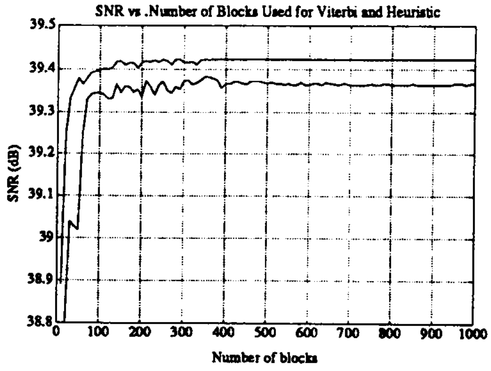

Figure 7: SNR Comparison between the Viterbi algorithm (top curve) and recursive 10% buffer-threshold Lagrangian heuristic for finite memory case. Note that the Viterbi algorithm is applied in a sliding window fashion, i.e. a on on a quantizer assignment is made based on the following n blocks and not on the whole sequence.

of blocks, simulations indicate that the heuristic approximation comes within 0.05 dB of the optimal value, while consuming about 1/20 of the CPU time.

5. CONCLUSIONS

In this paper, we have examined the problem of optimal buffer-constrained independent quantization for an additive cost criterion. The problem is formulated in an integer programming framework and a way of reaching the optimal solution using the Viterbi algorithm is described. The results obtained from the optimal solution are studied and a fast heuristic algorithm, based on recursive Lagrangian optimization using rate-distortion concepts, is proposed which provides a close approximation to the optimal solution with much lower computational complexity.

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COMPATIBLE CODING OF DIGITAL INTERLACED HDTV

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Abstract : This paper proposes a compatible coding scheme for interlaced HDTV. This scheme provides a standard definition signal and a second channel containing the additional information required to reach the HD resolution. The global scheme is of the pyramidal type : the standard TV is obtained by downsampling the HDTV signal. The former one is coded, decoded and upsampled. It is used as a prediction of the HD signal and the prediction error is coded in a second channel. In each branch, motion compensated hybrid DCT coding is used. In addition, the DCT is taken within the frame.

1 Introduction

As regards the digital secondary distribution CMTT/2 is presently concentrating on transmission over digital networks. Various formats may be delivered [4]. CTV (Conventional TV), EDTV (Enhanced definition TV) and HDTV (High-definition TV) formats are considered. The corresponding luminance signals are described in table 1 for 50 Hz based systems. It can be seen from this table that HDTV formats offer twice the spatial resolution of corresponding EDTV signals in both directions. CTV can be seen as a part of an EDTV-I format. An important requirement

Ref	Aspect ratio (h:v)	Active image		Interlace factor
		pels	lines	
HDTV-P	16:9	1920	1152	1:1
HDTV-I	16:9	1920	1152	2:1
EDTV-P	16:9	960	576	1:1
EDTV-I	16:9	960	576	2:1
CTV (601)	4:3	720	576	2:1

Table 1: TV and HDTV formats considered by CMTT/2 for compatible coding

has been considered : compatibility between CTV and EDTV on the one hand and HDTV on the other hand. As there is no present standard for digital distribution, this compatibility requirement is not of the backward type, i. e. it is not required that the HDTV scheme would be compatible with an existing digital CTV distribution standard. Compatibility is rather of the upward-downward type.

- downward compatibility : a receiver associated with the TV resolution should be able to extract a given part of the HDTV bit stream and display that information at its own resolution level,
- upward compatibility : a receiver associated with the HDTV resolution should be able to decode and display an image at TV resolution. Whether this image will be displayed on the full screen or a given part only is up to the receiver.

In this paper, we mainly focus our attention on the question of compatible coding of interlaced signals. We just consider EDTV-I and HDTV-I which will be referred to as TV and HDTV. For progressive formats, solutions have been proposed in [1,8,10,11]. As regards interlaced signals, solutions have been put forward in [2,12]. In the latter papers, the format conversion step is obtained by means of a subband decomposition. In this work, we propose a solution for the format conversion step which is based on the well-known pyramidal structure introduced by Burt and Adelson in [3]. In addition, the conversion of an HD frame to a TV frame is done by field processing only. This pyramid-based solution offers the advantage of less constrained filter design in comparison with subband coding, but at the price of a slight increase of the number of points : the pyramidal decomposition is a redundant representation. The TV signal obtained from HDTV is processed in the first channel. In the second channel, one encodes the difference between the original HDTV frame and the interpolated-decoded-TV frame.

