Hybrid Multichannel Analog/Digital CATV Transmission via Fiber Optic Link: Performance Limits and Trade-Offs

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***Abstract* - It is highly likely that a hybrid analog/digital system will be the format for future Cable transmission systems. The motivation, for using hybrid systems, is to maximize the tuning compatibility with existing consumer electronic devices, and such a mixed system may provide excellent digital video pictures, while still maintaining the quality of conventional analog (AM) signals. However, our research results have shown that in such mixed systems the digital signals may suffer a significant performance degradation due to the nonlinear, especially the clipping, operation of the laser transmitter when mixed signals are transmitted by a fiber optic link. Careful considerations should be given to available trade-offs in such hybrid system designs and realizations.**

1. INTRODUCTION

Research Target

The intent of this research was to develop cost-effective hybrid, analog/digital, transmission systems for new Cable, fiber-to-the-node or feeder architectures, (as shown in Fig. 1).

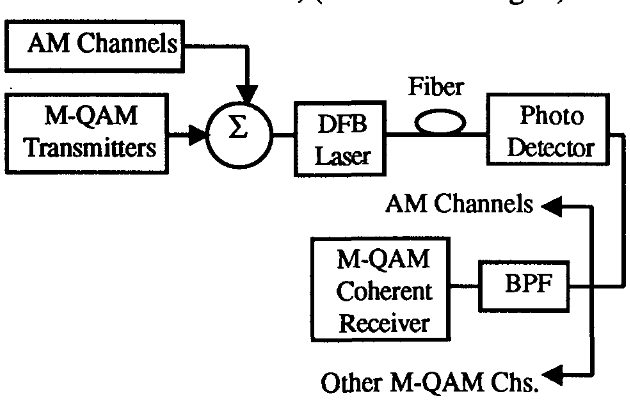


Fig. 1 Basic Configuration of the Hybrid SCM AM/QAM Fiber Optic System

General design targets are:

- 1.0 GHz., hybrid fiber optic/coax. transport to each Service Area, (~ 150 channels, 70 analog and 80 digital).
- One laser /Service Area, used to deliver all analog and digital downstream channels.
- Small Service Areas, (~ 500 homes ea.).
- Independent transmission of digital channels to each Service Area.

These targets would be required to implement cost-effective systems. All targets appear realizable, considering the recent developments in Cable technology:

- availability of high performance lasers,
- deployment of 1.0 GHz. coaxial transport,
- demonstration of compression of several television programs within 6 MHz. of bandwidth,
- development of digital video servers, and the
- possibility for telephony and interactive services by major Cable operators.

Problems for the Research.

Although, laser diodes have been shown to be able to transmit 80 analog channels with satisfactory Cable performance specifications, adding 80, or so, digital channels to the AM channels, using a single laser diode and fiber, would require further investigation. Theoretically the digital signals could be carried at power levels significantly lower than the analog AM signals.

Two initial problems were:

- A better model was needed for laser diodes used for analog multi-carrier Cable transport.
- A clear understanding of how laser diode induced multi-carrier nonlinear impairments would affect AM/digital transmission was needed.

Few experiments have been done and not many articles have been published for multi-carrier transmission of standard 6 MHz. AM Cable channels by lasers and fiber optic cables. Several models for the transmission of several analog channels over a single laser diode appeared [1]-[3]. These early models made several simplifying assumptions and the resulting models did not appear to agree closely with experimental results [6], [7]. This variation in experimental results and the theoretical models suggested that a more representative model should be developed.

Distortion - What is the Problem ?

The initial problem inhibiting the transmission of Cable quality signals by laser diodes was associated with the linearity of the laser diode. The major problems seemed to be low power and a lack of linearity in the operating region of the diode.

The classic distortion products produced by amplifier nonlinearities in multi-carrier analog Cable systems are composite second order distortion, (CSO), and third order distortion or composite triple beat, (CTB). These types of distortion are also produced by laser diodes. But, there were two new aspects to optical transmission by laser diodes, which were not prevalent in existing coaxial systems with standard trunk amplifier cascades:

- laser diodes are single ended and clip multi-carrier signals,
- the nature of these clipped signals is impulsive in nature and its effect on hybrid AM and digital transmission has not been throroughly investigated.

Limiting Factors

It has been shown that for laser diodes the limiting factor for multi-carrier transmission of VSB-AM signals is no longer the nonlinear operating region of the diode; it is the clipping effect [1]-[5]. This effect is caused by the clipping of the combined signals at the laser diode threshold when the modulation index is increased.

Further it has been shown that laser diode clipping creates a probabilistic impulse noise whose average value is small but, whose peak value is quite large [8]-[12]. These large peak values of impulse noise do not significantly affect AM transmission because of their low probability of occurrence, making their average value small. These large peaks, although they occur infrequently, do greatly affect the digital transmission portion of the hybrid system.

What Has Been Done

Several trade-offs are required to be able to do cost effective hybrid system designs that use the full capabilities of modern Cable components and architectures. In this paper we develop the basic analytic tools for analyzing hybrid systems and then show an iterative approach that can be used to design cost effective hybrid transmission systems.

In Section 2 we show a laser diode model that includes the contributions of clipping distortion, which are modeled as a Poisson impulse train. This model has been shown to agree closely with experiments done at other laboratories [6][7].

In Section 3 the effect of this clipping noise on the BER performance of M-QAM is modeled. Clipping-induced distortion is shown to be the most significant noise affecting BER, under specific conditions, and a noise floor is shown which limits the performance of digital transmission in hybrid fiber optic systems. Three cases are presented which show the BER performance of 16-QAM and 64-QAM in relationship to the impulsive noise generated by laser diode clipping.

In Section 4 we show several performance limits and trade-offs using 64-QAM and 256-QAM as examples. A design example is given to show how trade-offs can be made using typical components and system parameters.

