Digitizing Multichannel Video Signals for Lightwave Transmission

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***Abstract*—This paper presents a new approach to the design of lightwave multichannel CATV transmission systems using broadband digital coding to convert an all-analog multiplexed video signal to a string of pulses suitable for digital transmission. Four broadband coding methods have been discussed and compared, which are: uniform and optimal companding PCM, as well as uniform and adaptive DM. The principle advantages of this approach include the benefits obtainable from digital transmission and also direct compatibility with current CATV networks. CNR limits have been investigated for all of these methods.**

I. Introduction

OPTICAL fiber analog multichannel video transmission has so far been dominated by two modulation schemes, i.e., AM subcarrier multiplexing (AM–SCM) and FM–SCM [1]–[3]. The AM–SCM transmission systems, in which video signals are transmitted in vestigial-sideband (VSB) format, are especially convenient for connection with current CATV networks. However, like other AM systems, they are vulnerable to nonlinearity and noise, and thus generally have a low power budget and limited channel capacity.

Considerable efforts have been made in order to improve the system performance, which include the improvement of laser linearity [4], the reduction of the laser relative intensity noise (RIN) [5], the use of optical amplifiers, and the overmodulation of the laser to increase optical modulation index (OMI) [6]–[8]. An alternative approach employing a more robust FM–SCM scheme seem to be a feasible choice, because of the availability of the trade-off between bandwidth and signal-to-noise-ratio (SNR) for FM modulation. Obviously, most of the methods mentioned above for improving AM–SCM systems can also be applied to FM–SCM systems. However, the fundamental disadvantage of the inability to be directly compatible with AM–VSB equipment prevents the FM–SCM systems in large scale deployment in current CATV networks. In fact, present FM–SCM systems with expensive conversion from FM to AM and vice versa for each channel operate primarily in CATV supertrunks, where high-quality transmission is of paramount concern for further distribution.

A novel solution is presented in this paper, in which the whole analog multichannel AM–VSB video is digitized and then transmitted over a high-speed digital optical fiber link. The original analog signal can be recovered through an opposite digital-to-analog conversion in the receiver. This approach could provides a high degree of immunity to transmission distortion, interference, and noise because of the nature of digital transmission. Therefore a high optical power margin can be obtained and a regenerative repeater used for very long distance transmission and distribution. In addition, the analog-to-digital conversion is performed only once at the transmitter end, and no conversion of channel by channel, as in the case of FM–SCM systems, is required. This allows systems to be able to connect directly with the current CATV networks, since the signals at the interfaces remain in the same VSB–AM format.

Multichannel analog video signals typically occupy a bandwidth of several hundreds of MHz. Hence, the transmission rate for the proposed system would be in the order of multi-Gb/s or even more. For example, if a 600 Msps, 8-bit PCM system is chosen, the transmission rate will be at least 4.8 Gb/s. Furthermore, an extra rate may be needed for line coding, in order to provide efficient timing and synchronization as well as possible error detection and correction. With a delta modulation (DM) system, the transmission rate will be even higher for a full system performance. Therefore, advanced microwave circuits and subsystems will be indispensable for these systems. The operating rate of optical fiber digital transmission systems has increased dramatically over the past few years. A transmission system operating at 10 Gb/s has been demonstrated to be feasible. Research and early development work focuses on 20- and 40-Gb/s systems [9]. Also, the same period has seen a rapid progress in the conversion speed of analog-to-digital converters (ADC's). An 8-bit ADC with a sampling rate of 500 Msps and compatible digital-to-analog converters (DAC's) are already commercially available. Research in these two aspects is still in progress and will be able to provide a necessary basis for the proposed scheme.