2 Spectral description of interlaced signals

2.1 Mathematical description of interlaced signals

Interlaced signals can be characterized by means of the formalism developed by Dudgeon and Mersereau in [5] as multidimensional sampled signals. Let us consider the sampling lattice defined by the line positions of an interlaced signal. Such a lattice is shown by the left-hand side part of figure 1. The sampling along the horizontal direction is orthogonal to the $y-t$ domain and can be considered separately. We can see this sampling lattice as the union of 2 orthogonal lattices : a first one associated with the line positions of even fields, and the second one associated with the line positions of odd fields. The spectrum of the interlaced signal can be written as the addition of the spectra of the signals defined over the two orthogonal lattices. We know that due to the sampling, the spectrum of the analog signal is repeated around the nodes of the reciprocal lattice. It is therefore interesting to see how the periodicities associated with the signals defined over the orthogonal lattices combine together to provide the periodicity associated with the interlaced signal. Let us denote by $\mathbf{V}_i$ the sampling matrix associated with the interlaced signal and $\mathbf{U}_i$ the

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sampling matrix in the reciprocal domain. We can chose

$$\mathbf{V}_i = \begin{bmatrix} 2T & T \\ 0 & H \end{bmatrix} \quad \mathbf{U}_i = \begin{bmatrix} 1/2T & 0 \\ -1/2H & 1/H \end{bmatrix} \tag{1}$$

For both orthogonal lattices, we have $\mathbf{V}_t$ and $\mathbf{U}_t$ given by

$$\mathbf{V}_t = \begin{bmatrix} 2T & 0 \\ 0 & 2H \end{bmatrix} \quad \mathbf{U}_t = \begin{bmatrix} 1/2T & 0 \\ 0 & 1/2H \end{bmatrix} \tag{2}$$

The luminance signal of even fields is obtained by sampling the analog picture signal in lines located at $\mathbf{V}_t\mathbf{n}$. The sequence of luminance samples is then defined by :

$$x_e(\mathbf{n}) = x_a(\mathbf{V}_t\,\mathbf{n}) \tag{3}$$

In a similar way, the luminance signal of odd fields allows the definition of a sequence $x_o(\mathbf{n})$ such that

$$x_o(\mathbf{n}) = x_a\left(\mathbf{V}_t\,\mathbf{n} - \begin{bmatrix} T \\ H \end{bmatrix}\right) = x_a[\mathbf{V}_t\,(\mathbf{n} - \mathbf{n}_0)] \tag{4}$$

where $n_0 = [1/2 \;\; 1/2]^T$. If we want to combine those two sequences together to obtain the luminance signal on the interlaced sampling grid, we first have to define two new sequences. Let us characterize the interlaced sampling grid by the sampling matrix $\mathbf{V}_i$. Let us define x_1 and x_2 in this interlaced sampling grid, by

$$x_1(\mathbf{m}) = \begin{cases} x_e(\mathbf{n}) & \text{when } \mathbf{m} = \mathbf{E}\,\mathbf{n} \\ 0 & \text{otherwise} \end{cases} \tag{5}$$

$$x_2(\mathbf{m}) = \begin{cases} x_o(\mathbf{n}) & \text{when } \mathbf{m} = \mathbf{E}\,(\mathbf{n} - \mathbf{n}_0) \\ 0 & \text{otherwise} \end{cases} \tag{6}$$

Matrix $\mathbf{E}$ is called the upsampling matrix and gives the relationship which exists between the sampling matrices of the interlaced and the orthogonal sampling lattices : $\mathbf{V}_i = \mathbf{E}\,\mathbf{V}_t$. The luminance signal in the interlaced sampling grid is obviously obtained as $x_1 + x_2$. When we add x_1 and x_2 together, we obtain a signal whose spectrum can be written as :

$$\frac{1}{|\det \mathbf{V}_t|} \sum_{\mathbf{k}} X_a\{\mathbf{U}_t[(\mathbf{E}^{-1})^T\,\mathbf{\Omega}_t - 2\pi\mathbf{k}]\}\,(1 + \exp^{-2\pi j\,\mathbf{k}^T\,\mathbf{n}_0}) \tag{7}$$

with $\mathbf{\Omega}_t = \mathbf{U}_t^{-1}\,\omega$. We see that the versions of the initial spectrum repeated around the nodes with coordinates $\mathbf{k}$ such that $k_1 + k_2 = 2p$ in the orthogonal sampling grid cancel each other to produce the right periodicity in the interlaced sampling grid. Stated in other words, the two signals x_o and x_e individually suffer from aliasing but this aliasing is partially destroyed when they are added together to provide the signal sampled in the interlaced sampling grid.

