achieve 18 dB CNR is about 10 dBm. For \( N = 128 \), a high power LO laser is needed to achieve the acceptable CNR because of the serious branch loss. From both Figures we conclude that a common LO system is indeed applicable to local distribution system which can serve a number of subscribers with only a high power LO laser. It is partly because of the short transmission distance in a local optical network so that the path loss is not significant, and partly because we just need a medium LO power level to reach the acceptable CNR, that if the LO power at the centre is high enough it indeed can be shared by many receivers.

In addition to the use of simple and low cost receivers, there are other advantages provided by the common LO system. Since only one LO laser is employed and located at the distribution centre, its operating environment can be circumstantially controlled and we can adopt a very narrow linewidth and highly frequency stabilised laser to implement the LO, thus minimising the laser phase noise and the IF drift to improve the system performance. There are no active optical components placed at the receivers of the subscriber premises and we can also prepare a standby LO at the centre to enhance the system reliability. We can expand the system to more subscribers by employing optical amplifiers at appropriate locations. Since the SCM and LO signals are mixed together in the transmission fibre, they can be simultaneously amplified by an optical amplifier. In addition, we do not need additional LO lasers. The common LO system can thus be easily expanded.

In conclusion we have considered a coherent SCM star distribution system using a single LO laser at the distribution centre. The system is significantly different from long haul coherent systems where the LOs are generally located at the receiving end. Taking advantage of the short transmission path and the low loss nature of single-mode optical fibres, the common LO system is found to be possible and indeed provides many advantages. We expect a simple, low cost, reliable and easily expanded coherent distribution system can be implemented by using a common LO.

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FORMULA FOR THE STEADY-STATE GAIN OF A RECURSIVE ESTIMATOR

Indexing terms: Markov fields, Recursive estimation, Image processing

A formula for the direct computation of the steady-state gain of a recursive estimator is derived. The estimator is presented as a general 3-D recursive filter for noise smoothing of 3-D wide-sense Markov fields. The 1-D and 2-D estimators are special cases of the general filter presented. The filter with its steady-state gain computed from the derived formula is very useful because of its computational efficiency.

Noise is almost always present in recorded images or image sequences, therefore noise filtering will improve the visual quality of the images and the performance of subsequent image processing tasks (e.g., coding). In general, let \( y(n) \), where \( n = (n_1, n_2, n_3) \), be the observed noisy image sequence

\[
y(n) = x(n) + v(n)
\]

(1)

where \( x(n) \) (with variance \( \sigma_x^2 \) and \( v(n) \) (with variance \( \sigma_v^2 \) are the original image sequence and white Gaussian observation noise independent of \( x(n) \), respectively. The original image sequence is modelled by a wide-sense stationary process which has the following 3-D autocorrelation function:

\[
R_{x}(l_1, l_2, l_3) = \sigma_x^2 \rho_1^{l_1} \rho_2^{l_2} \rho_3^{l_3}
\]

(2)

This separable model represents a 3-D wide-sense Markov field and results in mathematically tractable expressions. The parameters \( \rho_1 \) and \( \rho_2 \) are the vertical and horizontal spatial correlation coefficients, respectively, and \( \rho_3 \) is the temporal correlation coefficient. When dealing with image sequences the motion-compensated frames should be used in place of the original frames in eqn 1. Otherwise, unacceptable smoothing of edges result because of the significant motion of large objects. Thus, \( \rho_3 \) represents the temporal correlation coefficient in the motion trajectory. 3 The 2D case has been extensively used for image processing with success. 4 Using the general model of eqn 2, other noise smoothing filters can be derived, such as purely spatial filters \( \delta(l_1, l_2) \) and purely temporal filters \( \delta(l_1) \) and \( \delta(l_2) \), where \( \delta \) is the delta function. 5 The values of the correlation coefficients may also be adapted to both the image content (edges, textures, etc.), and to specific motion features (e.g., occlusion).