II. System CNR Analysis

The block diagram of the system under consideration is shown in Fig. 1(a), where the input multichannel CATV signals are fed to a gain-adjustable wideband amplifier for a suitable amplitude to be applied to the following ADC. Basically there are two ways to digitize an analog signal. One is pulse-code modulation (PCM), and other is DM. PCM includes uniform and companding PCM, while DM comprises uniform and adaptive DM. PCM and DM techniques have found wide applications in various audio and video systems. ADC and DAC in Fig. 1(a) represent the relevant coding parts for both PCM and DM. Generally, the PCM scheme can offer a higher SNR than a DM system does for a limited available transmission rate, but involves more complex

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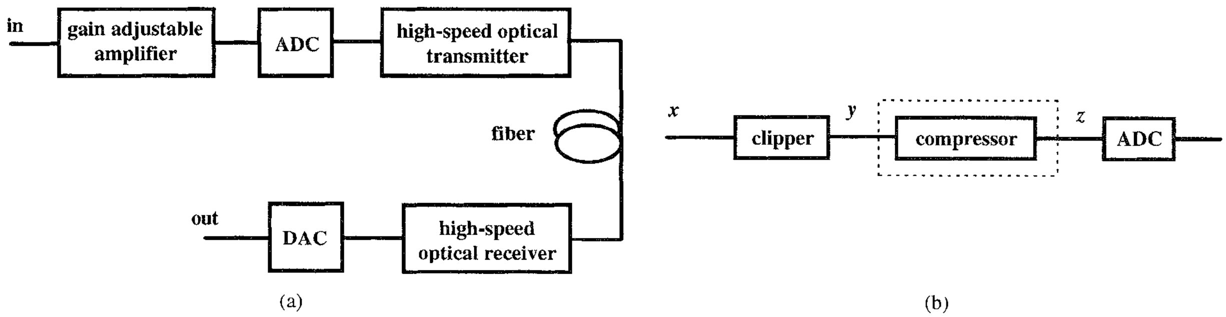


Fig 1. Block diagram of system: (a) configuration and (b) analysis model of an ADC at the transmitter for PCM schemes.

hardware. Hence the applications of DM are primarily limited to those cases where ease of implementation takes precedence over bandwidth considerations.

In this section, the limits of the carrier-to-noise-ratio (CNR) for both PCM (uniform and optimal companding) [10] and DM (uniform and adaptive) schemes will be investigated. Only fundamental noise, i.e., quantization noise and clipping noise for PCM systems, as well as overload and granular noise for DM systems, will be considered in detail. Other noise sources, such as conversion nonlinearity of ADC's and DAC's and noise in the transmission section also exist and cannot always be ignored, but they are not regarded as fundamental noise sources, and, therefore, will not be counted in the CNR calculation.

A. CNR of Uniform PCM System

The analog-to-digital conversion process at the transmitter may be modeled as shown in Fig. 1(b), where the clipper limits the amplitude of the multichannel signal within ± 1 V, the normalized input range of an AD converter. Although not existing in actual systems, it is introduced for simulation of the clipping action by the finite input range of ADC's. The transfer function of the clipper is

$$y = \begin{cases} x & |x| < 1 \\ \pm 1 & |x| \geq 1 \end{cases}. \tag{1}$$

Thus, nonlinear distortion (NLD) occurs at the clipper if the total instantaneous input voltage exceeds the range, which forms one of the two major noise mechanisms considered here.

Assuming that the normalized input signal, with respect to the half input range of ADC's, has a same amplitude, U, in each channel, the multiplexed input signal voltage to the ADC for an M-channel system may be written as

$$U_{in}(t) = \sum_{i=1}^{M} U \cos(2\pi f_n t + \theta_n) \tag{2}$$

where f_n is the center frequency for the nth channel, and θ_n is a random variable uniformly distributed over (0, 2π). The probability distribution for the amplitude of $U_{in}(t)$ can be described by a Gaussian process with good accuracy for $M \geq 10$ [8]. Because each channel is independent, the variance of $U_{in}(t)$ becomes $\sigma^2 = MU^2/2$, and apparently the mean is zero. Under the Gaussian assumption, the autocorrelation function of the clipper output y consists of a signal term, ${}_sR_o(\tau)$, and a noise term, ${}_nR_o(\tau)$, which are found to be [8]

