LiNbO₃ interlayer device for TM polarisation and the 2.5 μ m ZnS interlayer device which was polarisation insensitive within the limits of the test equipment. The response for the 10 μ m LiNbO₃ device for TE polarisation was similar in shape to Fig. 2 with the resonances displaced (by approximately 12 nm) as would be expected due to the material birefringence of LiNbO₃. The switching characteristics clearly show a periodic nature with the resonance separation increasing significantly for the thinner interlay [7]. The resonance separation for the 10 μ m LiNbO₃ interlayer device is approximately 55 nm and for the ZnS device approximately 215 nm, compared with the respective theoretical values obtained from eqn. 2 of 53 and 221 nm. An unexpected feature of the switching characteristics is the high loss (low output power from either fibre) region on the low wavelength side of each resonance. Presumably the power is confined to the interlay at these wavelengths. Further modelling of the device as a multilayer structure may provide a better understanding of this feature.

Fig. 3 Optical fibre switch characteristic

2.5 μ m zinc sulphide interlayer
— input arm
--- coupled arm

From a switch performance point of view, the important device parameters are the power transfer efficiency to the coupled arm η , the transmitted and coupled arm losses α_1 and α_2 , and extinction ratios e_1 and e_2 . These values can be obtained from the wavelength response graphs and are given in Table 1.

Table 1 IMPORTANT DEVICE PARAMETERS

Interlay material (polarisation)	η	α_1	α_2	e_1	e_2
	%	dB	dB	dB	dB
LiNbO ₃ (TM)	90	<0.5	<1	15	18
LiNbO ₃ (TE)	60	<0.5	<2	11	17
ZnS (—)	67	<0.5	<2	15	17

The values obtained for the extinction ratios are limited by the low signal to noise ratios produced by the monochromator/detection system. Other measurements taken with equipment of better dynamic range indicate that values of 25 dB or greater are attainable.

Conclusions: Efficient transfer of power between two side polished singlemode fibres via a high index interlay waveguides (10 μ m LiNbO₃ and 2.5 μ m ZnS) has been achieved by variation of the input wavelength. By tailoring the interlay thickness, it can be arranged that crosscoupling to the second fibre occurs at a desired wavelength such as 1.3 or 1.55 μ m. For the proposed switch structure, switching can then be induced by tuning the interlay mode structure via electro-optic variation of the material index (by the Pockels effect in the case of LiNbO₃), instead of wavelength variation (varying the interlay material index will result in similar tuning of the mode effective indices to that observed for wavelength variation). Hence, the feasibility of the proposed switch structure has been demonstrated. Furthermore, coupling efficiencies of up to 90%, losses in the region of 0.5–2 dB, and extinction ratios of >15 dB are obtainable. In addition to direct switching applications, the device can act as a wavelength selective/filtering

element or variable ratio coupler by biasing to the appropriate point on the switching characteristic.

The work reported in this Letter was based on passive structures for investigation purposes. The full proposed switch structure is presently being fabricated. Electro-optic switching/modulation has previously been demonstrated for a singlemode fibre in evanescent contact with an electro-optic multimode high index overlay. The results indicated that electric field strengths in the region of 4 V/ μ m are realistic for LiNbO₃ to move on/off resonance [3], implying that a switching voltage of 40 V would be possible for a switch structure with a 10 μ m interlay.

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DCT-BASED INTERFRAME CODING FOR VIDEO CODEC

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Indexing terms: Video coding, Image processing

A preprocessing method for the discrete cosine transform (DCT) in interframe coding schemes is proposed based on an observation that the statistical characteristic of block difference is quite different from that of ordinary pictures. This method makes block difference more suitable for DCTs, and significantly reduces the reconstruction errors of typical image sequences.

Introduction: The discrete cosine transform (DCT) [1] has been widely adopted in image and video coding systems. In video coding systems that process image sequences, the DCT is usually combined with motion estimation and DPCM to form interframe coding schemes. In these coding systems, the DCT processes the difference between two matched blocks in successive frames rather than original picture blocks. The interframe mode of the CCITT H.261 standard is a typical example [2].

The performance of DCT approaches that of the optimal KL transform for highly correlated data, and on the other hand, the DCT has poorer performance than several sinusoidal transforms when the correlation between data is not so

high [3]. The DCT performs excellently in intraframe coding because neighbouring pixels in ordinary pictures are usually highly correlated, but this is not so for block differences, where the statistical characteristics are quite different from those of the original pictures. Our approach is to process the block difference to raise the spatial correlation and thus promote the efficiency of the DCT, instead of considering other transforms for these data with low correlation. This enables us to improve the performance of well-established systems by only minor modifications.

