Lightwave Systems With Optical Amplifiers

N. A. OLSSON

Abstract—Recent advances have brought semiconductor amplifiers to the stage where lightwave systems employing amplifiers in some aspects clearly out-perform traditional systems. For example, the longest nonregenerated fiber transmission experiments use optical amplifier repeaters, and at very high data-rates (4 Gbit/s), receivers with optical pre-amplifiers are substantially more sensitive than their coherent and APD counterparts.

In this paper, fiber optic communication systems employing semiconductor laser amplifiers are investigated theoretically and experimentally. The noise and bit-error-rate characteristics of lightwave systems with optical amplifiers are calculated and the dependence of system performance on amplifier characteristics such as optical bandwidth, noise figure, gain, etc., is shown. Experimental results are presented on both a 4 Gbit/s optical pre-amplifier as well as coherent and direct detection systems with four in-line amplifiers.

INTRODUCTION

The traditional way of compensating for optical loss in lightwave communication systems has been the rather cumbersome procedure of regeneration. Regeneration includes photon-electron conversion, electrical amplification, retiming, pulse shaping and finally electron-photon conversion. In many applications, direct optical amplification of the light signal would be advantageous. Optical amplifiers can be used in any system that is loss limited; i.e., dispersion effects are not a limiting factor. This is the case for most systems operating near the dispersion minimum at 1.3 μm, and for coherent lightwave systems. Local area networks, where the main losses are from branching and taps, are also loss limited and can benefit from simple optical amplifiers.

Semiconductor laser amplifiers have been studied for a number of years. Significant work at 0.8-μm wavelength was done in the early 80's [1]-[2]. Recently major progress has been made in long wavelength devices. Optical amplifiers with high gain, low gain ripple, low noise, and high saturation output power have been reported [3]-[7]. Optical amplifier system applications have also been reported, both applications for preamplifiers [7]-[9] and in-line amplifiers [10]-[14].

As optical amplifiers have advanced to the stage that actual system use might be possible in the near future, it is important to know the system consequences, its advantages and limitations. In this paper we present a theoretical as well as experimental investigation of optical amplifier lightwave systems. Noise levels, bit-error-rate characteristics (BER), receiver sensitivities, and power penalties are calculated functions of the relevant optical amplifier parameters.

THEORY DIRECT DETECTION

The amplifier noise model presented below is based on the work by Mukai [2] and Simon [15] which we have extended and combined with the receiver models and BER calculations of Smith et al. [16]. The various symbols are defined in Table I and a schematic is shown in Fig. 1. The analysis presented here applies to traveling wave amplifiers (TWA) which are the technologically most important type of amplifier. However, extension of the analysis to resonant or Fabry-Perot amplifiers (FPA) is rather straightforward by modifying the optical bandwidths for the beat noise components and by including the excess noise factor from the mirror reflectivities as described in [2].

The spontaneous emission power at the output from an optical amplifier is given by (see Table I for definitions of the symbols):

\[ P_{sp} = N_{sp}(G - 1)h\nu B_{\nu}. \] (1)

For an ideal amplifier, \( N_{sp} = 1 \). For semiconductor laser amplifiers, however, \( N_{sp} \) ranges from 1.4 to more than 4 depending both on the pumping rate and the operating wavelength [3], [6]. In the following we will write the optical powers as their photo current equivalent, i.e., as the photo current that would be generated by detecting the optical power with a detector with unity quantum efficiency.

The photo current equivalent of the spontaneous emission power is:

\[ i_{sp} = P_{sp}e/\hbar\nu = N_{sp}(G - 1)e\nu B_{\nu}. \] (2)

After square law detection in the receiver, the received signal power is given by:

\[ S = (GI_{in}\eta_{out}L)^2. \] (3)

The noise terms are:

\[ N_{shot} = 2B_{\nu}\epsilon_{out}L(GI_{in} + I_{sp}) \] (4)
\[ N_{s - sp} = 4GI_{in}\eta_{out}L^2B_{\nu}I_{sp}/B_{\nu} \] (5)
\[ N_{sp - sp} = (I_{sp}\eta_{out}L)^2B_{\nu}(2B_{\nu} - B_{\nu})/B_{\nu} \] (6)
\[ N_{th} = I_{th}^2. \] (7)

And the total noise is:

\[ N_{tot} = N_{shot} + N_{s - sp} + N_{sp - sp} + N_{th}. \] (8)
where

The BER is given by

\[ P_{\text{BER}} = \frac{1}{\sqrt{2\pi}} \exp\left(-\frac{Q^2}{2}\right) \] (11)

where \( Q \) is given by:

\[ Q = \frac{\sqrt{S(1)} - \sqrt{S(0)}}{\sqrt{N_{\text{tot}}(1) + N_{\text{tot}}(0)}} \] (12)

\( S(1), S(0) \) and \( N_{\text{tot}}(1), N_{\text{tot}}(0) \) are the signal and total noise for a mark and space, respectively. A BER of \( 10^{-9} \) requires \( Q = 6 \). Equations (1)–(12) form the basis from which the performance of the amplifier systems are evaluated.