In Section 5 techniques are discussed for reducing the effects of clipping distortions, with channel coding. Examples of common, Reed-Solomon codes are selected for one and two symbol correction. The effects of these codes on clipping distortion noise is then shown, respectively, for 64-QAM and 256-QAM signals combined with AM signals in hybrid systems.

2. MODEL AND CHARACTERISTICS OF CLIPPING DISTORTION AND NOISE

Distortion (Intermodulation) Products

It is known that the clipping-induced distortion in a multi-carrier fiber system generates many higher order in-band intermodulation products in addition to the classical CSO and CTB products [4][5]. Several models have been developed to calculate the in-band distortion power. Earlier models [1]-[3] have, however, made simplifying assumptions and appeared not to agree with experimental results [6][7].

It is this discrepancy between the earlier models and experiments that have led our effort to develop a model more accurately accounting for the clipping distortion. The model, as described below, can predict the clipping distortion power at any channel for all orders of distortion, including the CSO and CTB [4].

Briefly stated, for an N-channel signal input to an ideal laser, the model predicts that the total carrier to distortion power ratio (C/NLD) for a channel is given by [4]:

$$(C/NLD)^{-1} = \frac{1}{2\pi h_1^2} \sum_{n=2}^{\infty} \frac{K_n}{N^{n-1} 2^n} \cdot e^{-2/m^2 N} G_{n-2}^2(m\sqrt{N}) \quad (1)$$

where K_n is the product count of the n-th NLD at that channel, m is optical modulation index (OMI) per channel, $h_1 = [1 + \mathrm{erf}(1/m\sqrt{N})]/2$, erf(·) being the error function, and

$$G_{n-2}(m\sqrt{N}) = \sum_{k=0}^{[(n-2)/2]} (-1)^{n-k} \frac{(n-2)!}{(2k)!(n-2-2k)!} \cdot \left(\frac{\sqrt{2}}{m\sqrt{N}}\right)^{n-2-2k} C_k, \quad n \geq 2 \quad (2)$$

with $C_k = 1$ if $k = 0$ and $C_k = 1 \cdot 3 \cdot 5 \cdots (2k-1)$ if $k \geq 1$. The CSO and CTB are then obtained from (1) by letting n = 2 and n = 3, respectively, i.e.,

$$CSO = \frac{K_2}{8\pi N h_1^2} e^{-2/m^2 N} \quad (3a)$$

$$CTB = \frac{K_3}{8\pi N^2 h_1^2} e^{-2/m^2 N} \quad (3b)$$

To give an example, in Fig. 2 we show the CSO obtained from (3a) for a 42-channel system, and compared with experimental results [6][7]. The figure also includes the simulation results based on the earlier models [1][2].

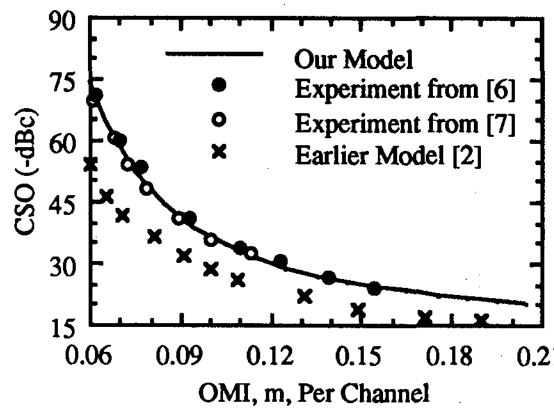


Fig. 2 CSO vs. OMI for 42-Channel Case

The figure shows that the CSO given by our model closely agrees with the experimental data, and is 5-15 dB less than that estimated by the earlier models.

Clipping Distortion as "White" Noise

From the frequency viewpoint, the clipping-induced distortion is wideband, and may be considered as "white" noise to the system. This can be verified by the power spectral density (PSD) of the clipping noise which is given by [4][8]:

$$P_W(f) = \sum_{k=2}^{\infty} \frac{h_k^2}{k!} \int_{-\infty}^{\infty} \left[\frac{R_{in}(\tau)}{R_{in}(0)}\right]^k e^{-j2\pi f\tau} d\tau \quad (4)$$

where $R_{in}(\tau)$ is the autocorrelation function of the laser input signal and h_k is given by [4]

$$h_k^2 = \frac{R_{in}(0)}{2\pi} e^{-2/m^2 N} G_{n-2}^2(m\sqrt{N}),\ k \geq 2 \quad (5)$$

Fig. 3 shows a plot of $P_w(f)$ when the input signal consists of 70 AM and 80 digital (64-QAM) signals. Fig. 3 shows that the spectrum of the clipping noise is roughly flat below 1 GHz - the band of interest for CATV. The in-band noise power at a channel is then given by $NLD_{in} = P_w(f_c) \cdot B$, f_c being the carrier frequency of that channel and B is the bandwidth. Note that this NLD should be the same as that given by (1); that is, that we have two alternative ways to calculate the NLD induced by clipping.

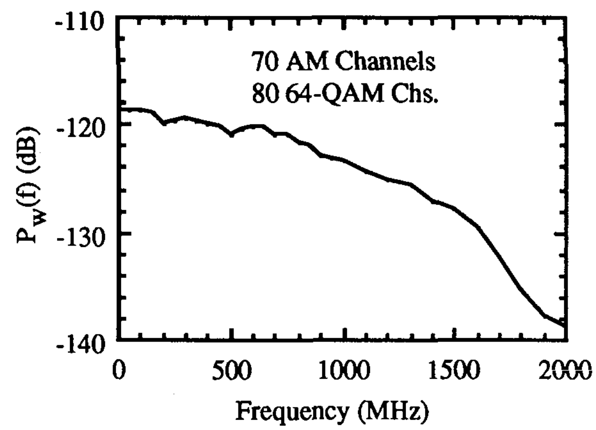


Fig. 3 PSD of the Clipping Noise

<u>Clipping Distortion as "Impulse" Noise</u>

While the clipping noise exhibits wideband spectrum, it really behaves like an impulsive noise infrequently occurring over random time intervals as experimentally demonstrated in [9] [10] and theoretically proved in [11][12]. In particular, for small probability of clipping, the clipping noise is shown to possess the following statistical properties [11][12]:

