Semiconductor optical amplifiers in WDM star networks

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Abstract: The placement of optical amplifiers in WDM star fibre-optic networks has been considered. Both postamplifier and preamplifier configurations have been investigated. The number of users supported by the topology has been obtained when reflection noise and amplifier gain saturation (cross-saturation) are present. The effect of extinction ratio has also been considered. In general the analysis shows that the preamplifier arrangement is better than that of the postamplifier.

1 Introduction

The high information carrying capacity of the optical fibre can be utilised using one of the following three techniques: wavelength division multiple access (WDMA), optical time division multiple access (TDMA) and optical code division muitiple access (CDMA). The WDMA technique has the advantage that its implementation is relatively easier as compared with other two. To implement the WDMA in local access network, many physical topologies such as star, bus, folded bus, ring, etc. can be used.

Star is the most suitable topology as it distributes the transmitted power equally among the receivers. The number of users in the star is limited by the power budget and can be increased by using (i) higher transmitter power level, (ii) higher receiver sensitivity, (iii) components having lower losses, (iv) electronic regenerator (i.e. active hub) and (v) optical amplifiers (OAs). Options (i), (ii) and (iii) will demand considerable technological effort to improve the devices and components. Option (iv) will require wavelength division multiplexers and demultiplexers; multiple detectors; electronic amplifiers and sources. Option (v) would be the preferred choice as only one amplifier can be used for the amplification of multiple channels in the fibre. The optimum use of OAs in the star topology depends on their placement. In this paper, performance of star network in terms of number of users supported has been investigated for postamplifier and preamplifier configurations.

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2 System description

In an $N \times N$ star network every user is provided with a transmitter and receiver pair. In the simplest arrangement [1], each user is assigned a dedicated transmit wavelength. The receiver takes one channel at a time using a wavelength-selective filter, which can be fixed or tunable depending on the media access control (MAC) protocol. In this topology, optical amplifiers can be placed either after the transmitter as postamplifiers or before the receiver as preamplifiers. It will increase the number of users or decrease the required transmitter power for given requirements. However, the presence of noise and the gain saturation effect of the amplifiers will partially reduce these advantages.

In a multichannel transmission system, the total light intensity varies randomly since in each channel, light is independently modulated. Thus, the signal output of one channel varies according to the gain fluctuation induced by modulation of other channels, even when the input in the channel is constant. This is crosstalk induced by gain saturation in the amplifier and is often referred as cross-saturation. The cross-saturation effect is maximum when all N wavelengths are used in the system. So this gives the upper bound on the degradation.

In general, implementation of a $N \times N$ star coupler will require $\log_2 N$ stages, each containing $(N/2)$ number of 2×2 couplers. The insertion loss L_{TI} of such a coupler is given by

$$L_{TI} = L_i \log_2 N \text{dB} \tag{1}$$

where L_i is the insertion loss of a single 2×2 coupler.

In this paper, operating wavelength is considered to be in the 1.5μm region and the signalling scheme to be binary intensity modulation. Semiconductor laser amplifiers (SLAs) are commonly available in this wavelength region, which can be integrated easily with the integrated optic implementation of transmitters and receivers. Therefore SLAs have been considered in the following analysis.

3 Amplifier model

An SLA can be represented by a travelling wave amplifier (TWA) model if $G\sqrt{[R_1 R_2]} < 0.17$ [2], where R_1 and R_2 are the reflection coefficients of two facets of the amplifier, and G the single-pass saturated gain. For a TWA, the gain is given by [2]

$$G = G_0 \exp\left(-[G-1]\frac{P_{in}}{P_{sat}}\right) \tag{2}$$

where G_0 is the unsaturated gain, and P_{in} and P_{sat} the input and saturation power levels, respectively. Parameters G_0 and P_{sat} depend on amplifier and biasing conditions. This equation is derived from a steady-state analysis and ignores dynamic effects, which may be important for high bit rates and possible four-wave mixing (FWM). To simplify the analysis, the dynamic effects have not been considered.