### RESULTS

**Preamplifier**

Of particular interest for preamplifier applications is the receiver sensitivity dependence on the amplifier gain, noise figure, and optical bandwidth. The application of optical preamplifiers is at very high data rates where avalanche photo detectors are limited by their gain–bandwidth product [17]. Therefore, we have analyzed a 5-Gbit/s receiver with 2.5-GHz electrical bandwidth. First we calculate the achievable receiver sensitivity versus amplifier gain. The results are shown in Fig. 2. To make the calculations realistic, we have used measured values for the receiver and amplifier parameters. The value chosen for the thermal noise current corresponds to a base receiver sensitivity of \(-25 \text{ dBm}\). At low amplifier gains, the receiver is limited by the thermal noise and consequently the receiver sensitivity improves 1 dB for every decibel of gain. For higher gains, the signal-spontaneous and spontaneous-spontaneous beat noise become dominant, and the best achievable receiver sensitivity depends on the optical bandwidth of the amplifier. The ultimate limit is achieved when \( B_a = 2B_r \) and is in this case equal \( -39.4 \text{ dBm} \). This corresponds to 181 photons/bit and is close to the expected value of 41 \( N_{\text{sp}}/\eta_{\text{in}} \). The power penalty versus optical bandwidth with the gain as parameter is shown in Fig. 3. The penalty is defined as the decrease in receiver sensitivity as compared to \( B_a = 2B_r \) for each given gain. As expected, the penalty is larger for higher gains when the receiver is closer to the ultimate limit. The receiver sensitivity dependence on the spontaneous emission coefficient \( N_{\text{sp}} \) is shown in Fig. 4 for an optical bandwidth of 10 A and for amplifier gains of 15, 20, and 35 dB. For low amplifier gains, the sensitivity dependence on \( N_{\text{sp}} \) is quite weak because we are far from the ultimate sensitivity and the receiver is still dominated by thermal noise. At high gain, in the signal-spontaneous beat noise limit, however, the sensitivity is inversely proportional to \( N_{\text{sp}} \). A 3-dB increase in \( N_{\text{sp}} \) gives a 3-dB reduction in sensitivity. Like APD receivers, optical preamplifier receivers have signal dependent noise. Therefore, the sensitivity degradation from a nonperfect extinction ratio is worse than in a traditional p-i-n receiver.

The effect of the extinction ratio is shown in Fig. 5 as the sensitivity penalty at amplifier gains of 15 dB and 35 dB and an optical bandwidth of 5 A. The extinction ratio is here defined as the ratio of the optical power in the "1" state to that in the "0" state. The higher the gain, the more prominent is the signal-spontaneous beat noise \( (N_{\text{sp}}) \) term and consequently the higher is the extinction ratio penalty. An extinction ratio of 20:1 or better will in all circumstances give a penalty of 1 dB or less.

Overall, using very reasonable and achievable values for the amplifier and system parameters \( (G = 25 \text{ dB}, \eta_{\text{in}} = \ldots) \),
Fig. 2. Optical preamplifier sensitivity versus amplifier gain. Calculated for $B_p = 2500$ MHz, $N_{sp} = 1.4$, $\eta_m = 0.31$, $\eta_{in} = 0.26$, and $\lambda = 1.55 \mu m$.

Fig. 3. Optical preamplifier sensitivity penalty versus optical bandwidth. Same amplifier parameters as Fig. 2.

Fig. 4. Optical preamplifier sensitivity versus spontaneous emission factor $N_{sp}$. Same amplifier parameters as Fig. 2.

Fig. 5. Optical preamplifier sensitivity penalty versus extinction ratio. Same amplifier parameters as in Fig. 2.

One of the more promising applications of optical amplifiers is the in-line amplifier, where the amplifier compensates for losses in the system. The losses may be due to fiber losses in a long haul system, tap and splitting losses in a local area network, or coupling and switch losses in an optical switch. For this application, the main concerns are how the system performance is affected by gain, optical bandwidth and the input power of the amplifier. Also of interest is the ultimate number of amplifier stages that can be cascaded while maintaining reasonable system performance. Several of these issues have been addressed previously [2]. However, prior work has been concerned only with optical signal-to-noise ratios which give an indication of upper limits on the system performance but do not provide information about major system parameters such as power penalties and noise induced error floors. In this section we will give a detailed analysis of the performance of lightwave systems utilizing optical in-line amplifiers.