1. It is modeled as a Poisson impulse train.
2. Its pulse duration follows a Rayleigh distribution.
3. Its pulse shape is approximated by a parabolic arc.
4. Its mean pulse width is less than a nanosecond.
5. It is non-Gaussian distributed; its probability density function (PDF) is of the type [8][11]:

$$f(n) = (1 - \gamma)f_g(n) + \gamma f_w(n) \qquad (6)$$

where $f_g(n)$ is the PDF of the Gaussian component, $f_w(n)$ is the PDF of the non-Gaussian component, and γ is known as the clipping index (or impulsive index) which is equal to the product of the average number of impulses in a second and the mean impulse duration.

6. Its variance can be obtained from (1) or (4) or from the simplified equation [8][11]:

$$\sigma_w^2 \doteq \frac{8\bar{\tau}}{3} (\pi\rho)^{-3/2} e^{-\rho} R_{in}(0) \qquad (7)$$

where $\rho = 1/m^2N$ and $\bar{\tau}$ is the mean impulse duration. The in-band noise power is then $NLD_{in} = \sigma_w^2 \cdot B \approx P_w(f_c) \cdot B$.

To illustrate the impulse behavior of the clipping noise, in Fig. 4 we conceptually show how the impulse noise train is generated by the clipping phenomenon.

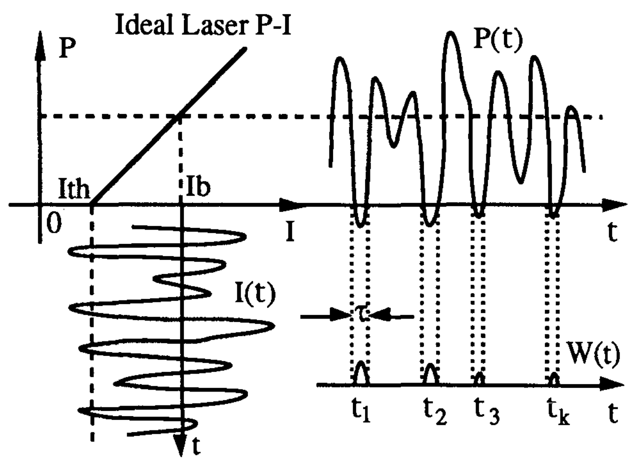


Fig. 4 Clipping-Induced Impulse Noise Train

The figure illustrates that the laser output signal which has been clipped may be decomposed into a sum of the attenuated input signal and a noisy impulse train w(t), viz.:

$$P(t) = KI(t) + w(t) \qquad (8)$$

where attenuation coefficient K can be obtained from the laser P-I curve and is equal to $K = h_1$. We note that this impulse-like time domain waveform responsive to clipping has also been observed by recent experiments employing the hybrid SCM signals [9][10].

3. BER PERFORMANCE OF M-QAM SUBJECT TO CLIPPING NOISE

BER Model

Having described the impulsive behavior of the clipping noise, we now study the performance of a digital M-QAM system impaired by the clipping noise in conjunction with a Gaussian background noise. First, we devise a BER model of M-QAM that can be used in such events. To this end, we use an analytic model of Fig. 1 which, as shown in Fig. 5, specifically indicates an M-QAM system embedded in both the clipping and Gaussian noises.

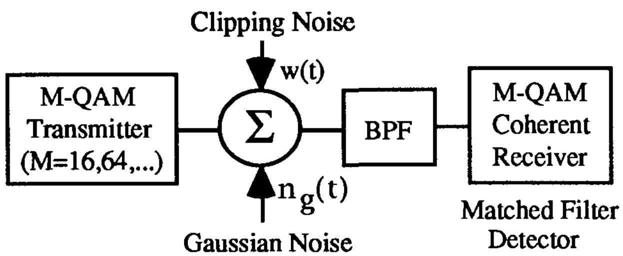


Fig. 5 Analytic Model for M-QAM Signal in the Presence of Clipping and Gaussian Noise Sources

In Fig. 5, the Gaussian noise refers to the additive background noise sources existant in the optical link including laser intensity noise RIN, shot noise, and thermal noise i_n; its variance is given by [7]-[11]:

$$\sigma_g^2 = i_n^2 + 2q(rP_0) + RIN(rP_0)^2 \qquad (9)$$

where q is the electron charge, r is the photodetector responsitivity, and P_0 is the received optical power. The clipping noise has a variance given by (7). For the hybrid system, the input correlation function $R_{in}(\tau)$ consists of both the AM and M-QAM signals. Thus (7) can be rewritten as [8][12]:

$$\sigma_w^2 = \frac{4\bar{\tau}}{3}\pi^{-3/2}\mu^5 e^{-1/\mu^2}(rP_0)^2 \qquad (10)$$

where $\mu^2 = N_a m_a^2 + N_q m_q^2 = \mu_a^2 + \mu_q^2$, N_a and N_q being the number of AM and QAM channels, and m_a and m_q being the OMI's of AM and QAM, respectively. From (10), we see that the clipping noise has contributions from both the AM and QAM clipping distortions. If μ_q is small compared to μ_a, which usually is the case, the AM distortion dominates the clipping noise.