3.1 Reflection from amplifier

The path difference between transmissions and reflections from the amplifier cavity is less than the coherence length of the laser light. Therefore multiple reflections and transmissions can be added coherently. Under this condition, reflection coefficient of the amplifier is given by

$$R_{amp} = \frac{[G\sqrt{R_2} - \sqrt{R_1}]^2}{[1 - G\sqrt{R_1 R_2}]^2} \tag{3}$$

3.2 Amplifier noise model

The OA generates noise owing to amplified spontaneous emission (ASE). It is considered to be white with a single-sided power spectral density (PSD) [3, 4]

$$S_{sp} = n_{sp}[G - 1]h\nu \tag{4}$$

Here, n_{sp} is the spontaneous emission factor which lies between 1.4 and 4.0 for SLAs [3], h is Planck's constant and ν the optical frequency. At the photodetector, the ASE noise beats with itself and other optical signals, i.e. information signal, echo signal. The variances of different beat noise current components are:

ASE-ASE beat noise [3, 4]:

$$\sigma^2_{ASE-ASE} = R_0^2 S_{sp}^2 [2B_e B_o - B_e^2] \tag{5}$$

ASE-signal beat noise [3, 4]:

$$\sigma^2_{ASE-sig} = 4R_0^2 P_R S_{sp} B_e \tag{6}$$

In the presence of an echo signal along with ASE, an additional ASE-echo beat noise component is produced. It is determined by replacing the received power level P_R with the echo signal power P_{echo} in eqn. 6. In eqns. 5 and 6, R_0 is the responsivity of photodetector, B_e the electrical bandwidth of receiver and B_o the optical bandwidth. ASE-ASE beating also produces a DC component which gives rise to shot noise. The shot noise variance is [3, 4]

$$\sigma^2_{ASE-shot} = 2eR_0[S_{sp}B_o]B_e \tag{7}$$

where e is the electron charge.

4 Computation of minimum required transmitter power

Once the relationship between bit error rate (BER) and transmitter power P_T is known for a given N, minimum required P_T can be determined using this relationship. Therefore models to determine BER as a function of P_T and N are formulated as follows, when bits 1 and 0 occur with equal probability.

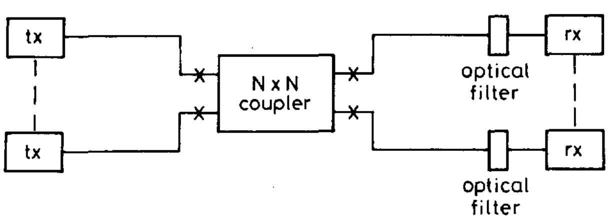

Fig. 1 *Star network without amplifier*

5 Star without amplifier

In the star without amplifiers (Fig. 1), power received at photodetector for bit i (i is either 1 or 0) is given by

$$P_R(i) = \frac{P_s(i)L_{TR}}{N} \tag{8}$$

where:

$$P_s(1) = \frac{P_T}{\epsilon + 1} \tag{9a}$$

$$P_s(0) = \frac{P_T\epsilon}{\epsilon + 1} \tag{9b}$$

In these equations, $P_T/2$ represents the average transmitter power and ε the extinction ratio. The loss L_{TR} between transmitter and receiver is given by

$$L_{TR} = 10^{-\left(\frac{\alpha 2L + L_{TI} + L_{FI} + L_{CV} + 3L_{sp}}{10}\right)} \tag{10}$$