First we will address the question: what power penalties are incurred at the receiver when a single in-line amplifier is inserted somewhere in the fiber line? For this case, the most important parameter is the optical input power to the amplifier. The calculation is made as follows. For each given input power to the amplifier, the loss between the amplifier and receiver is varied such as to produce a signal power at the receiver that gives an BER of $10^{-9}$. No approximations are made and all of the noise terms in equations (4)-(7) are used to evaluate the BER. This receiver sensitivity is then compared to the baseline receiver sensitivity (no amplifiers) and the power penalty is taken as the sensitivity difference. This calculation is shown in Fig. 6 and Fig. 7. In Fig. 6 for electrical bandwidths of 2500, 500, and 50 MHz, all for an optical bandwidth of 50 A. The corresponding base-line sensitivities are: -25, -32, and -42 dBm respectively. Fig. 7 is a calculation of the actual receiver sensitivity for a electrical bandwidth of 2500 MHz and for optical bandwidths of 300, 30, and 3 A, respectively. Note that for this calculation we have used $\eta_m = 1$ so the amplifier input power refers to the power actually coupled into the amplifier. To give some guidance of what power penalty is acceptable for a system, we show in Fig. 8 a family of BER versus received power curves with the amplifier input power as parameter. The amplifier input powers have been chosen such that the $10^{-9}$ BER receiver power penalties are $-0$, $-35.2$ dBm. This is still 9 dB from the ultimate limit but is about 4 dB better than the best reported APD results. As have been pointed out earlier, optical preamplifier receivers are a viable alternative to APD receivers but mainly at high data rates [17]. The above analysis should give some guidance in the relative importance of the various system and amplifier parameters.
DIRECT DETECTION

To modify (1)-(12) to correspond to a coherent detection system is straightforward. For an amplitude shift keying (ASK) system, which we will use as our model coherent system, the signal term (3) is replaced by

$$S_{coh} = 2I_{lo} I_{in} \eta_{in} L G$$

(13)

where $I_{lo}$ is the photocurrent equivalent of the local oscillator power $P_{lo}$:

$$I_{lo} = P_{lo} e/(\hbar v).$$

(14)

To the total noise power in (8), we must add the local oscillator shot noise: $2I_{lo} B_e e$, and the local oscillator-spontaneous beat noise:

$$I_{lo-sp} = 4I_{lo} L(G - 1) N_{sp} e \eta_{out} B_e.$$

With these modifications, we can calculate the receiver sensitivity versus amplifier input power analogous to Fig. 7 for the direct detection case. The result is shown in Fig. 9 for a 500-MHz electrical bandwidth and a local oscillator power of 9 dBm. The receiver thermal noise corresponds to a direct detection sensitivity of $-32 \text{ dBm}$. The coherent baseline receiver sensitivity is $-52.8 \text{ dBm}$ or $41 \text{ photons/bit (1 Gbit/s)}$. A 2-dB power penalty is incurred for an amplifier input power of $-44.5 \text{ dBm}$. Contrary to the direct detection case, for coherent detection the power penalty is independent of the optical bandwidth. In Fig. 10 we show the calculated BER curves for the coherent detection system. We have chosen the amplifier input powers such that the penalties at $10^{-9}$ BER are $-0, 1, 2, 3, 4,$ and $5 \text{ dBm}$. In this case, even for a $10^{-9}$ penalty of 2 dBm, no error floor exist for BER $> 10^{-9}$. The difference between the direct detection and coherent detection stems from the quadratic dependence of the electrical signal on optical power for direct detection and linear dependence in the case of coherent detection.

It is clear from the above analysis that a single amplifier can be inserted in both direct detection and coherent detection systems with no or very small penalty provided the amplifier input power is sufficiently large.
Next we will analyze a system with multiple in-line amplifiers. We will assume that the gain of each amplifier exactly equals the loss in between two amplifiers. With this assumption, the cumulative effect of the amplifier noise is obtained by replacing $N_{sp}$ in equations (2)-(12) by $N N_{sp}$, where $N$ is the number of amplifiers. First we calculate the power penalty at the receiver as function of $N$ and with the amplifier input power as a parameter. Similarly to the calculation for Fig. 6; for each input power and optical bandwidth, we vary the loss after the last amplifier as to produce a signal power at the receiver that gives a BER of $10^{-9}$. Fig. 11 show the calculated results for a system with 500-MHz electrical bandwidth and an optical bandwidth of 3 A. For amplifier input powers of $-30$, $-20$, and $-10$ dBm, the maximum (1-dB penalty) number of amplifiers are 28, 280, and 2800, respectively. This example with very narrow optical bandwidth and low data rate ($1$ Gbit/s) gives the upper limit on the number of amplifiers. If we increase the data rate to $-5$ Gbit/s ($B_e = 2500$ MHz) and the optical bandwidth to 300 A, which is approximately the unfiltered bandwidth of a 1.3-µm amplifier, the maximum number are 3, 30, and 300 for input powers of $-30$, $-20$, and $-10$ dBm, respectively, as shown in Fig. 12.