For M-QAM corrupted by an additive noise, the BER is generally expressed by [13]:

$$BER = \frac{1}{\log_2\sqrt{M}}\left(1 - \frac{1}{\sqrt{M}}\right)\int_{d_m/2}^{\infty} f(n)dn \qquad (11)$$

where d_m is the minimum distance between adjacent states in the M-QAM constellation and is given by $d_{min}/2 = \sqrt{[3T_sP_{av}/(M-1)]}$, $T_s = 1/B$ is the symbol time and $P_{av} = m_q^2(rP_0)^2/2$ is the average signal power of M-QAM, and f(n) is the PDF of the noise. Hence, for the hybrid system, using the joint PDF of the clipping and Gaussian noises as represented by (6), the BER of M-QAM may be devised as follows [8]:

$$BER = \frac{1 - 1/\sqrt{M}}{\log_2\sqrt{M}}\left\{(1-\gamma_T)\mathrm{erfc}\left(\sqrt{\frac{1.5\Gamma_g}{M-1}}\right) + \gamma_T\left[\mathrm{erfc}\left(\frac{\Delta_1}{\sqrt{2}}\right) + \frac{9\gamma_T}{4}\cdot\frac{\Phi_3(\Delta_2)}{(2+\gamma_T G)^2} + \frac{21\gamma_T}{4}\cdot\frac{\Phi_5(\Delta_3)}{(3+\gamma_T G)^3}\right]\right\} \qquad (12)$$

where $\gamma_T = T_s\lambda = \gamma T_s/\bar{\tau}$ is the clipping index per symbol interval, $G = \sigma_g^2/\sigma_w^2$ is the variance ratio of the Gaussian to clipping noise component, $\Gamma_g = P_{av}T_s/\sigma_g^2$ is signal to Gaussian noise ratio, $\Delta_i = \sqrt{[3\Gamma_g/(M-1)(1+ i/\gamma_T G)]}$ (i=1,2,3), and $\Phi_k(z) = H_k(z)\exp(-z^2)/\sqrt{2\pi}$ (k = 1,2,3), $H_k(z)$ being the Hermit polynomials.

The first term of (12) indicates the effect of the Gaussian noise only, while the second term indicates the effect of both the clipping and Gaussian noises. Further, Eq. (12) shows that the BER of M-QAM in this case is specified by three parameters: Γ_g, γ_T, and G. From (9) and (10), we note that these three parameters all depend on system parameters such as the number of channels and OMI's of both the AM and M-QAM signals. Hence, the BER model

(12) can be utilized to evaluate the digital system performance under various conditions.

BER Performance

We now present several example cases, and also compare our results with the reported experimental results [9][10].

In general, when many M-QAM signals are added with AM signals, especially with higher-level modulations such as 64- or 256-QAM, the effect of the clipping of the QAM signals can not be ignored; the effect will not only affect the QAM signals themselves, but also the AM signals. This multi-effect will be addressed in the next section. In the following examples, we consider the AM to QAM interference only. That is, we consider the scenario in which a few M-QAM channels are multiplexed with the AM signals. For this case, the effect of clipping is mainly due to the clipping of the AM signals, since QAM requires much lower power, (e.g., 10~20 dB below), than AM for reliable transmission. This AM dominance on the clipping noise components has also been verified by one of the recent experiments [10].

In case 1, we use 42 AM carriers (55.25-337.25 MHz) and a 16-QAM located above the AM carriers with 6 MHz bandwidth. The BER results are shown in Fig. 6 versus the OMI of 16-QAM for three different AM OMI's: $m_a = 0$ (no AM carriers), $m_a = 5\%$ (C/NLD = 69 dB) and $m_a = 6\%$ (C/NLD = 53 dB). The solid BER curves are the analytic results obtained by (12), and the black dots are the experimental data.

Fig. 6 shows that for 42 AM carriers, even with a typical 5% AM OMI, the effect of clipping on BER is significant; the BER increases greatly as compared to the no AM case. For a higher 6% AM OMI, the BER begins to show a floor. Increasing the OMI of 16-QAM, i.e., increasing the signal power will no longer reduce the BER. For this case, other means may have to be applied, such as coding techniques.

Fig. 6 also shows a strong correlation between the model and the experiments; both show a similar BER behavior and the two appear to agree for small OMI's of 16-QAM and for the high clipping case. The difference between the two results may possibly lie in the hardware limitations and other effects that are not dealt with by the model.

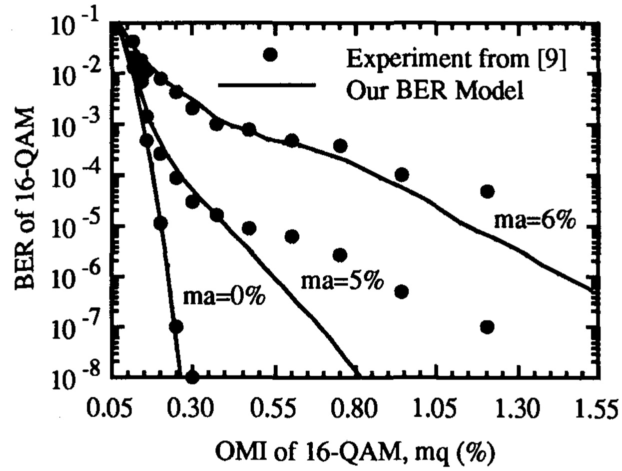


Fig. 6 BER Performance of a 42-AM/16-QAM System. RIN=-150 dB/Hz. i_n=35 pA/$\sqrt{\text{Hz}}$. r = 0.87 A/W. P_0=-1 dBm.

In case 2, we use 60 AM carriers (55.25-439.25 MHz) and a 64-QAM with 15 MHz bandwidth. Again, three different values of AM OMI's are used: $m_a = 0$, m_a=4%, and m_a=4.6%. The BER's obained from (12) along with the experimental results extracted from [10] are plotted in Fig. 7 versus the OMI of 64-QAM.