$${}_sR_o(\tau) = \frac{U^2}{2} \sum_{i=1}^{M} h_1^2 \cos(2\pi f_n \tau) \tag{3}$$

$${}_nR_o(\tau) = \sum_{k=3,\, odd}^{\infty} \frac{h_k^2}{k!} \operatorname{sinc}^k(\pi M B \tau)\, \sigma^{2k} \cos^k(2\pi f_0 \tau) \tag{4}$$

where f_0 is the center frequency of the signal band. h_k representing the nonlinear coefficients of the clipper is determined by the transfer function as well as σ and given by [8]

$$h_k = \begin{cases} \operatorname{erf}\left(\dfrac{1}{\sqrt{2}\sigma}\right) & k = 1 \\ -\dfrac{2\exp\left(-\dfrac{1}{2\sigma^2}\right)}{\sqrt{\pi}\,(\sigma\sqrt{2})^{k-1}} H_{k-2}\left(\dfrac{1}{\sqrt{2}\sigma}\right) & k = 3, 5, 7, \cdots \end{cases} \tag{5}$$

where $H_n(k)$ is the Hermitian polynomial. However, it should be noted that the h_k expression here is slightly different from that given by [8].

We extend the analysis [8] which concerns the carrier-to-NLD-ratio (CNLD) in the center channel to all other channels by adding the $\sum$ term in (3), so that CNLD calculation for each channel can be made. As later will be shown, the worst CNLD channel will no longer be fixed at the center channel, since the NLD spectrum may not be symmetry about the center frequency with the worst NLD in the center channel.

Equation (3) shows that the power of y signal in each channel is equal to $0.5U^2h_1^2$. Since the NLD spectrum, $W(f)$, which is the Fourier transform of ${}_nR_o(\tau)$, scarcely change with f in each single channel, $2BW(f_n)$ would be a good approximation of the noise power in the nth channel, where B denotes the channel bandwidth. Thus, the CNLD at the nth channel is given by

$$\text{CNLD} = \frac{h_1^2 \sigma^2}{2BM\text{W}(f_n)} \tag{6}$$

where

$$W(f_n) = \frac{2}{\pi M B} \sum_{k=3,\, odd}^{\infty} \frac{h_k^2 \sigma^{2k}}{k!} \int_0^{\infty} \operatorname{sinc}^k(u) \cdot \cos^k\left(\frac{2 f_0}{MB} u\right) \cos\left(\frac{2 f_n}{MB} u\right) du. \tag{7}$$

The other main noise considered here is quantization noise (QN), which is inherent in the PCM system as a result of the limited number of quantum levels. Given the assumption that the quantization error probability is equal for any signal amplitude, it can be shown that the spectrum density of QN is flat and equals $2^{-2q}/3f_s$, where f_s is the sampling rate and q the bit resolution [11]. Therefore the received carrier-to-QN-ratio (CQN), assuming errorless transmission of the digits, is obtained as

$$\mathrm{CQN} = \frac{3f_s\sigma^2 h_1^2}{BM2^{-2q+1}}. \tag{8}$$

Thus the CNR can be found by

$$\mathrm{CNR} = \left(\frac{1}{\mathrm{CNLD}} + \frac{1}{\mathrm{CQN}}\right)^{-1} \tag{9}$$

Since CNLD and CQN, hence CNR, are functions of σ which is an adjustable parameter in practical systems, the performance may be optimized by selecting an appropriate σ.