Coherent Detection

The penalty versus number of amplifiers for a coherent detection system is shown in Figs. 13 and 14. Fig. 13 for a 500-MHz electrical bandwidth, and Fig. 14 for 50-MHz electrical bandwidth. As in the direct detection case, the maximum number of amplifiers that can be used before a given power penalty is reached, is directly proportional to the amplifier input power and very large number of amplifiers can be cascaded without excessive power penalties. For example, with an electrical bandwidth of 50 MHz, and an amplifier input power of $-30$ dBm, 250 amplifiers can be used with a penalty of about 2 dB. In all the above calculations, we assumed an amplifier gain of 25 dB per stage. The results, however, as shown in Figs. 11-14 are independent of the amplifier gain as long as one makes no distinction between $G$ and $G - 1$.

A rough estimate of the maximum number of amplifiers can be made from the condition that amplifiers can be added as long as the amplifier noise is less than the dominant noise term of the receiver. For a direct detection receiver, the dominant noise term is the thermal noise and the amplifier added noise comes from the spontaneous-spontaneous beat noise for large optical bandwidths and from the signal-spontaneous beat noise for narrow optical bandwidths. In the first case the requirement becomes

$$N_{th} > (N_{max} N_{sp} G e L_0^2 B_e 2 B_o)$$  \(15\)
where $L$ is the loss between the last amplifier and the receiver. By using

$$P_{\text{sens}} = P_{\text{in}} G L$$

(16)

where $P_{\text{sens}}$ is the sensitivity of the receiver, (15) can be written:

$$N_{\text{max}} = \frac{P_{\text{in}}}{17N_{\text{sp}} h v / \sqrt{B_{\text{o}} B_{\text{o}}}}$$

(17)

where we have used the fact that the signal to noise ratio at $10^{-9}$ BER is $\sim 144$. For $B_{\text{o}} = 2500$ MHz, $B_{\text{o}} = 300$ A, $P_{\text{in}} = -20$ dBm, and $N_{\text{sp}} = 1.4$, (17) gives $N_{\text{max}} = 33$ in good agreement with the exact calculation of Fig. 12. When a narrow optical filter is used, the main amplifier added noise is the signal–spontaneous beat noise. In this case the condition on $N_{\text{max}}$ can be written:

$$N_{\text{max}} < \frac{P_{\text{in}}}{(576B_{\text{o}} h v N_{\text{sp}})}$$

(18)

yielding a $N_{\text{max}} = 40$ for $B_{\text{o}} = 2500$ MHz, $P_{\text{in}} = -20$ dBm, and $N_{\text{sp}} = 1.4$. The crossover between these two cases occurs when

$$B_{\text{o}} = \frac{P_{\text{in}}}{(2N_{\text{sp}} h v)}$$

(19)

giving $B_{\text{o}} = 45$ A for $N = 50$ and $P_{\text{in}} = -20$ dBm. For coherent detection, finally, the dominant receiver noise is the local oscillator shotnoise and the amplifier added noise is the local oscillator–spontaneous beat noise. In this case the restriction on $N_{\text{max}}$ is:

$$N_{\text{max}} < \frac{P_{\text{in}}}{(2P_{\text{sens}} N_{\text{sp}})}$$

(20)

which for an input power power of $-20$ dBm and a sensitivity of $-52.8$ dBm gives a $N_{\text{max}}$ of the order 600.

It is clear from the above analysis that very large numbers of amplifiers can be cascaded without prohibitive noise penalties. In most cases, in real systems, the limitations will not come from amplifier noise but from other considerations. In long-haul systems, for example, dispersion will accumulate and eventually give rise to the need for a regenerator. In this case, the maximum number of amplifiers will be determined by the spectral purity of the transmitter laser and the dispersion of the fiber.

As we have seen in the preceding analysis, the number of amplifiers is directly proportional to the amplifier input power. The input power, however, can not be made arbitrarily large. When the amplifier output power becomes large, the gain saturates. This restrains the amplifier input power to:

$$P_{\text{out}} = GP_{\text{in}} < P_{\text{sat}}$$

(21)

where $P_{\text{sat}}$ is the saturation output power. Typically, $P_{\text{sat}} = +7$ dBm, hence the maximum input power to a 25-dB gain amplifier is $-18$ dBm. In an amplifier chain were the spontaneous emission adds from stage to stage, the total accumulated spontaneous emission will eventually be sufficiently strong to saturate the amplifiers. If we require that the amplified spontaneous emission from the last amplifier should be less than the saturation output power, the maximum number of amplifiers is limited by:

$$N_{\text{max}} < \frac{P_{\text{sat}}}{(10hvN_{\text{sp}}G)}$$

(22)

For $P_{\text{sat}} = 7$ dBm, $B_{\text{o}} = 300$ A, $N_{\text{sp}} = 1.4$, and $G = 25$ dB, (22) gives $N_{\text{max}} = 20$ which is far more restrictive than that calculated from noise considerations in Fig. 12. Thus, the main impetus for restricting the optical bandwidth in an amplifier chain system is not for noise reduction but rather for preventing the accumulated spontaneous emission from saturating the amplifiers.