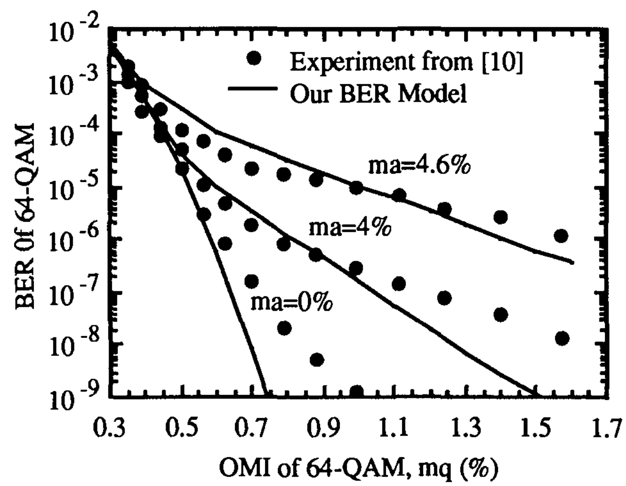


Fig. 7 BER Performance of a 60-AM/64-QAM System. RIN=-155 dB/Hz. i_n = 30 pA/$\sqrt{\text{Hz}}$. r = 0.9 A/W. P_0 =-1 dBm.

Fig. 7 shows the similar BER behavior to Fig. 6; i.e., the BER degrades significantly with the increase of the AM OMI. Again, the analytic and experimental results exhibit the similar BER decreasing trend, and agree reasonably well for small OMI's of 64-QAM.

Notice that from Figs. 6-7, one can also find the (average) signal level of the 16/64-QAM relative to AM, at a certain BER. As an example, in Fig. 7 consider BER=10^{-8} and m_a = 4%. Then, from Fig. 7 the model gives m_q = 1.25%, about 10 dB ($20 \cdot \log_{10} m_q/m_a$) below the AM carriers, whereas the experiment [10] gives m_q = 1.53%, about 8 dB below the AM carriers.

In case 3, we show how the BER of M-QAM is further influenced by the OMI of AM. The case is shown in Fig. 8 using 64-QAM, with m_q = 0.75% which gives a BER of 10^{-9} with no AM carriers and is 13 dB below the AM carriers (m_a = 3%). For comparison purpose, Fig. 8 also includes the results obtained by assuming the clipping noise as Gaussian noise; the results are indicated by dashed lines.

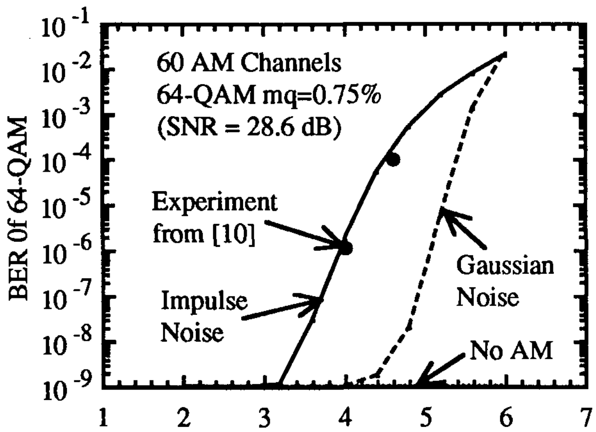


Fig. 8 BER of 64-QAM vs. the OMI of AM .

Fig. 8 shows that to have a required BER, the OMI of AM needs to be operated at a lower level than it would otherwise; e.g., the OMI of AM reduces from 4% to 3% for a 10^{-9} BER. One may also keep the AM level but increase the QAM level to obtain a desired BER. The trade-offs are that the former will reduce the CNR of the AM signals, while the latter will limit the number of QAM channels that can be applied. These trade-offs will be studied in the next section. Finally, we see that the BER obtained based on the impulse noise model agrees more closely to the experiments, and is much higher in value than predicted by the Gaussian noise model.

4. PERFORMANCE LIMITS AND DESIGN TRADE-OFFS

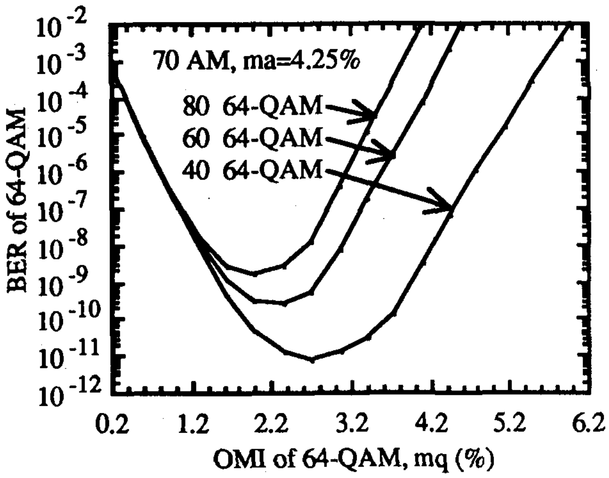
We have shown that our BER model is capable of predicting the BER behavior of M-QAM under various clipping conditions, and agrees reasonably well with the experiments. Here, we show that the model can be used to determine the transmission limits of M-QAM and perform trade-off studies.

Transmission Limits of M-QAM

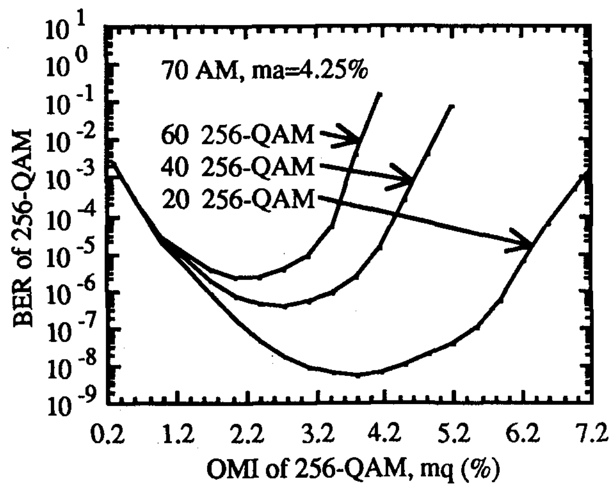
For Cable engineers, the limiting conditions of system parameters, such as the number of channels, OMI, and required optical power, are essential to system designs. In analog systems, CNR and C/NLD (or CSO/CTB) are used to obtain these transmission limits [1]-[4]. In digital systems, the BER is also required, in addition to CNR and C/NLD [8]. Analytically, the transmission limits may be obtained by minimizing the BER model of (13) with respect to, say, the OMI of M-QAM for a specified CNR and C/NLD. However, due to the complexity of the equation, there may be no closed-form solution. A simple means would be a graphic solution, as illustrated below.