This does not include the split loss of star coupler which has been accounted by the factor $1/N$ in eqn. 8. In the above equation, α is the attenuation coefficient of fibre in dB/km, L the length of fibre from user to star coupler in km, L_{FI} represents the filter insertion loss, L_{CV} the loss due to nonuniformity in the power splitting by the star coupler, and L_{sp} the splice loss in decibels. The signal current and noise variance for bit i (i = 0 or 1) at photodetector output are given by:

$$I_{sig}(i) = R_0 P_R(i) \tag{11a}$$

$$\sigma^2(i) = 2eR_0 P_R(i) B_e + \frac{4kTB_e}{R_L} \tag{11b}$$

where k is the Boltzman constant, T the temperature in Kelvin, and R_L the load resistance of the photodetector. The first term on the right-hand side of eqn. 11b represents shot noise and the second term thermal noise. The average probability of error considering the threshold level which equalises the BER for bit 1 and 0 is

$$P_e = \frac{1}{2} erfc\left(\frac{I_{sig}(1) - I_{sig}(0)}{\sqrt{2}[\sigma(0) + \sigma(1)]}\right) \tag{12}$$

6 Star with postamplifier

In this Section the performance of a star network with an ideal postamplifier is analysed. Subsequently, the effect of amplifier gain saturation and reflection is also considered (Fig. 2).

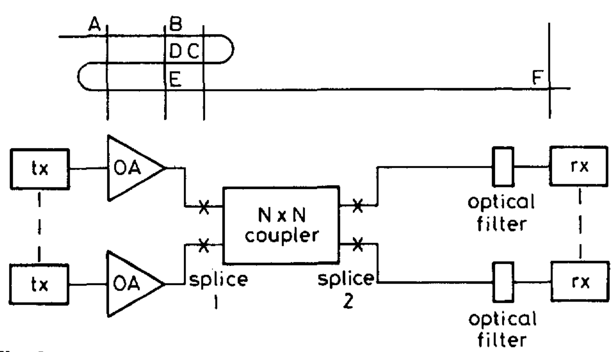

Fig. 2 *Star network with postamplifiers*
A, B, ... F is echo path

6.1 Ideal postamplifier

Let the gain of the ideal postamplifier be G_0; then

$$P_R(i) = \frac{P_S(i)G_0 L_{TR}}{N} \tag{13}$$

The ASE noise PSD at the receiver considering N amplifiers and $N \times N$ star coupler will be

$$S_{sp} = n_{sp}[G_0 - 1]h\nu L_{TR} \tag{14}$$

The signal current and noise variance for bit i using eqns. 5–7 are given by:

$$I_{sig}(i) = R_0 P_R(i) \tag{15a}$$

$$\sigma^2(i) = 2eR_0[P_R(i) + S_{sp}B_o]B_e + 4R_0^2 P_R(i) S_{sp} B_e + R_0^2 S_{sp}^2[2B_oB_e - B_e^2] + \frac{4kTB_e}{R_L} \tag{15b}$$

Eqns. 12 and 15 are used to determine average P_e.

6.2 Effect of gain saturation

Let the amplifier gain corresponding to bit 1 and 0 be $G(1)$ and $G(0)$, respectively. Therefore the received signal power level for bit i will be

$$P_R(i) = \frac{P_s(i)G(i)L_{TR}}{N} \tag{16}$$

The received ASE noise PSD is different for bit 1 and 0. When the desired channel has bit 1, one half of the remaining (N–1) channels are expected to have bit 0 and other half, bit 1. The ASE noise from each amplifier gets distributed equally among the outputs. Therefore at any output ASE noise PSD from each amplifier will be reduced by a factor of $1/N$. The total ASE noise PSD at the receiver for bit i is

$$S_{sp}(i) = \frac{n_{sp}[G(i)-1]h\nu L_{TR}}{N} + \frac{[N-1]}{2}\frac{n_{sp}[G(0)-1]h\nu L_{TR}}{N} + \frac{[N-1]}{2}\frac{n_{sp}[G(1)-1]h\nu L_{TR}}{N} \tag{17}$$

The first term corresponds to the desired channel, the second term to channels having bit 0 and the third term to channels having bit 1. The signal current and noise variance for bit i are given by:

$$I_{sig}(i) = R_0 P_R(i) \tag{18a}$$

$$\sigma^2(i) = 2eR_0[P_R(i) + S_{sp}(i)B_o]B_e + 4R_0^2 P_R(i) S_{sp}(i) B_e + R_0^2 S_{sp}^2(i)[2B_oB_e - B_e^2] + \frac{4kTB_e}{R_L} \tag{18b}$$

The average P_e in this case can be determined using these equations and eqn. 12.