Discrete random fields with the above autocorrelation function are stationary autoregressive sources of the type

\[
x(n) = \sum_{s \in S} a(s)x(n - s) + u(n)
\]

(3)

where \( s = (s_1, s_2, s_3) \), \( S \) is the 3D causal support of the AR model which contains seven points and extends one point in each direction, and \( u(n) \) are the prediction coefficients. By fitting the model of eqn 2 into the data, we get

\[
a(s) = -(\rho_1 |s_1|)(\rho_2 |s_2|)(\rho_3 |s_3|) \quad s \in S
\]

(4)

where \( u(n) \) is a zero-mean white noise input process with variance equal to the mean-squared prediction error, \( \epsilon^2 \), which is given by

\[
\epsilon^2 = E[u(n)^2] = (1 - \rho_1^2)(1 - \rho_2^2)(1 - \rho_3^2) \sigma_x^2
\]

(5)

Based on the model of eqn 3, a recursive spatio-temporal noise smoothing filter is derived which represents an esti-
s(n) = \left[ 1 - F(n) \right] \sum_{i=0}^{\infty} a(n) s(n - i) + F(n) y(n)

(6)

Bounds for the gain \( F \) have been obtained and the stability of the filter has been studied. By conducting a performance study similar to that in Reference 1 for the 2D case, we concluded that the upper bound of the filter performance does not yield a bound on the gain. The 3D filter performs best for \( 0 < F < 1 \), where it is always stable. The steady-state gain of the estimator depends on the model coefficients \( \rho_2 \) and the signal-to-noise ratio (\( SNR = \sigma^2_1/\sigma^2_2 \)). In deriving an expression for the steady-state gain, the minimisation of the functional

\[ \mathcal{W}(x(n)) = \frac{1}{2} \| s(n) \|^2 + \frac{1}{\sigma^2_2} \| \epsilon(n) \|^2 \]

(7)

with the use of eqn. 5 gives the following expression:

\[ F = \frac{\phi SNR}{1 + \phi SNR} \]

(8)

This expression does not give good results when applied to the recursive filtering because it does not take into consideration the causality of the estimator. A different computation of the steady-state gain is derived below.

Let us assume that the estimator can be represented by a filter with input \( y(n) \) and output \( s(n) \), as was done in Reference 5. The 3D z-transform of the filter is given by

\[ H(z_1, z_2, z_3) = \frac{F}{D(z_1^{-1}, z_2^{-1}, z_3^{-1})} \]

(9)

where

\[ D(z_1^{-1}, z_2^{-1}, z_3^{-1}) = 1 - (1 - F) \sum_{s, t, u} a(s, t, u) z_1^{-s} z_2^{-t} z_3^{-u} \]

(10)

For a causal system, \( H(z_1, z_2, z_3) \) can be expanded as follows:

\[ H(z_1, z_2, z_3) = F \sum_{i_1=0}^{\infty} \sum_{i_2=0}^{\infty} \sum_{i_3=0}^{\infty} h(i_1, i_2, i_3) z_1^{-i_1} z_2^{-i_2} z_3^{-i_3} \]

(11)

where \( h(i_1, i_2, i_3) \) is the impulse response of the filter and \( h(0, 0, 0) = 1 \). Then

\[ F \left\{ \sigma^2_2 \sum_{i_1=0}^{\infty} \sum_{i_2=0}^{\infty} \sum_{i_3=0}^{\infty} h(i_1, i_2, i_3) \rho_1^{i_1} \rho_2^{i_2} \rho_3^{i_3} + \sigma^2_2 \right\} + \left\{ \sum_{i=0}^{\infty} \sum_{k=0}^{\infty} \sum_{\ell=0}^{\infty} h(i, k, \ell) \rho_1^i \rho_2^k \rho_3^\ell \right\} \]

(12)

for all \( i_1 \geq 0, i_2 \geq 0 \) and \( i_3 \geq 0 \). For \( i_1 = i_2 = i_3 = 0 \), Eqn. 12 becomes

\[ F \left\{ \sum_{i_1=0}^{\infty} \sum_{i_2=0}^{\infty} \sum_{i_3=0}^{\infty} h(i_1, i_2, i_3) \rho_1^{i_1} \rho_2^{i_2} \rho_3^{i_3} + \sigma^2_2 \right\} = \sigma^2_2 \]

(13)

The filter is causal and stable. We can therefore evaluate its z-transform at \( (z_1, z_2, z_3) = (\rho_1^{-1}, \rho_2^{-1}, \rho_3^{-1}) \), which belongs to it's region of convergence. Then, using eqns. 9 and 11 we have

\[ \sum_{i_1=0}^{\infty} \sum_{i_2=0}^{\infty} \sum_{i_3=0}^{\infty} h(i_1, i_2, i_3) \rho_1^{i_1} \rho_2^{i_2} \rho_3^{i_3} = \frac{1}{D(\rho_1, \rho_2, \rho_3)} \]

(14)

where

\[ D(\rho_1, \rho_2, \rho_3) = \phi + F(1 - \phi) \]

(15)

Finally, by substituting eqn. 14 into eqn. 13 we get

\[ F \left[ \frac{1}{\phi + F(1 - \phi)} + \frac{1}{SNR} \right] = 1 \]