B. CNR of Optimal Companding PCM System

The companding, i.e., the joint use of signal compression and expansion, has been shown to be an effective technique to reduce QN, where the probability density function (PDF) of the input signal is not uniformly distributed. Companding circuits, composed of a compressor at the transmitter and a complimentary expander at the receiver, as well as a uniform ADC and DAC, together perform a nonuniform quantization, as shown in Fig. 1(a) (ADC part). Since the PDF of the multiplexed signal is approximately a Guassian one and the signal mostly often appear around zero, a CNR improvement could be expected by using an appropriate companding function.

Let $f_Y(y)$ be the PDF of signal y, the optimal compressor transfer function has the form [12]

$$Z(y) = A\left[\frac{2\int_{-\infty}^{y}[f_Y(\eta)]^{1/3}\,d\eta}{\int_{-\infty}^{\infty}[f_Y(\eta)]^{1/3}\,d\eta} - 1\right] \tag{10}$$

where A is the maximum value of $Z(y)$. Since signal clipping is rare, $f_Y(y)$ is essentially same as the PDF of signal x, $f_X(x)$, for $|y| \le 1$ and can expressed as a Gaussian function

$$f_Y(y) = \frac{1}{\sqrt{2\pi\sigma^2}}\exp\left(-\frac{y^2}{2\sigma^2}\right) \quad |y| \le 1. \tag{11}$$

Substituting it in (10), and normalizing the optimal function at the condition of $|z(\pm 1)| = \pm 1$

$$Z(y) = \mathrm{sgn}\,(y)\,\frac{\mathrm{erf}\left(\frac{|y|}{\sqrt{6}\sigma}\right)}{\mathrm{erf}\left(\frac{1}{\sqrt{6}\sigma}\right)}. \tag{12}$$

With companding, the CQN could be improved by a factor k [13], where k is

$$\begin{aligned} k &= 2\int_0^1 \frac{f_Y(y)}{[Z'(y)]^2}\,dy \\ &= 3\pi\sigma^2\,\mathrm{erf}^2\left(\frac{1}{\sqrt{6}\sigma}\right)\int_0^1 f_Y(y)\exp\left(-\frac{y^2}{3\sigma^2}\right)dy. \end{aligned} \tag{13}$$

Assuming that the power spectral density of QN remains flat, the output CQN for each channel is obtained by dividing (8) by k

$$\mathrm{CQN} = \frac{3f_s\sigma^2 h_1^2}{BMk2^{-2q+1}}. \tag{14}$$

The CNLD in this case is just the same as in uniform PCM. Similarly, the CNR can also be obtained by (9).

C. CNR of Uniform DM System

Uniform DM is a simple type of predictive quantization. There are two types of quantizing noise, i.e., granular and overload. Granular noise is similar to the quantization noise of PCM and occurs because the samples can assume only discrete values, which in DM are multiples of the height h of the output digital pulses. Overload noise is a result of the maximum slope a DM system may produce being limited to hf_s, where f_s is the sampling frequency.

Experimental studies confirm that the power spectrum of granular noise is essentially flat over $|f| \le f_s$ [13]. On the other hand, for a given signal rms, a reasonably large value of hf_s results in negligible overload noise. On this condition, the maximum CNR for each channel is given by [13]

$$\mathrm{CNR} = \frac{6}{\pi^2}\left(\frac{f_{\max}}{\Delta f_{rms}}\right)^2 \frac{b^3 f_{\max}}{MB\,\ln^2(2b)} \tag{15}$$

where $f_{\max}$ and MB are the highest frequency and the bandwidth of the input signal, respectively. $b = f_s/2f_{\max}$ and Δf_{rms} is signal's rms bandwidth defined as

$$\Delta f_{rms} = \frac{1}{\sigma}\left[\int_{-\infty}^{\infty} f^2 G(f)\,df\right]^{1/2} \tag{16}$$

$G(f)$ is the power spectrum of the input signal x. As the signal in each channel has the same intensity, and is treated as a random noise, it may be assumed that the signal spectrum is uniformly distributed over the entire bandwidth. This results in

$$G(f) = \frac{\sigma^2}{2BM}. \tag{17}$$

Therefore

$$\Delta f_{rms} = \frac{1}{\sqrt{3BM}}\left(f_{\max}^3 - f_{\min}^3\right)^{1/2} \tag{18}$$

where $f_{\min}$ is the lowest frequency of the input signal band.