In Figs. 9(a)-(b), we show the BER's of 64-QAM and 256-QAM, respectively, for different numbers of channels, multiplexed with 70 AM carriers. In each case, the AM OMI m_a = 4.25% which yields a CNR of 52.5 dB and a C/NLD of 60 dB for the analog channels with digital signals absent [4]. Both cases show that a minimum BER exists for each channel number. The reason is due to the fact that for small digital OMI's, the clipping effect from AM signals dominates the BER, and for large digital OMI's, then the clipping effect from QAM signals dominates the BER, as indicated by the figures. Hence, a minimal BER exists when the two effects are equal.

Thus, one way to determine the channel capacity is to first specify a minimum BER, and then to search by means of the BER model for an optimal OMI of QAM and a maximal channel number that meets the specified BER. For example, given a minimum BER of 10^{-9}, Fig. 9(a) shows a limit of eighty 64-QAM channels operating at 2% OMI, and Fig. 9(b) shows that less than twenty 256-QAM channels can be accommodated.



(a) 64-QAM



(b) 256-QAM

Fig. 9 BER of M-QAM as a Function of the Number of M-QAM Channels for (a) M = 64 and (b) M = 256. RIN = -155 dB/Hz. i_n = 12 pA/$\sqrt{\text{Hz}}$. r = 0.9 A/W. P_0 = 0 dBm. B=6 MHz.

Trade-Offs Among System Parameters

The trade-off study is another crucial step in system design. In a hybrid analog/digital SCM system, the trade-offs exist when we consider the interference of analog to digital signals or digital to analog signals.

For the effect of AM on QAM, as we have shown, the clipping of the AM signals yields significant BER degradations. To meet a specified BER, the OMI of M-QAM must be operated at a higher value than the no clipping case. This sets a limit for the number of digital channels to be used. For example, Fig. 10 shows that for an OMI of 256-QAM as small as 0.55%, there is virtually no BER difference between the one and one hundred channel cases. But, if a 2.5% OMI is used, the BER difference is visible even when 20 channels of 256-QAM are applied.

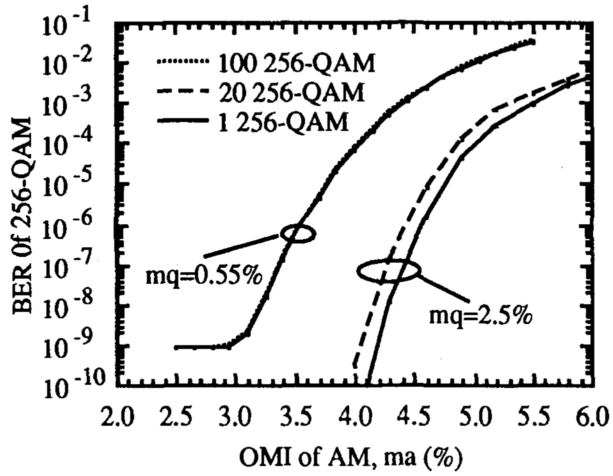


Fig. 10 BER of 256-QAM vs. OMI of AM (70 AM)

For the effect of QAM on AM, operating the M-QAM power (or the OMI) at large values in order to meet the required BER, will inevitably reduce the specified C/NLD (or CSO/CTB) of the AM signals. On the other hand, if the OMI of QAM is kept small, the signal level of AM (i.e., the OMI of AM) must be reduced to obtain the desired BER, as shown in Fig. 10. This will affect the CNR of the AM signals. Considering, again, Fig. 10, for BER = 10^{-9} a 20-channel 256-QAM operating at 2.5% OMI, which is only 4.5 dB below AM, would give a 5 dB C/NLD (or similarly CSO/CTB) reduction in AM channels. If, however, a 0.55% OMI is used, Fig. 10 shows that the OMI of AM must be reduced to 3% from the previous 4.25% to have the same BER of 10^{-9}. This implies a reduction of 3 dB in CNR.

In short, the above mentioned trade-offs for both analog and digital channels must be taken into account in a hybrid system design. One approach would be to first specify the CNR and C/NLD (or CSO/CTB) requirement for AM with a margin of a few dB's and the BER requirement for M-QAM, and then determine the other system parameters by means of the

BER model. We will illustrate this idea via the following design cases.

Design Study Cases

The architecture we consider is the popular fiber-to-the-feeder (FTF) architecture, with one fiber per node and one laser per fiber.

The objective is to provide for system designers and/or Cable engineers the reference values of system parameters involving both the analog and digital channels and the trade-offs among various parameters, given certain requirement specifications.

Specifically, we here consider the following scenario: given a link budget and the number of channels for both AM and M-QAM systems, how would a designer choose a laser diode and what sacrifice he has to make in order to meet the system specifications?

First, we look at what is given (the numbers are for illustration purpose):

1. **Link budget**; assume 8 dB link budget which would give a 20 km transmission with 0.4 dB/km link loss including fiber and splice loses.
2. **Channel capacity**; assume 70 AM channels and 80 M-QAM channels, making a total of 150 channels.
3. **Optical detector**; assume responsitivity $r = 0.9$ A/W and thermal current density $i_n = 12$ pA/$\sqrt{(Hz)}$.

Next, we set up requirement specifications:

1. **Analog**: CSO/CTB $\leq$ -70 dBc (C/NLD $\geq$ 55 dBc).
2. **Digital**: (Uncorrected) BER ~ (on the order of) 10^{-5}.