(16)

The steady state gain is computed from the following quadratic equation:

\[ (1 - \phi)F^2 + \phi(SNR + 1)F - \phi SNR = 0 \]

(17)

and the positive root is chosen for \( F \), that is

\[ F = \frac{\sqrt{(SNR + 1)^2 \phi^2 + 4 \phi (1 - \phi) SNR - \phi (SNR + 1)}}{2(1 - \phi)} \]

(18)

In Fig. 1, the gain \( F \) computed from eqns. 8 and 18 is plotted as a function of the SNR (dB) for \( \rho_1 = \rho_2 = \rho_3 = \rho \), where \( \rho \) is equal to 0.8 and 0.9. The values of \( F \) evaluated from eqn. 18 are greater than those obtained from eqn. 8. The observation is therefore weighted more in filtering since the estimator uses causal prediction. Some results of the application of the 2D estimator to image noise smoothing are presented. The image of a woman of size 256 \times 256 \times 8 was used which was distorted by adding white Gaussian noise. Two cases were considered where \( \sigma^2_1 = 273.52 \) (SNR = 10 dB) and \( \sigma^2_1 = 864.98 \) (SNR = 5 dB). The coefficients \( \rho_1 = 0.92 \) and \( \rho_2 = 0.91 \) were computed from the original image and \( \rho_3 = 0.95 \). The noisy images were restored for steady-state values of the gain ranging from 0 to 1. The figure of merit is the mean-squared error (MSE) improvement (dB) defined by

\[ I_{MSE} = 10 \log_{10} \left( \frac{\sum_{n=0}^{\infty} s(n)^2}{\sum_{n=0}^{\infty} e(n)^2} \right) \]

(19)

where \( s(n) \) is the average of four estimates obtained by filtering the image in different directions of recursion. In Fig. 2, \( I_{MSE} \) is shown as a function of \( F \) for both cases. It is clear that the maximum \( I_{MSE} \) is obtained when \( F \) is evaluated according
and for $SNR = 5\, dB$ we get $F = 0.0694$ which results in $I_{MSE} = 3.2371\, dB$. The distorted image with $SNR = 5\, dB$ is shown in Fig. 3. The restored images with $F$ evaluated from eqns. 8 and 18 are shown in Figs 4 and 5, respectively. It is clear that Fig. 5 represents a better result in terms of both visual quality and $MSE$ improvement. The proposed formula (eqn. 18) gives the steady-state gain of the recursive estimator of eqn. 6 which results in the minimum mean-squared filtering error.

Fig. 5  Image restored using eqn. 18

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3.2GHz, 0.2μm GATE CMOS 1/8 DYNAMIC FREQUENCY DIVIDER

Introduction: CMOS devices scaled down to deep submicrometre sizes have been reported for the purpose of developing ULSIs which integrate 100 million transistors. 1-2 These CMOS devices offer improved high-speed circuit performance with a very low power consumption as well as high packing density. MultigigaHertz CMOS dual-modulus prescalars ICs integrated in bulk silicon 3 using 0.4 to 0.5μm gate CMOS technology have recently been reported.

High-speed circuits such as prescalars and dividers have been widely applied to ICs such as multiplexers and demultiplexers for high-speed transmission systems, and as frequency synthesizers for mobile telephone systems. These systems consist of both low and high-frequency components which are usually fabricated with CMOS and GaAs or Si bipolar technologies, respectively. If the high-frequency parts could be made using CMOS technology instead of GaAs or Si bipolar technologies, the process compatibility with the low-frequency parts would permit the system integration on a common silicon substrate. As a result, total power dissipation, cost and development time would be significantly reduced for system development.

This letter describes the performance of a 1/8 dynamic frequency divider fabricated with subquarter-micrometre gate CMOS technology 1-4 and designed with a 0.6μm design rule except for the gate length. The divider IC has the highest speed performance ever reported for a digital CMOS circuit integrated in bulk at a low supply voltage below 2.0 V.

Deep submicrometre CMOS process technology and divider circuits: The deep submicrometre CMOS process technology was used for divider fabrication. The detailed fabrication steps have been previously reported. 1-4 CMOS devices were designed with a 0.6μm rule except for the gate lengths. CMOS device parameters and measured $I/V$ characteristics are listed in Table 1. This process yields an inverter chain propagation-delay time of 34 ps at a supply voltage of 2.0 V.

Fig. 1a shows the block diagram of the 1/8 frequency divider IC, which contains three dynamic binary frequency dividers, an internal buffer, an an output buffer. They consist of fully complementary CMOS circuits. Two types of 1/8