Proposed approach: As shown in Fig. 1, a preprocessing unit to raise the data correlation is inserted just before the DCT, and its inverse operation is added just after the IDCT to recover the data. Several types of preprocessing method are considered and evaluated. A simple but effective method is taking the absolute values of input data and assigning a bit of flag for each number to indicate the sign of the original value. The simulation result of this method is presented below.

An exact-bit emulator of a CCITT H.261 source coder/decoder is built based on Reference Model 8 (RM8) [4], and then the proposed preprocessing and post-processing functions are applied to compare the reconstruction errors. Test sequences include two low bit rate video benchmarks, 'Salesman' and 'Claire', a sequence of random generated blocks, and a computer generated sequence that simulates the propagating ripples. Fig. 2 illustrates the effect of the proposed pre-

processing. Fig. 2a shows the DCT coefficients of a typical difference block of sequence 'Salesman'. After taking the absolute values of input data, the energy of the DCT coefficients is concentrated toward the left upper corner, as shown in Fig. 2b. The input data can now be transmitted and recovered by reserving fewer DCT coefficients, for a specified accuracy. Similar results are found in the cases of 'Claire' and random data. On the other hand, the effect is not so significant for a ripple sequence. This is because the data are extremely regular in both pattern and motion, which results in a highly correlated block difference that is suitable for DCT processing.

Results and conclusion: Table 1 compares the original system and proposed modification by showing the reconstruction errors of each sequence when various number of DCT coefficients are reserved. The proposed preprocessing provides average MSE reductions of 66.8, 29.3 and 76.8 for sequences 'Salesman', 'Claire' and random data, respectively. In the case of 'Ripple', the original system performs quite well and thus the proposed modification improves only when very few coefficients are reserved. These results show that although pixels in ordinary pictures are highly correlated and can be efficiently compressed by DCT, the block difference results from motion estimation and the DPCM loop behaves more like random data. The proposed preprocessing effectively makes block difference more suitable for DCT processing, and thus significantly reduces the reconstruction errors of typical image sequences.

As mentioned above, a drawback of the preprocessing is the bit map of sign flags. The bit map can be compressed by two-tone coding schemes to reduce the overhead. The results in Table 1 show that even considering the overhead, the preprocessing still improves the accuracy under certain bit rates, or reduces the bit rate under specified accuracy. To eliminate the overhead, the bit map can even be predicted in the post-processing by using the spatial correlation of the original pictures. In this predictive scheme the improvement of preprocessing may be counteracted by the prediction errors,

Fig. 1 Interframe coding scheme with pre/postprocessing for DCT and IDCT

Fig. 2 Example: DCT coefficients of block difference and absolute values of block difference
a Block difference
b Absolute values of block difference

Table 1 SIMULATION RESULTS

Number of reserved coefficients	Reconstruction errors (MSE)							
	Salesman		Claire		Random		Ripple	
	Original	New	Original	New	Original	New	Original	New
1	20.07	9.64	3.95	2.41	36.08	9.11	12.00	3.61
3	17.65	7.99	3.29	1.90	34.95	8.82	8.69	2.98
6	15.47	6.84	2.74	1.56	33.24	8.38	3.42	2.54
10	13.53	7.79	2.27	1.26	30.92	7.80	0.79	1.90
15	11.50	4.99	1.71	1.01	28.05	7.07	0.25	1.09
21	9.77	4.14	1.21	0.78	24.60	6.21	0.15	0.53
28	7.08	3.41	0.83	0.59	20.57	5.19	0.10	0.26
36	5.01	1.64	0.52	0.41	16.05	4.03	0.076	0.16
43	4.13	1.14	0.32	0.27	12.08	3.02	0.056	0.10
49	3.47	0.79	0.19	0.18	9.64	2.16	0.040	0.07
54	2.95	0.54	0.11	0.11	5.76	1.44	0.026	0.04
58	2.35	0.34	0.06	0.06	3.46	0.86	0.016	0.02
61	1.95	0.19	0.02	0.02	1.72	0.42	0.008	0.01
63	0.67	0.07	0.01	0.01	0.57	0.14	0.003	0.004
64	0	0	0	0	0	0	0	0

and thus more sophisticated techniques should be applied to avoid the error accumulation and propagation.

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SIMPLE AND EFFECTIVE APPROACH TO ARRAY BEAMFORMING IN PRESENCE OF STEERING ERROR

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Indexing terms: Antenna arrays, Adaptive antennas

A simple and yet effective approach is presented for array beamforming in the presence of steering error. This approach consists of two stages. The first stage is the Capon beamformer. The second stage maximises the residual power in the vicinity of the look direction using an iterative first order Taylor series approximation of the steering vector for estimating the steering error.