Note that the first requirement is set based on the clipping distortion only; it is intended to leave some room for normal nonlinearity of a laser diode which when considered would bring down the CSO/CTB to -65 dBc or -60 dBc. The second requirement is set without any error correction such as coding applied.

Finally, based on the given conditions and the requirement specifications, we can then determine the desired analog and digital system parameters by means of the developed clipping distortion and BER models. The parameters may include the CNR and OMI of AM and the SNR and OMI of M-QAM. Note that iteration steps may have to be used to obtain the desired parameters that satisfy all requirements simultaneously. Table 1 lists a set of these parameter values obtained for two DFB lasers with different output powers and RIN's, respectively, using 64-QAM as an example.

8 dB Link Budget. 150 AM/64-QAM Channels

Lasers Opt. Received Power	4 mw -2 dBm		8 mw 1 dBm	
RIN (dB/Hz)	-150	-155	-150	-155
CNR of AM (dB) OMI of AM (%)	50.7 4.45	52.5 4.45	52.3 4.45	55.2 4.45
SNR of QAM (dB) OMI of QAM (%) Level Below AM (dB)	33.6 0.84 14.5	34.92 0.84 14.5	34.77 0.84 14.5	36.3 0.84 14.5

Table 1. System Parameters for Two DFB Lasers

From Table 1, we observe that (a) the OMI's of AM and QAM are unchanged in all cases, (b) for same RIN, the higher the laser power the higher the CNR (or SNR), and (c) the 4 mw laser with -155 dB/Hz RIN gives a slightly higher CNR (SNR) than does the 8 mw laser with -150 dB/Hz RIN. This implies that a low power laser with smaller RIN may be traded off for a better performance with a high power laser with a larger RIN.

Other trade-offs can also be made, such as reducing the analog or digital requirements. Table 2 gives two other cases in which the CSO/CTB $\leq$ -65 dBc or BER ~ 10^{-3} using the 8 mw laser with a RIN of -155 dB/Hz. As expected, we see that in both cases, the CNR and OMI of AM do get improved. In the former case, the SNR and OMI of 64-QAM also improve.

8 dB Link Budget. 150 AM/64-QAM Channels

COS/CTB BER	$\leq$ -70 dBc ~ 10^{-3}	$\leq$ -65 dBc ~ 10^{-5}
CNR of AM (dB) OMI of AM (%)	55.36 4.53	55.5 4.6
SNR of QAM (dB) OMI of QAM (%) Level Below AM (dB)	27.5 0.3 23.6	37.0 1.1 12.4

Table 2. Different Trade-off Conditions

As a final remark, we point out that the above cases use one set of parameters and conditions to illustrate the idea. Practically, one may need to use different system parameters and requirements to satisfy the designer's needs.

5. CODING TO REDUCE THE EFFECT OF CLIPPING DISTORTION/NOISE

All the BER's of M-QAM presented above, as an effect of clipping noise, were assumed uncorrected. We now show how coding would improve the performance.

Channel Coding - Error Correcting Codes

Channel coding employing error correcting codes are widely used in digital communication systems to reduce the effect of noise. Traditionally, coding has been used in an additive Gaussian noise channel. Here, we demonstrate that coding is also effective for reducing the impact of the clipping noise.

Specifically, we consider the use of a Reed-Solomon (RS) code [13], which is developed for multi-level (nonbinary) digital systems and is capable of correcting both the random and burst types of symbol errors. Consider a RS code with N total symbols and K information symbols. Then, the code is guaranteed to correct up to t = [(N-K)/2] symbol errors [13]. We denote this code as RS(N,K,t) code. The code rate of a RS(N,K,t) code is defined as $R_c = K/N$. Note that in a coded digital communica-tion system, the information data rate or the bandwidth of the system is increased by an amount of $1/R_c$ since channel coding introduces redundant symbols in the data. This poses another trade-off factor for the hybrid system design, as demonstrated below.

Decoded BER and Coding Gain

For RS codes, if hard decision decoding is used, the decoder output BER can be expressed by [13]:

$$BER_c = \frac{N+1}{2N^2} \sum_{i=t+1}^{N} \binom{N-1}{i-1} SER^i (1 - SER)^i \quad (13)$$

where SER is the symbol error probability at the decoder input. For M-QAM, the SER and BER (uncoded) are related by [13]

$$SER = \log_2 M \cdot BER \quad (14)$$

One important parameter to measure the benefit of coding is the coding gain, defined as the difference of the signal to noise ratios before and after coding, for the same specified BER. In a system design, this quantity can be utilized to determine a proper error correction code or codes to compensate certain BER loss.

Example One - BER Improvement

In this case, we consider a hybrid system of 70 AM carriers and one 64-QAM channel. We use a RS code with N = 63, t = 1 (single-error correction) and t = 2 (double-error correction), respectively. Both uncoded and coded BER's are shown in Fig. 11. It shows that even a one-error correcting RS code can reduce the BER significantly. For instance, at m_q = 0.8%, the BER is reduced from 10^{-6} to 10^{-9}. The higher the error correction capability, the lower the decoded BER. However, the BER improvement will be traded off with an expansion of bandwidth as mentioned earlier.

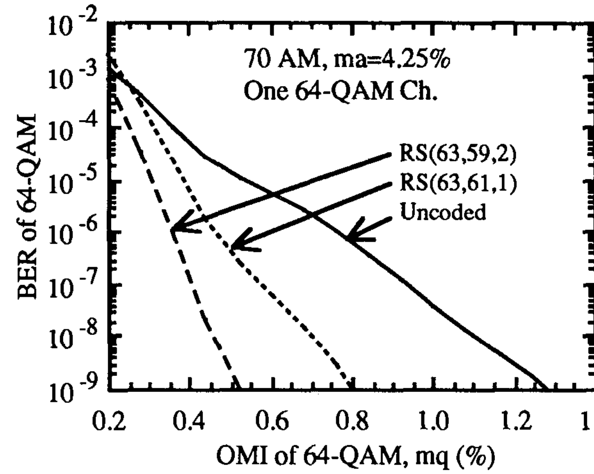


Fig. 11 Uncoded/Coded BER's of 64-QAM with 70 AM. The optical parameters are the same as those in Fig. 9.