Here we will make a rough estimate of the ultimate capacity of a system with multiple amplifiers and multiple channels. As a measure of the capacity we will use the product: $N_{\text{max}} GMB_{\text{o}}$, i.e., the product of the number of amplifiers multiplied by the gain for each amplifier times the number of channels times the bandwidth of each channel. We will assume that the channel spacing is 10 times the bandwidth of each channel. This rather conservative requirement will minimize crosstalk penalties. Thus using (22) with $B_{\text{o}} = 10 MB_{\text{o}}$ from the spontaneous emission overloading restriction we obtain:

$$N_{\text{max}} B_{\text{o}} M G < \frac{P_{\text{sat}}}{(10hvN_{\text{sp}}G)} \sim 2.8 \times 10^{15}.$$  

(23)

If we use the noise limitations on a coherent system in (20) and require that $MP_{\text{in}} G < P_{\text{sat}}$, and that the sensitivity is 60 photons/bit ($P_{\text{sens}} = 60hvB_{\text{o}}/2$), we get

$$N_{\text{max}} B_{\text{o}} M G < \frac{P_{\text{sat}}}{(60hvB_{\text{o}})} \sim 0.5 \times 10^{15}.$$  

(24)
For the signal-spontaneous beat noise limited case, we get from (18)
\[ N_{\text{max}} B_r M \cdot G < \frac{P_{\text{sat}}}{576 h \nu \cdot N_{\text{sp}}} \sim 5 \cdot 10^{13}. \]  
(25)

Finally, using (17) for a direct detection system in the spontaneous-spontaneous beat noise limit and using \( P_{\text{in}} GM = P_{\text{sat}}, \) and \( B_r = 10 B_e \) we get:
\[ N_{\text{max}} B_r M G < \frac{P_{\text{sat}}}{(54 h \nu N_{\text{sp}})} \sim 0.5 \cdot 10^{15}. \]  
(26)
The gain-bandwidth product (GBP) of the amplifier chain is given by:
\[ \text{GBP} = MB_r G^{N_{\text{max}}}. \]  
(27)

Taking the restrictions (23)-(26) into account, the GBP is maximum for \( G = 2.71 \) (= e) [18]. This is an impractically low gain and a multiamplifier system is likely to be suboptimum with respect to the GBP. In Fig. 15 we show the ultimate GBP that can be achieved, plotted versus the gain for each amplifier stage. As can be seen in Fig. 15, the achievable GBP is very large. For a single channel system with 1-GHz bandwidth, the maximum GBP is \( \sim 10^{11} \) Hz for an amplifier gain of 2.72. With a more realistic gain of 25 dB, the GBP is still an impressive \( 10^{10} \) Hz. The main conclusion from this hawing exercise is that the capacity is directly proportional to \( P_{\text{sat}}/N_{\text{sp}}, \) which is the true figure of merit for an optical amplifier.

**EXPERIMENTAL**

**Device Characteristics**

The key device parameters for a traveling wave amplifier are: the facet reflectivity, the spontaneous emission factor, and the saturation output power. Below we will give experimental data on these parameters on amplifiers made in our laboratory. Although data is presented only for a 1.5-\( \mu \)m amplifier, virtually identical results have been obtained for 1.3-\( \mu \)m devices.

The amplifiers were fabricated from channel substrate buried heterostructure lasers having cavity lengths of 500 \( \mu \)m. The long cavity length, about twice that of a laser, reduces the wavelength shift of the gain peak after AR coating and increases the maximum gain that can be achieved. The AR coatings were applied by thermal evaporation of SiO in an oxygen atmosphere. The partial pressure of the oxygen was adjusted to give the optimum index of refraction of the coatings [19]. The film thickness was controlled by in situ monitoring of the light output from the laser being coated [20]. Before AR coating, the devices had typical threshold currents of 17 mA. The spontaneous emission spectrum (TE) at 100-mA drive current is shown in Fig. 16. At 1.51-\( \mu \)m wavelength, the peak gain is 29.8 dB. As can be seen in Fig. 16, the reflectivity, as judged from the Fabry–Perot ripple in the spontaneous emission, has a minimum around 1.54 \( \mu \)m where the ripple vanishes. The actual facet reflectivity was obtained by measuring the resonant and antiresonant gain across the gain curve. From this measurement the geometric mean of the two facet reflectivities \( R = \sqrt{R_1 R_2} \) can be obtained as:
\[ R = \frac{G_{\text{max}} - G_{\text{min}}}{4G_{\text{max}} G_{\text{min}}} \]  
(28)

where \( G_{\text{max}} \) and \( G_{\text{min}} \) are the resonant and antiresonant gain, respectively. The result is shown in Fig. 17. At the reflectivity minimum at 1.54 \( \mu \)m, the gain ripple is less than the experimental uncertainty 0.2 dB and the reflectivity is less than \( 3 \cdot 10^{-5} \). This is about one order of magnitude less than the best previously published results for the average reflectivity [5]. The spectral region over which the reflectivity is less than 1 \( \cdot 10^{-4} \) is about 300 A.