6.3 Effect of reflection noise

When the splices in the network are not good, back reflection results in echoes. In general a signal which has undergone $2n$ reflections will generate n-pass echo. Generally the reflection coefficient of splices is below –20dB; echoes with two and more passes can be neglected.

When delays suffered by echoes are greater than the coherence time of the optical signal, coherence between signal and echoes need not be considered. In the network under consideration, distance between the reflection points is much greater than coherence length. Hence, incoherent addition of signal and echo power has been considered.

In the network shown in Fig. 2, the first reflection point is the amplifier and the second can be the splice before the coupler i.e. splice 1 or the splice after the coupler i.e. splice 2. The loss between OA and splice 2 (insertion loss of star coupler + splice loss + fibre loss) is much higher than the loss between OA and splice 1. Therefore the strength of echo between OA and splice 2 will be much lower than the strength of echo between OA and splice 1. Hence, echo signal arising due to splice 2 can be neglected.

For simplicity, let the bits in the desired signal and the echo signal be synchronised. This corresponds to a worst-case situation and the analysis will give an upper bound of degradation owing to echo. Let the echo of bit b_r interfere with signal corresponding to bit b_c. The received echo power considering the echo path A, B, ..., F (Fig. 2) is given by

$$P_{echo} = \frac{P_s(b_r)G_p(b_r)L_{ASI}^2 R_{sp}R_{amp}L_{TR}}{N} \tag{19}$$

where L_{ASI} is the fibre loss between amplifier and splice 1, R_{sp} the splice reflection coefficient. The parameter $G_p(b_r)$ represents the amplifier gain when the signal corresponding to bit b_r passed through the amplifier:

$$G_p(b_r) = G_0 \exp\left(-[G_p(b_r) - 1] \times \frac{[P_s(b_r)[1-R] + P_s(b_{rr})G_p' L_{ASI}^2 R_{sp}[1-R]^2]}{P_{sat}}\right) \tag{20}$$

where R is the facet reflectivity of the amplifier ($R = R_1 = R_2$). In this equation, it is considered that the echo of bit b_{rr} interfered with the signal corresponding to bit b_r. The parameter G_p' is the gain when the signal corresponding to bit b_{rr} was amplified. To simplify the analysis, it is assumed that no echo was present when the signal corresponding to b_{rr} was amplified. Further, b_{rr} is assumed to be zero for maximum echo signal. In the equation, the signal corresponding to b_r and echo of b_{rr} are added together and used as P_{in} in eqn. 2. As such, the two optical signal powers cannot be added as these are entering the amplifier from opposite ends. But the addition corresponds to the worst-case scenario.

The amplifier gain G for the signal corresponding to bit b_c under the same assumption is

$$G = G_0 \exp\left(-[G-1] \times \frac{P_s(b_c)[1-R] + P_s(b_r)G_p L_{ASI}^2 R_{sp}[1-R]}{P_{sat}}\right) \tag{21}$$

It is seen from eqns. 3 and 21, R_{amp} is function of G, which in turn depends on b_c and b_r. The received signal power at the photodetector input will be

$$P_{sig}(b_c, b_r) = \frac{P_s(b_c)G(b_c,b_r)[1-R_{sp}]L_{TR}}{N} \tag{22}$$

The ASE noise PSD at the photodetector is given by

$$\begin{aligned} S_{sp}(b_c, b_r) &= n_{sp}[G(b_c,b_r) - 1]h\nu[1-R_{sp}]\frac{L_{TR}}{N} \\ &+ n_{sp}[G(b_r,b_{rr}) - 1]h\nu L_{ASI}^2 R_{sp} R_{amp}(b_c,b_r)\frac{L_{TR}}{N} \\ &+ \sum_{i=1}^{N-1} n_{sp}[G(b_{ci},b_{ri}) - 1]h\nu[1-R_{sp}]\frac{L_{TR}}{N} \\ &+ \sum_{i=1}^{N-1} n_{sp}[G(b_{ri},b_{rri}) - 1]h\nu L_{ASI}^2 R_{sp} R_{amp}(b_{ci},b_{ri})\frac{L_{TR}}{N} \end{aligned} \tag{23}$$