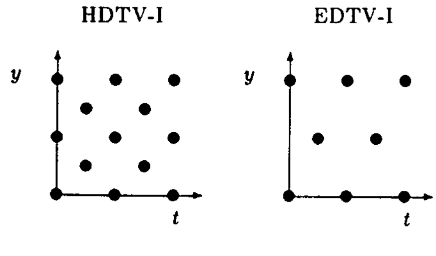

Figure 1: Sampling lattices corresponding to the line positions of HDTV-I and EDTV-I images

2.2 Choice of a signal baseband

As regards the interlaced signal, the study of the previous section shows that the nodes around which the input signal spectrum is repeated form a quincunx lattice. This means that the definition of a baseband is not unique. Actually, there is even an infinity of possible basebands.

On the other hand, we know that the pre-filtering performed in front of the camera in order to avoid aliasing effects is very poor. There isn't any normalized filter before the sampling operation to which scanning corresponds. Some spatial filtering is produced when the camera spot is defocused. The spot also produces some temporal filtering because of its remanence. We just can say that the temporal and spatial filtering operations are separated. Therefore the baseband is more likely close to a rectangle than a diamond-shaped region.

2.3 Vertical downsampling

Let us consider the sampling grids corresponding to an HD signal and its compatible version. By compatible version, we mean a signal with half the vertical and the horizontal resolution of HD. For the reasons we mentioned before, the horizontal direction is not critical. Let us focus on the $y - t$ domain. The HDTV and TV sampling grids are given by the left-hand side part and the right-hand side part of figure 1 respectively. If a field processing is considered, we see that odd and even fields have to be processed differently, because the line positions of the odd fields in the compatible image are not a subset of those of the HD odd field.

On the other hand, it is also well-known that the spatial filtering performed by the camera optics makes the vertical resolution smaller than that allowed by the number of lines. It results from a compromise between alias and spatial resolution. The ratio between vertical resolution and vertical definition is called the Kell factor and is about 0.6-0.7. Considering the formalism developed in section 2.1, the maximal vertical resolution is given by $1/2H$ if H is the distance between 2 lines within a frame. If one takes into account the Kell factor denoted by K, this resolution reduces to $K/2H$. So, if we want to produce an interlaced image with the

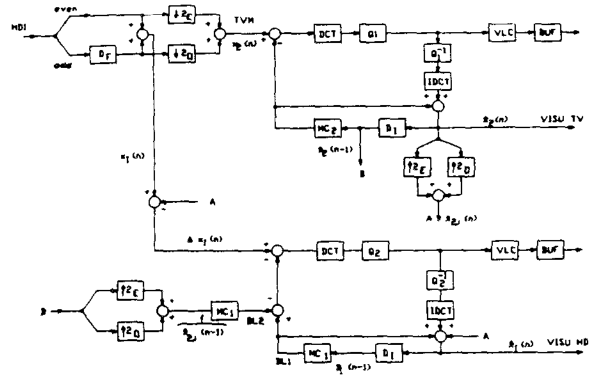

Figure 2: Global compatible coding scheme for interlaced HDTV

same Kell factor, we have to apply a vertical pre-filtering with $K/4H$ as cutoff frequency prior to the vertical decimation.

Coming back to a field-based approach, the formalism developed in section 2.1 shows that a vertical filtering of an interlaced image can be achieved by processing independently signals x_1 and x_2, i. e. the fields themselves. As regards odd fields, the filtering operation has to be combined with a resampling operation, because of the line positions.

From a realistic point of view, vertical downsampling of the HDTV signal inside fields is perfectly acceptable to provide the TV resolution since it performs the kind of anti-aliasing filtering that would be performed in front of a TV camera. It is completely free of temporal artifacts and provides a very good motion rendition.