Next, the saturation output power was obtained from a measurement of the gain versus amplifier output power. At 1.52-\( \mu \)m wavelength and a gain of 31 dB, the 3-dB saturation output power is +6.5 dBm in the close agreement with previously published results [3], [5].

The spontaneous emission coefficient depends on the relative position of the operating wavelength with respect to the gain peak [6]. Here, the spontaneous emission factor was measured using a heterodyne method in which the amplifier output is heterodyned with a local oscillator [21]. The result of this measurement is shown in Fig. 18 showing the measured \( N_{\text{sp}} \) as function of wavelength at a drive current of 80 mA. At this current, the peak gain at 1.52 \( \mu \)m was 28 dB. The spontaneous emission factor decreases with increasing wavelength and reaches a minimum value of 1.4 at a wavelength 300 A shorter than the gain peak. At 1.51 \( \mu \)m, \( N_{\text{sp}} \) is almost 2 times higher or 2.5. This measurement show that, for noise considerations, the longer wavelength side of the amplifier spec-
The two main applications for optical amplifiers in lightwave communication systems are as in-line amplifiers and as receiver pre-amplifiers. In in-line applications, the amplifier operates as a simple repeater providing gain to increase the allowable loss between the transmitter and receiver. When operating as a preamplifier, the optical amplifier boosts the optical signal immediately before the photo detector thereby increasing the receiver sensitivity.

**4-Gbit/s Optical Preamplifier**

At 1.5-μm wavelength, optical preamplifiers have been demonstrated both at 500 Mbit/s [8], 2 Gbit/s [7], and up to 4 Gbit/s [22]. Noise measurements have also been made on a 10-GHz receiver [9]. Optical preamplifiers are at advantage mainly at very high data-rates where the gain bandwidth product of APD detectors is a limiting factor. Therefore, the preamplifier was evaluated at the fairly high data-rate of 4 Gbit/s to demonstrate its potential advantage over APD receivers.

The experimental setup is shown in Fig. 19. The amplifier chip had the reflectivity and noise characteristics as shown in Figs. 17 and 18. Optical isolators were inserted both before and after the amplifier for more stable operation and a grating provided optical filtering. The input and output coupling efficiencies were 0.31 and 0.26, respectively. The net fiber-to-fiber gain was 14.2 dB at the 1.55-μm operating wavelength and the optical filter bandwidth was 10 Å. The source laser was a DFB laser directly modulated with a 2^{15} - 1 NRZ bit stream at 4 Gbit/s. The receiver had a p-i-n detector and a high impedance (5 kΩ) front end and the sensitivity without the optical pre-amplifier was -25 dBm at 4 Gbit/s [23].

The BER characteristics for the receiver with the optical pre-amplifier is shown in Fig. 20. The received power in Fig. 20 refers to the power in the input fiber to the preamplifier. With the optical preamplifier, the receiver sensitivity is -34.3 dBm which is about a factor of two better than the best published APD receiver results at the same data-rate [24]. It is also interesting to note that the best (and only) published coherent receiver sensitivity at 4 Gbit/s is -31.3 dBm, again substantially worse than the sensitivity of the optical preamplifier receiver presented here [25].

The optical preamplifier was operated at a wavelength 400 Å longer than the gain maximum. As discussed above, this is advantageous because of the substantially lower noise figure at longer wavelengths. For this particular amplifier, the noise figure was about 3 dB lower at the operating point at 1.55 μm than at the 1.51-μm gain maximum.

Using the known device and system parameters, the calculated receiver sensitivity is -35.2 dBm (see Fig. 2). The discrepancy between theory and experiments is believed to be due to intersymbol interference (ISI) from both the modulation characteristics of the laser and from
Fig. 21. Experimental setup for long-haul transmission experiment using four in-line amplifiers. SCBR = silicon chip Bragg reflector transmitter laser, ISO = optical isolator, PC = polarization controller, ECL = external cavity laser, AMP = electrical amplifier, DBX = double balanced mixer, and LPF = low pass filter.

the nonperfect frequency response of the receiver electronics. ISI effects, which are increasingly important at higher data-rates, have not been accounted for in the theory.

In-Line Amplifiers with Coherent Detection

As an example of an in-line amplifier application, we will describe a recent coherent long haul system experiment [13]. In this experiment, four optical amplifiers were used and spaced about 65 km in the transmission path. The detailed experimental setup is shown in Fig. 21. The system operated at 400 Mbit/s and used coherent detection with a frequency shift keying modulation format. Optical isolators were inserted in between all amplifiers. These isolators eliminate cross-interaction between the amplifiers and improve the stability of the system. Both the isolators and the optical amplifiers are sensitive to the input polarization of the light; and the polarization was adjusted before each amplifier stage by means of manual fiber polarization controllers.