In this equation, the first term corresponds to bit b_c in the desired channel, the second term to bit b_r in the desired channel, the third term to signals from the remaining N–1 amplifiers, and the last term to echoes from the remaining N–1 amplifiers. Further, bits b_{ci}, b_{ri} and b_{rri} are bits in the ith channel corresponding to bits b_c, b_r and b_{rr}, respectively, in the desired channel. In the worst case situation, bits b_{rr} and b_{rri} can be taken as 0. The b_{ci} and b_{ri} can be either 1 or 0 with equal probability. When P_{sig}, P_{echo} and S_{sp} are known the signal current and noise variance are given by:

$$I_{sig}(b_c, b_r) = R_0 P_R \tag{24a}$$

$$\begin{aligned}\sigma^2(b_c, b_r) &= 2e[I_{sig}(b_c, b_r) + R_0 S_{sp}(b_c, b_r) B_o] B_e \\ &+ 4R_0^2 P_R(b_c, b_r) S_{sp}(b_c, b_r) B_e \\ &+ R_0^2 S_{sp}(b_c, b_r)^2 [2B_0 B_e - B_e^2] + 2R_0^2 P_{sig} P_{echo} \\ &+ \frac{4kTB_e}{R_L}\end{aligned} \tag{24b}$$

where

$$P_R = P_{sig}(b_c, b_r) + P_{echo}$$

The fourth term on the right-hand side of eqn. 24b represents signal-echo beat noise. It is evident from eqn. 24a that there are two signal levels corresponding to bit 1. The same is true for bit 0 also. The threshold is obtained using the highest level corresponding to bit 0 and the lowest level corresponding to bit 1. Using these levels and corresponding noise variances, the threshold that equalises the BERs for bit 1 and 0 is given by

$$D_{th} = \frac{\sigma(1,0) I_{sig}(0,1) + \sigma(0,1) I_{sig}(1,0)}{\sigma(1,0) + \sigma(0,1)} \tag{25}$$

With this threshold, the probabilities of error for bit 1 and 0 are given by:

$$P_e(1, b_r) = \frac{1}{2} erfc\left(\frac{I_{sig}(1, b_r) - D_{th}}{\sqrt{2}\sigma(1, b_r)}\right) \tag{26a}$$

$$P_e(0, b_r) = \frac{1}{2} erfc\left(\frac{D_{th} - I_{sig}(0, b_r)}{\sqrt{2}\sigma(0, b_r)}\right) \tag{26b}$$

Since bit 1 and 0 are equiprobable, average probability of error is given by

$$P_e = \sum_{b_r=0}^{1} \frac{1}{2}\left[\frac{1}{2} P_e(1, b_r) + \frac{1}{2} P_e(0, b_r)\right] \tag{27}$$

7 Star network with preamplifier

In this arrangement, all the wavelengths are amplified by each preamplifier and one of them is selected by the filter. As in the case of the postamplifier, performance of star network with ideal preamplifiers is analysed. This analysis is extended to include the effect of amplifier gain saturation and reflection.

7.1 Ideal amplifier

The received power for bit i is

$$P_R(i) = \frac{P_s(i) L_{TA} G_0 L_{AR}}{N} \tag{28}$$

where

$$L_{TA} = 10^{-\left(\frac{2\alpha L + L_{TI} + L_{cv} + 3L_{sp}}{10}\right)} \tag{29}$$

is the loss between transmitter and amplifier without the split loss of coupler. The L_{AR} represents the loss between amplifier and receiver which consists of only the insertion loss of filter.