From Fig. 11, we can obtain the coding gain at a certain BER. For example, at BER = 10^{-9} the RS(63,61,1) and RS(63,59,2) codes give 4 dB and 7 dB coding gains, respectively.

Example Two--Coding to Increase Capacity

In this case, we show that coding also increases the (digital) channel capacity. We consider a hybrid 70-AM and sixty 256-QAM system. For this case, we use a RS code with N = 255. Fig. 12 shows the results using double-error correction ($t = 2$). Also shown is the case of 75 channels of 256-QAM with the same 2-error correcting RS code.

We observe that the coding reduces the minimum BER, enlarges the workable BER range, and increases the number of channels. For example, the figure shows that the minimum BER is reduced from larger than 10^{-6} (uncoded) to less than 10^{-9} (coded). If a coded BER of 10^{-9} is required, seventy-five (75) 256-QAM channels can be used, implying a 15 channel increase compared to the uncoded case. Increasing the t value will increase the channel capacity. For example, with $t = 3$, i.e., a triple-error correcting RS code, then 125 channels of 256-QAM can be accommodated given the same BER of 10^{-9}; this gives more than twice capacity increase over the uncoded case. Again, the benefit has to be traded-off with the bandwidth expansion or an increase in the information data rate.

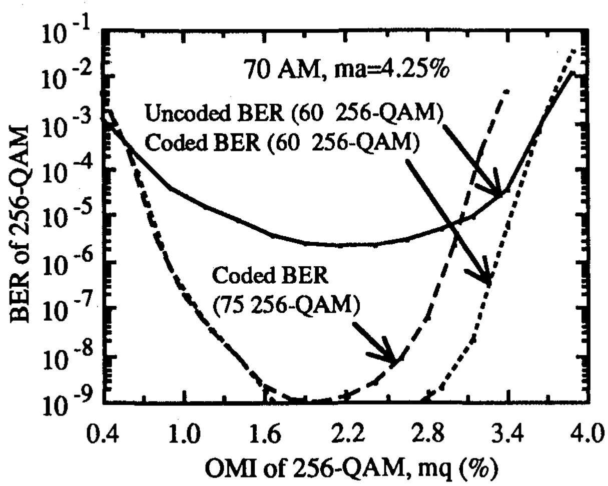


Fig. 12 Uncoded/Coded BER's of 256-QAM in hybrid AM/QAM. The parameters are the same as Fig. 9.

Finally, we note that concatenated codes [13], comprised of inner and outer codes, may also be employed in such environments. Using a concatenated code, particularly with a trellis code as inner code and a RS code as outer code [13], gives an advantage of no further bandwidth expansion in the inner code part and yet produces a lower BER, a higher coding gain, or a higher channel capacity than that using the RS code alone.

6. CONCLUDING REMARKS

We have developed a model that can be utilized to characterize the clipping-induced distortion impairments encountered in a hybrid SCM analog/digital fiber optic system. Specifically, the model is developed for Cable engineers to estimate both the analog and digital performances when implementing a hybrid fiber optic system. In terms of analog (AM) signals, the model estimates the C/NLD and CSO/CTB induced by the clipping distortion. In terms of digital signals, the model is then used to predict the BER of a digital channel, subject to the clipping-induced noise.

The results herein presented have shown that the model agrees reasonably well with the experimental results involving both the analog and digital signals. The results have also shown that in a hybrid AM/M-QAM system, the M-QAM signals may suffer a significant (BER) performance degradation due to the presence of the impulsive clipping noise. In certain cases, BER floors have been obtained. These effects have also been demonstrated in recent experiments involving hybrid optical transmission [9][10].

A major benefit of an analytic model is that one can use the model to evaluate system performance under various conditions, without conducting extensive experiments and computer simulations. To illustrate this benefit, the model, particularly, the BER model herein developed, has thus been employed to determine the transmission limits, such as channel capacity and OMI's, and to perform trade-off analysis for both the AM and M-QAM signals. A design example has also been presented using practical system parameters and requirements. The results presented indicate that as a result of the impulsive non-Gaussian clipping noise, only a limited number of M-QAM channels can be accommodated if high-level M-QAM signals, such as 64-QAM or 256-QAM, are used. Further, the transmitted power of M-QAM must also be higher than in the Gaussian noise case. The trade-off analysis and the design example show that both the analog and digital requirements (e.g., CNR, C/NLD or CSO/CTB, and BER) have to be carefully balanced for doing a hybrid optical transmission system design.

Means have also been sought to reduce the effect of the clipping distortion/noise. In particular, we have shown that applying coding techniques, i.e., error-correction techniques, can effectively reduce impact of the clipping noise on M-QAM signals and therefore improve the BER performance significantly. The examples presented using the RS codes show that the BER of M-QAM (e.g., 64-QAM) can be reduced significantly even with a single-error correcting RS code. The results further show that coding also increases channel capacity and enlarges the workable range of BER (see the 256-QAM example). The benefit of coding must, however, be sacrificed with an expansion of bandwidth or information rate increase. The trade-off should also be considered in system design stages.

Finally, we note that another digital modulation scheme known as multi-level vestigial sideband modulation (M-VSB), has been developed [14] and is becoming popular in Cable community, due to some advantages over the M-QAM system. The M-VSB system still employs amplitude modulation and exhibits similar BER behavior to M-QAM, though a slight BER improvement is observed [14]. Hence, M-VSB will still be subject to the impairments of the clipping distortion/noise when it is used in a hybrid AM/M-VSB fiber optic environment. Hence, our model herein developed for M-QAM may also be modified without great difficulty to deal with the M-VSB signals. This work is currently under way in our laboratories.

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