D. CNR of Adaptive DM System

Adaptive DM involves additional hardware designed to provide a variable step size, thereby reducing slope-overload effects without increasing the granular noise. In adaptive DM, a typical 8–14 dB improvement in CNR may be obtained [13]. Therefore, the CNR of this system may be obtained simply by adding 8–14 dB to (15).

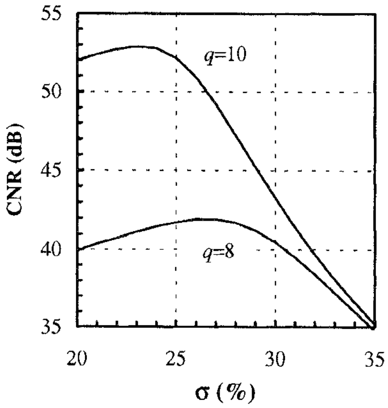


Fig. 2. Output CNR against the normalized signal rms σ for the 8-bit and 10-bit uniform PCM systems.

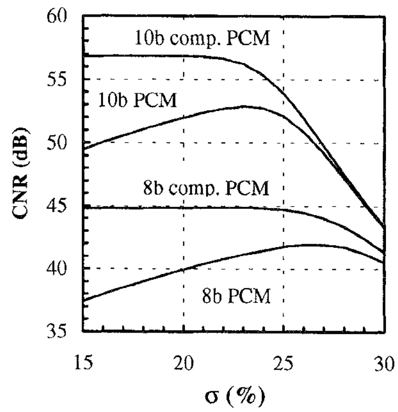


Fig. 3 Comparison of the optimal companding PCM with the uniform PCM.

III. Results and Discussions

In order to determine what the performance of the proposed scheme can attain, some specific cases are studied in the following, where the input signal is a 40-channel multiplexed NTSC video signal ranging from 55.25 MHz–289.25 MHz, with $B = 6$ MHz intervals and no guard band. For PCM systems, the sampling rate, f_s, is set to 600 MHz and the bit resolution q is chosen to be 8 and 10, respectively.

Fig. 2 shows that the peak CNR's for an 8-bit and 10-bit uniform systems are 41.9 dB, at $\sigma = 0.27$, and 52.9 dB at $\sigma = 0.23$, respectively. The optimal σ will become smaller if q increases. This is understandable, because QN is lower for bigger q and, to achieve the maximum CNR, σ must be shifted downwards to avoid the constraint by CNLD. Notice that the CNR's in Fig. 2 represent the worst values among all channels and, therefore, can be regarded as the system performance limits.

Fig. 3 shows the CNR's for the optimal companding PCM systems, where the uniform cases are also plotted for comparison. By employing optimal companding, peak CNR's improve approximately 3 dB and 4 dB, for $q = 8$ and 10, allowing them to be 44.8 dB and 56.8 dB. An interesting phenomenon for optimal companding systems is that the CNR could be kept almost constant over a fairly wide range of σ, when σ is less than a certain value corresponding to the peak CNR. This is due to a greater improvement of CNR in the lower σ region by nonuniform quantization. However, the optimal companding function changes with σ, as indicated in (12). In order to simplify circuit design, it is helpful to fix the parameters of the companding circuit and then only change the intensity of the input signal by adjusting the gain of the input amplifier.

As stated before, the worst CNR channel differs with σ. Fig. 4 shows the channel location number n of the worst CNR against σ for either of the PCM systems. Since the noise spectrum of QN is flat over the frequency band concerned, the worst channel is solely decided by the CNLD. When σ is low, the worst channel will reside at the higher end of the band. Increasing σ pushes the worst channel firstly toward the lower end and then to the central region of the signal band. It is also the CNLD that determines the profile of CNR distribution over all the input bandwidth. Fig. 5 shows the CNR plotted against the channel number n at $\sigma = 0.23$ for a 10-bit uniform PCM system. This kind of profile is different from that given by [8], where the worst channel permanently stays at the center channel.