The ASE noise PSD S_{sp} at the photodetector is determined from eqn. 14 by replacing L_{TR} with L_{AR}. It will be the same for bit 1 and bit 0. Substituting $P_R(i)$ and S_{sp} in eqn. 15, the signal currents ($I_{sig}(1)$ and $I_{sig}(0)$) and noise variances ($\sigma^2(1)$ and $\sigma^2(0)$) for bit 1 and 0 are obtained, and the average P_e is computed using eqn. 12.

7.2 Effect of gain saturation

The degradation due to cross-saturation arises owing to (i) decrease in the average gain because of increase in the number of channels and (ii) gain fluctuations due to randomness in the number of channels having bit 1. In this section, degradation due to average gain reduction and gain fluctuations is analysed.

7.2.1 Average gain reduction: To determine the average gain, the probability distribution of number of channels having bit 1 is considered to be binomial. Let N_1 channels out of total N channels are on. The probability that N_1 channels are having bit 1 is [8]

$$P_{NI} = \binom{N}{N_1}\left[\frac{1}{2}\right]^N \tag{30}$$

The input power to the amplifier corresponding to bit i is

$$P_{in}(i) = \frac{P_s(i) L_{TA}}{N} \tag{31}$$

Therefore the total input power to the amplifier is $N_1 P_{in}(1)$ + (N–N_1)$P_{in}(0)$. The corresponding saturated gain of the amplifier is

$$G(N_1) = G_0 \exp\left(-[G(N_1) - 1] \times \frac{N_1 P_{in}(1) + (N - N_1) P_{in}(0)}{P_{sat}}\right) \tag{32}$$

Using the probability defined by eqn. 30, the average gain is

$$G_{av} = \sum_{N_1=0}^{N} P_{NI} G(N_1) \tag{33}$$

The signal power received for bit i will be

$$P_R(i) = P_{in}(i) G_{av} L_{AR} \tag{34}$$

As G_{av} is independent of the signal bit, the ASE noise PSD for both bit 1 and bit 0 will be same. It is given by

$$S_{sp} = n_{sp}[G_{av} - 1] h\nu L_{AR} \tag{35}$$

Eqns. 34 and 35 are used to determine signal currents ($I_{sig}(1)$ and $I_{sig}(0)$) and noise variances ($\sigma^2(1)$ and $\sigma^2(0)$) in conjunction with eqn. 15. The average P_e is computed using eqn. 12.

7.2.2 Gain fluctuations: The probability of N_1 channels out of N–1 interfering channels having bit 1 is given by

$$P'_{NI} = \binom{N-1}{N_1}\left[\frac{1}{2}\right]^{N-1} \tag{36}$$

The corresponding total amplifier input power for the bit i in the desired channels is $P_{in}(i)$ + $N_1 P_{in}(1)$ + (N – 1 – N_1)$P_{in}(0)$. Therefore the saturated gain of the amplifier is given by

$$G(i, N_1) = G_0 \exp\left(-[G(i, N_1) - 1] \times \frac{[P_{in}(i) + N_1 P_{in}(1) + (N - 1 - N_1) P_{in}(0)]}{P_{sat}}\right) \tag{37}$$

For bit i, the signal power and ASE noise PSD at the photodetector are:

$$P_R(i, N_1) = P_{in}(i)G(i, N_1)L_{AR} \tag{38a}$$

$$S_{sp}(i, N_1) = n_{sp}(i)[G(i, N_1) - 1]h\nu L_{AR} \tag{38b}$$

The signal current and noise variance corresponding to bit i are:

$$I_{sig}(i, N_1) = R_0 P_R(i, N_1) \tag{39a}$$

$$\begin{aligned}\sigma^2(i, N_1) &= 2eR_0[P_R(i, N_1) + S_{sp}(i, N_1)B_o]B_e \\ &\quad + 4R_0^2 P_R(i, N_1)S_{sp}(i, N_1)B_e \\ &\quad + R_0^2[S_{sp}(i, N_1)]^2[2B_oB_e - B_e^2] + \frac{4kTB_e}{R_L}\end{aligned} \tag{39b}$$