The amplifiers used in this experiment were of the near traveling wave type with facet reflectivities of the order $5 \cdot 10^{-3}$. The total gain of the four amplifiers was 500,000,000 or 87 dB. However, because of coupling losses, the net useful gain as 50,000 or 47 dB. Despite the heavy coupling losses, the amplifiers provided enough net gain to compensate for the loss in 200 km of fiber, extending the total transmission distance to 372 km, the longest nonregenerated transmission distance for any fiber communication system. The power budget for this system experiment is given in Table III. Fig. 22 shows the variation of the optical signal power along the fiber path. In between the amplifiers, the signal power drops exponentially with distance (linear in decibels) with a slope given by the fiber loss. At each amplifier site, the power is boosted an amount equal to the net gain of the amplifier. A baseline system performance was established by measuring the BER characteristics with only a short length of fiber and no optical amplifiers. This measurement is shown in Fig. 23 as curve $A$, and gives a receiver sensitivity of $-50.0 \text{ dBm}$. With the full length of fiber (372 km) and with four amplifiers, the receiver sensitivity is $-48.5$ dBm as shown by curve in Fig. 23. Hence, the penalty is only 1.5 dB and the BER decreases monotonically with increased power without any evidence of an error floor.

Although we used coherent detection in this experiment, direct detection can also be used. For example, the same four amplifiers described above were successfully used in a 313-km 1-Gbit/s direct detection system [14].

CONCLUSIONS

We are now at the point where systems with optical amplifiers clearly out-perform traditional systems in terms
of transmission distance and allowable line loss. For example, previously, the longest transmission distance for direct detection around 1 Gbit/s was 171 km, substantially shorter than the 313 km discussed above. And, as we have seen, at high data-rates, receivers with optical pre-amplifiers are considerably more sensitive than both coherent and APD receivers. Although the outlook for optical amplifiers seems bright, it is not without problems. One of the major tasks in the near future is to design and fabricate an optical amplifier that is polarization independent. Today's amplifiers have several decibels difference in gain for TE and TM polarized light. Both the facet reflectivity and internal gain is polarization dependent. Progress has been made in addressing these problems, but no truly polarization independent amplifier has been reported yet. A second potential problem is associated with possible crosstalk in amplifiers. A major advantage of an optical amplifier is that it is a multichannel device; i.e., an amplifier can simultaneously and independently amplify several channels in the same device. However, the simultaneous presence of several channels gives rise to the possibility of crosstalk between the channels. Of particular concern is crosstalk caused by gain saturation [26]-[28] and four-wave mixing [29]-[31]. These effects will not prevent multichannel operation; however, they may put additional restraints on the channel spacing and maximum output power from the amplifiers. Additional research is required to clarify these concerns.

**Appendix A**

**Derivation of the Signal-Spontaneous and Spontaneous-Spontaneous Beat Noise Terms at the Output of an Optical Amplifier**

This derivation of the amplifier noise is based on a talk given by P. S. Henry, who pointed out some inconsistencies in the literature and derived the correct results.

For simplicity, we assume an optical amplifier with unity coupling efficiency, uniform gain $G$, over an optical bandwidth $B_o$, and an input power of $P_{in}$ at optical frequency $w$, centered in the optical passband $B_o$. The spontaneous emission power in the optical bandwidth $B_o$ is given by:

$$P_{sp} = N_{sp}(G-1)hvB_o.$$  \hfill (A1)

Writing the electric field $E_{sp}$, representing the spontaneous emission as a sum of cosine terms:

$$E_{sp}(t) = \sum_{k = -B_o/(2\delta v)}^{B_o/(2\delta v)} \sqrt{2N_{sp}(G-1)hv\delta v} \cdot \cos ((\omega_0 + 2\pi k\delta v)t + \phi_k)$$  \hfill (A2)

where $\phi_k$ is a random phase for each component of spontaneous emission. With

$$N_{sp}(G-1)hv = N_o$$  \hfill (A3)

and

$$\frac{B_o}{2\delta v} = M$$

de the total electric field at the output of the amplifier is

$$E(t) = \sqrt{2GP_{in}} \cos (\omega_0 t + \sum_{k = -M}^{M} \sqrt{2N_{sp}\delta v} \cdot \cos ((\omega_0 + 2\pi k\delta v)t + \phi_k).}$$  \hfill (A4)

The photocurrent $i(t)$ generated by a unity quantum efficiency photodetector is proportional to the intensity:

$$i(t) = E^2(t) \frac{e}{hv}$$  \hfill (A5)

where the bar indicates time averaging over optical frequencies. Hence

$$i(t) = GP_{in} \frac{e}{hv} + \frac{4e}{hv} \sum_{k = -M}^{M} \sqrt{GP_{in}N_{sp}\delta v} \cdot \cos (\omega_0 t + 2\pi k\delta v + \phi_k) + \frac{2\pi N_{sp}\delta v}{hv} \cdot \left[ \sum_{k = -M}^{M} \cos ((\omega_0 + 2\pi k\delta v) + \phi_k) \right]^2.$$  \hfill (A6)

The three terms in (A6) represent, signal, signal-spontaneous beat noise, and spontaneous-spontaneous beat noise, respectively.