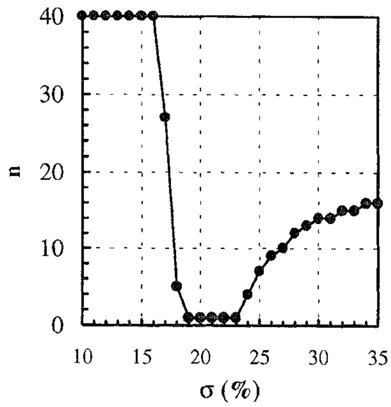


Fig. 4. Channel location number n of the worst CNR against the normalized signal rms σ for the PCM systems.

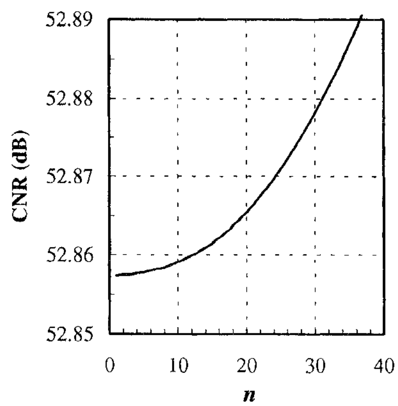


Fig. 5. Distribution of CNR against channel location number n for the 10-bit uniform PCM system when the normalized signal rms σ =0.23.

For uniform DM system, the output optimal CNR for each channel verses b is displayed in Fig. 6, where the CNR is shown to grow with b, that is, with increasing sampling rate f_s, since $b = f_s/(2f_{max})$. From (15), it is clear that the only adjustable parameter concerned here is b. When $b = 24$, the

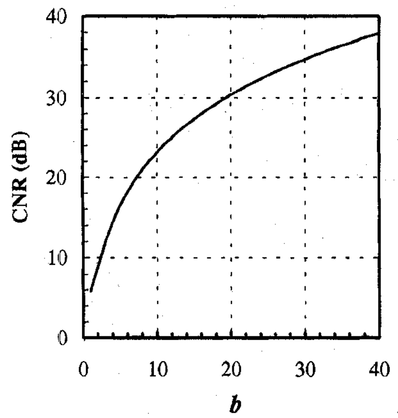


Fig. 6. CNR against $b[= f_s/2f_{\max}]$ for the uniform DM system.

sampling rate $f_s = 14.4$ GHz, which is more than twice the signal rate of the 10-bit PCM system, the CNR being only 32.3 dB, far below that obtainable by PCM systems for the same signal rate. However, the circuits of DM systems should be simpler. On the other hand, by employing adaptive DM, the CNR cited above for uniform DM could be raised to 40.3–46.3 dB. Furthermore, both uniform and adaptive DM systems performances can be improved with increasing signal rate.

IV. Conclusion

Broadband digital coding of an input-multiplexed, VSB–AM video signal has been proposed for lightwave transmission. Four schemes, i.e., uniform and optimal companding PCM as well as uniform and adaptive DM, have been analyzed and compared. The advantages of the proposed scheme over conventional methods include all the benefits expected from digital transmission, such as a high degree of immunity to transmission distortion, interference and noise, high power budget and possible errorless relay, as well as direct compatibility with the current VSB–AM CATV network. The proposed scheme is based on recent advances in the two key technologies, i.e., high-speed ADC's and DAC's, including related high speed microwave circuits, as well as high-speed lightwave transmission.