Introduction: Optimal beamforming with multiple linear constraints is now a well established technique. In the simplest case, a single linear constraint is imposed, namely unity gain response in the look direction [1]; the weight vector is then calculated by minimising the beamformer output power subject to this constraint. For the beamformer to reject noise and interference without rejecting the signal, it is necessary to make some assumptions about the signal characteristics. In actual operating conditions, the assumptions of plane wave signals, perfectly matched sensor channels, exact knowledge of sensor locations and perfect match between assumed look direction and actual direction of arrival of the desired signal do not hold. Under these nonideal conditions, the performance of the single linear constraint beamformer degrades rapidly [2]. To combat against beamsteer error, the use of multiple linear constraints [2-4] and quadratic inequality constraints [5] has been described in the literature. The method of artificial noise injections [6] and hybrid method [7] have also been proposed by various researchers. Recently, a robust adaptive array based on signal subspace approach has been proposed in Reference 8.

In this Letter, a simple and effective approach to array beamforming in the presence of steering errors is presented. The basic idea is based on the fact that the output power of an optimised beamformer has a maximum if the steering vector coincides with that of the desired signal. Hence, the approach iteratively searches for the correct steering vector by maximising the array mean output power with respect to the steering error. By approximating the steering vector by its first order Taylor series expansion in terms of the steering angle, the maximisation reduces to a one-dimensional optimisation problem. The improved performance of the approach is illustrated with numerical examples.

Simple and effective approach for robust beamforming against steering error: In the linearly constrained minimum power

beamforming technique [2], the processor weights are chosen to minimise the processor mean output power subject to a unity gain response in the look direction as an indirect way of rejecting noise and interference incident on the array. Mathematically, the optimum weight vector w is the solution to the following constrained optimisation problem:

$$\min_w w^H R w \quad \text{subject to } w^H s(f_0, \theta) = 1 \quad (1)$$

where R is the array covariance matrix defined by $R = E\{x(t)x^H(t)\}$ with E denoting the expectation operator, and $s(f_0, \theta)$ is the N -dimensional steering vector $s(f_0, \theta) = [e^{j2\pi f_0 \tau_1}, \dots, e^{j2\pi f_0 \tau_N}]^T$ where $\{\tau_i, i = 1, 2, \dots, N\}$ are the propagation delays between the plane wavefront and the antenna elements. Using the method of Lagrange multiplier, it is easy to show that the optimum weight vector which solves eqn. 1 is given by

$$\hat{w} = \frac{R^{-1} s(f_0, \theta)}{s^H(f_0, \theta) R^{-1} s(f_0, \theta)} \quad (2)$$

and the optimum output power of the array is given by

$$P(\hat{w}) = \frac{1}{s^H(f_0, \theta) R^{-1} s(f_0, \theta)} \quad (3)$$

It can be shown that the output power of an optimised beamformer achieves a maximum when the steering vector coincides with that of the desired signal. A simple and yet effective approach is proposed to search for the correct steering vector.

It is clear from eqn. 3 that maximising the mean output power with respect to the steering vector is equivalent to

$$\min_s s^H R^{-1} s \quad (4)$$

In a small region $\delta\theta$ about the assumed look direction θ_0 , s can be approximated by a first order Taylor series

$$s(\theta) = s(\theta_0 + \delta\theta) \approx s(\theta_0) + \delta\theta s_1(\theta_0) \quad (5)$$

where $s_1(\theta_0) = ds(\theta)/d\theta|_{\theta=\theta_0}$. Use of eqn. 5 in eqn. 4 results in

$$\min_{\delta\theta} [s(\theta_0) + \delta\theta s_1(\theta_0)]^H R^{-1} [s(\theta_0) + \delta\theta s_1(\theta_0)] \quad (6)$$

which is a simple one-dimensional optimisation problem. It can be easily shown that the optimum $\delta\theta_0$ is given by

$$\delta\theta_0 = - \frac{\text{Re} [s^H(\theta_0) R^{-1} s_1(\theta_0)]}{s_1^H(\theta_0) R^{-1} s_1(\theta_0)} \quad (7)$$

where $\text{Re}(\cdot)$ denotes the real part of the complex quantity. The optimum steering vector can then be found by substituting eqn. 7 into eqn. 5.

If the pointing error is very small, then the steering vector found using eqn. 7 is a true steering vector in the sense that it points to a certain direction. However if the pointing error is quite large, then there is no guarantee that the vector found is a true steering vector. To overcome this, an iterative algorithm that updates the steering error $\delta\theta$ is proposed. Let θ_0 be the assumed (initial) look direction.

Algorithm:

For $i = 0, 1, 2, \dots$

$$\delta\theta_i = - \frac{\text{Re} [s^H(\theta_i) R^{-1} s_1(\theta_i)]}{s_1^H(\theta_i) R^{-1} s_1(\theta_i)}$$

If $\delta\theta_i \leq \epsilon$ then

stop

else

$$\theta_{i+1} = \theta_i + \delta\theta_i$$

repeat