3 Pyramidal coding scheme

Once the HDTV to TV conversion is defined, it becomes easy to define a pyramidal coding scheme producing both formats. Let us assume that HDTV is being transmitted. The input picture is first decimated to the TV format. The latter image is coded and decoded at the emitter side. This decoded version is upsampled to HDTV size (by means of a processing similar to that required for the decimation) and is used as a prediction of the HDTV image. The spatial prediction error is afterwards encoded in a second channel. Let us now describe the coding method which is applied in both loops. The global coding scheme is given in figure 2. In this scheme, the boxes $\downarrow 2_E$ (resp. $\downarrow 2_O$) stand for decimation within an even (resp. odd) field. The box D_F means a delay of one field.

3.1 Compatible channel coding

The picture to be coded in this channel has the CCIR 601 format (except as regards the aspect ratio). The coding scheme which is used on that picture is an extended version of the MPEG 2 proposal number 30. It is based on motion compensated hybrid DCT coding. In addition, the DCT is taken within the frame, i. e. the two fields of a same frame are merged together.

3.1.1 Quantization

The quantization is linear and perceptually matched to the eye sensitivity function; this means that the quantization step size of each transform coefficient depends on the eye sensitivity to this transform coefficient. This weighting of the quantization is a global adaptation. It assumes that the TV image will be displayed as a quarter of an HD image. In addition, the quantization is locally adapted to the picture content [6]. For each block of transform coefficients (8×8 blocks), a criticality value is determined. Based on this criticality measurement, the nominal quantization stepsize is increased or decreased. The criticality parameter has of course to be transmitted for each block.

3.1.2 Temporal prediction

The motion estimation is performed at HD level with half-pel accuracy. Vectors are determined for HD-macroblocks of size 16×32 (v $\times$ h). Each vector is valid for an area of size 8×16 in a TV image. This area defines the TV macroblock. In addition, an accuracy of a quarter TV pel is allowed. In figure 2, the box MC_2 stands for the TV prediction memory. The decision whether the block itself or the prediction error will be coded is taken at the TV-macroblock level. This decision is taken on the basis of an energy criterion.

3.1.3 Variable length coder

The variable length coder is the U-VLC proposed in [7]. It encodes transform coefficients of a given order taken in successive blocks. As many vectors are constituted as there are coefficient orders. In addition, each table is scanned at the bit level : one starts with the MSB's and ends with the LSB's. In the scanning, when a 1-bit is encountered, its position is sent by means of an ATRL [9] code, the less significant bits are sent uncoded and the coefficient is skipped in the following scanning steps. Performances of this VLC have been established and demonstrated in [7].

3.2 Enhancement channel

The second channel is in charge of the difference between the HDTV image and the decoded TV image upsampled to HD size. Therefore, this channel provides additional resolution and quantization accuracy about the first channel. The coding scheme is the second channel is really close to that of the first channel : it is frame DCT based, and improved with motion compensation. Let us see in more details how it works.

3.2.1 Temporal prediction

The enhancement image has to be predicted from the previous decoded TV image and the previous decoded HDTV image. For each HD macroblock, it is possible to find the prediction block in the previous HD decoded image by means of the motion vector. This block is called "BL1" in figure 2. On the other hand, the previous decoded TV image (denoted by A in figure 2) can be upsampled to HD size (it is already known, because it is the input signal of the second channel at the previous instant). In this HD image of TV resolution, it also possible to find the prediction block addressed by the HD motion vectors on an HD macroblock basis. This block is called "BL2" in figure 2. The difference between these two blocks (BL1-BL2) is used as a prediction of the enhancement macroblock to be coded. An inter/intra decision is taken for each macroblock (HD macroblock size). This decision is independent from that taken in the first channel for the same physical area (TV macroblock).

3.2.2 Quantization

The quantizer is the same as in the first layer. As regards the eye sensitivity function, when TV is displayed as a quarter of HDTV, and watched from $6 \times H_{\mathrm{TV}}$ and HDTV is watched from watched from $3 \times H_{\mathrm{HDTV}}$, i. e. the same physical distance, the viewing angle between two adjacent pels is the same in both images and the weighting should be the same. Concerning the criticality, as the DCT size is the same as in the first layer coder, it was also taken identical to that of the first channel.

3.2.3 Variable length coder

The VLC in the enhancement channel is again the UVLC.

4 Conclusions

This paper has presented an efficient solution for the compatible coding of interlaced TV and HDTV signals. It is based on a pyramidal structure, with decimation performed in the fields. The first layer provides the compatible image and the second one encodes both a complement of resolution and precision. Results are promising.

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