In the analysis of system CNR limits, fundamental noise sources have been defined as clipping and quantization noise for PCM systems, as well as overload and granular noise for DM systems. For a given input of a 40-channel multiplexed video signal, the optimal CNR's of the uniform PCM system are 41.9 dB and 52.9 dB, at a 600 MHz sampling rate, for 8-bit and 10-bit resolution respectively. For the optimal companding PCM system, the CNR's are 44.8 dB and 56.8 dB. In comparison, at 14.4 GHz sampling rate, the uniform DM system can only offer a 32.3 dB CNR, while the adaptive DM approach can produce a 40.3–46.3 dB CNR. Both DM systems need far higher sampling rates to achieve a comparable performance of PCM systems. However, their simple circuit hardware may have some attraction in the design of practical systems.

Finally, only quantization noise has been considered for ADC's and DAC's. However, for available commercial products, the conversion nonlinearity at high frequency, i.e., the limited dynamic range, may seriously degrade the system performance. This is something which needs to be resolved in the future.

References

[1] T. E. Darcie, "Subcarrier multiplexing for lightwave networks and video distribution systems," *IEEE J. Select. Areas Commun.*, vol. 8, no. 7, pp. 1240–1248, 1990.

[2] J. A. Chiddix, H. Laor, D. M. Pangrac, L. D. Williamson, and R. W. Wolfe, "AM video on fiber in CATV systems: Need and implementation," *IEEE J. Select. Areas Commun.*, vol. 8, no. 12, pp. 776–777, 1990.

[3] R. Olshansky, V. Lanzisera, and P. Hill, "Subcarrier multiplexed lightwave systems," *J. Lightwave Technol.*, vol. 7, pp. 1329–1342, 1989.

[4] G. Morthier, "Influence of the carrier density dependence of the absorption on the harmonic distortion in semiconductor lasers," *J. Lightwave Technol.*, vol. 11, pp. 16–19, 1993.

[5] K. Sato, "Intensity noise of semiconductor laser diodes in fiber optic analogue video transmission," *IEEE J. Quantum Electron.*, vol. QE-19, pp. 1380–1391, 1983.

[6] I. M. I. Habbab and A. A. M. Saleh, "Fundamental limitations in EDFA-based subcarrier-multiplexed AM–VSB CATV systems," *J. Lightwave Technol.*, vol. 11, pp. 42–48, 1993.

[7] A. A. M. Saleh, "Fundamental limit on number of channels in subcarrier-multiplexed lightwave CATV systems," *Electron. Lett.*, vol. 25, no. 12, pp. 776–777, 1989.

[8] K. Alameh and R. A. Minasian, "Optimum optical modulation index of laser transmitters in SCM systems," *Electron. Lett.*, vol. 26, no. 16, pp. 1273–1275, 1990.

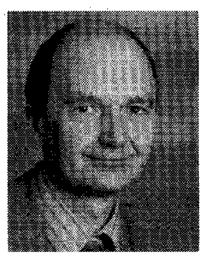
[9] H. Ichino, M. Togashi, M. Ohhata, Y. Imai, N. Ishihara, and E. Sano, "Over-10-Gb/s IC's for future lightwave communications," *J. Lightwave Technol.*, vol. 12, pp. 308–319, 1994.

[10] Q. Pan and R. J. Green, "AM–SCM–PCM lightwave CATV transmission systems," *Electron. Lett.*, vol. 30, no. 14, pp. 1155–1156, 1994.

[11] K. C. Pohlmann, *Advanced Digital Audio.* SAMS, 1991.

[12] J. G. Proakis and M. Salehi, *Communication Systems Engineering.* Englewood Cliffs, NJ: Prentice-Hall, 1994.

[13] A. B. Carlson, *Communication Systems: An Introduction to Signal and Noise in Electrical Communications,* 3rd ed. New York: McGraw-Hill, 1986.



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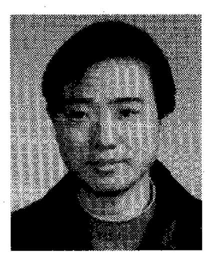

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