**Signal-Spontaneous Beat Noise**

Examining the signal-spontaneous beat noise term first:

$$i_{sp}(t) = \frac{4e}{hv} \sum_{k = -M}^{M} \sqrt{GP_{in}N_{sp}\delta v} \cdot \cos (\omega_0 t + 2\pi k\delta v + \phi_k)$$

$$= \frac{2e}{hv} \sqrt{GP_{in}N_{sp}\delta v} \sum_{k = -M}^{M} \cos (2\pi k\delta v + \psi_k)$$  \hfill (A7)

where terms $-\cos (2\omega_0 t)$ which average to zero have been neglected. For each frequency, $2\pi k\delta v$, in (A7), the sum has two components but with a random phase. Therefore, the power spectrum of $i_{sp}(t)$ is uniform in the frequency interval $0 - B_o/2$ and has a density of:

$$N_{sp} \frac{4e^2}{hv^2} \cdot \frac{GP_{in}N_o}{2} \cdot \frac{1}{2} \cdot \frac{1}{2}$$

$$= \frac{4e^2}{hv} P_{in}N_{sp}(G-1)G.$$  \hfill (A8)

**Spontaneous-Spontaneous Beat Noise**

From (A6) we have

$$i_{sp}(t) = \frac{2N_o\delta v e}{hv} \sum_{k = -M}^{M} \cos (\omega_0 + 2\pi k\delta v) + \phi_k)^2$$

$$= 2N_o \frac{\delta v e}{hv} \sum_{k = -M}^{M} \cos (\beta_k) \sum_{j = -M}^{M} \cos (\beta_j)$$  \hfill (A9)
The terms \( \cos (\beta_k + \beta_j) \) have frequencies \( \sim 2\omega_o \) and average to zero. Rewriting (A11) gives:

\[
I_{sp-sp}(t) = \frac{N_o \delta e \nu}{\hbar \nu} \sum_{k=-M}^{M} \sum_{j=-M}^{M} \cos (k-j)2\pi \delta \nu t + \Phi_k + \Phi_j. \tag{A12}
\]

The dc term is obtained for \( k = j \) and there are \( 2M \) such terms:

\[
I_{sp-sp}^d = \frac{e}{\hbar \nu} N_o \delta e \nu 2M = N_{sp}(G-1)\nu B_o. \tag{A13}
\]

Organizing the terms in (A12) according to their frequencies we get:

<table>
<thead>
<tr>
<th>frequency</th>
<th># terms</th>
</tr>
</thead>
<tbody>
<tr>
<td>(-2M \delta \nu)</td>
<td>1</td>
</tr>
<tr>
<td>(-2 \delta \nu)</td>
<td>2M-l</td>
</tr>
<tr>
<td>(-2 \delta \nu)</td>
<td>2M-1</td>
</tr>
<tr>
<td>(2 \delta \nu)</td>
<td>2M-1</td>
</tr>
<tr>
<td>(2 \delta \nu)</td>
<td>2M-l</td>
</tr>
<tr>
<td>((2M-1) \delta \nu)</td>
<td>1</td>
</tr>
</tbody>
</table>

The terms with same absolute frequency but of opposite sign add in phase. Therefore, the power spectrum of the spontaneous-spontaneous beat noise extends from 0 to \( B_o \) with a triangular shape and a power density near dc of

\[
N_{sp-sp} = \frac{2N_0^2 \delta e \nu}{\hbar \nu} \left( \frac{B_o}{\delta \nu} - 1 \right) \cdot \frac{1}{\nu} = 2N_0^2(G-1)\nu^2 B_o. \tag{A14}
\]

In Fig. 24 we summarize the results of (A8), (A13), and (A14) and show the electrical power spectrum of the beat noise.

**ACKNOWLEDGMENT**

It is the author’s pleasure to thank P. S. Henry for several illuminating discussions of optical amplifier noise, and especially for the material in Appendix A.

**REFERENCES**


* N. Anders Olsson was born in Soiffeke, Sweden, on April 13, 1952. He received the "Civilingenjorsexamen" degree in engineering physics from Chalmers University of Technology, Gothenburg, Sweden, in 1975, and the M.Eng. and Ph.D. degree in electrical engineering from Cornell University, Ithaca, NY, in 1979 and 1982, respectively.

Between 1975 and 1977 he was with Schlumberger Overseas SA, Singapore, as Field Engineer assigned to the Far East Asian region. After six months as a Post-Doctoral Research Associate at Cornell University, he joined Bell Laboratories in 1982. His research interests include semiconductor lasers for optical communication, especially single-frequency sources and optical amplifiers.