Both $I_{sig}(1)$ and $I_{sig}(0)$ will have N levels depending on N_1. The threshold corresponding to the highest level of $I_{sig}(0)$ and the lowest level of $I_{sig}(1)$ will be

$$D_{th} = \frac{\sigma(0,0)I_{sig}(1, N-1) + \sigma(1, N-1)I_{sig}(0,0)}{\sigma(0,0) + \sigma(1, N-1)} \tag{40}$$

The probability of error under the condition that N_1 interfering channels are at bit 1 is

$$\begin{aligned}P_e(N_1) = \frac{1}{2}\Bigg[&\frac{1}{2}erfc\left(\frac{I_{sig}(1, N_1) - D_{th}}{\sqrt{2}\sigma(1, N_1)}\right) \\ &+ \frac{1}{2}erfc\left(\frac{D_{th} - I_{sig}(0, N_1)}{\sqrt{2}\sigma(0, N_1)}\right)\Bigg]\end{aligned} \tag{41}$$

The average P_e will be

$$P_e = \sum_{N_1=0}^{N-1} P'_{Nl}P_e(N_1) \tag{42}$$

7.3 Reflection noise

To analyse the effect of reflection noise, the echo path A, B, ..., G as shown in Fig. 3 has been considered. The amplifier is saturated by the signal and echo power in the desired channel and also the signal and echo power in interfering channels. Let the echo of bit b_r affect the signal corresponding to bit b_c. Further, assume that when the signal corresponding to b_r was reflected by OA to form the echo, M_1 interfering channels were having bit 1. The average echo power at the amplifier input in the desired channel is given by

$$P_{ae}(b_r) = \sum_{M_1=0}^{N-1} P_{in}(b_r)L_{S2A}^2 R_{sp}[1 - R]P'_{M1}R_{amp} \tag{43}$$

The average echo power in the interfering channels will be

$$\begin{aligned}P_{aei}(b_r) = \sum_{M_1=0}^{N-1} \{&[M_1P_{in}(1) + (N - 1 - M_1)P_{in}(0)] \\ &\times L_{S2A}^2 R_{sp}[1 - R]P'_{M1}R_{amp}\}\end{aligned} \tag{44}$$

P_{M1}' is obtained from eqn. 36 by replacing N_1 by M_1. The L_{S2A} represents the loss between splice 2 and amplifier. R_{amp} depends on the gain $G(b_r, M_1)$, which is

$$G(b_r, M_1) = G_0 \exp\left(-[G(b_r, M_1) - 1] \times \frac{[P_{in}(b_r) + M_1P_{in}(1) + [N - 1 - M_1]P_{in}(0)]}{P_{sat}}\right) \tag{45}$$

Let N_1 channels out of N–1 interfering channels be at bit 1 when the signal corresponding to bit b_c is amplified. The gain of amplifier for this signal is

$$G(b_c, b_r, N_1) = G_0 \exp\left(-[G(b_c, b_r, N_1) - 1] \times \frac{[P_{in}(b_c) + P_{ae}(b_r) + [N_1P_{in}(1) + [N-1-N_1]P_{in}(0)] + P_{aei}(b_r)]}{P_{sat}}\right) \tag{46}$$

In the numerator of the argument of this exponential function, the first term is the signal in the desired channel, the second term echo in the desired channel, the third term signal power in interfering channels, and the last term echo power in interfering channels. The signal power, the echo power and ASE noise PSD at the receiver are:

$$P_{sig}(b_c, b_r, N_1) = P_{in}(b_c)G(b_c, b_r, N_1)L_{AR} \tag{47a}$$

$$P_{echo}(b_c, b_r, N_1) = P_{ae}(b_r)G(b_c, b_r, N_1)L_{AR} \tag{47b}$$

$$S_{sp}(b_c, b_r, N_1) = n_{sp}[G(b_c, b_r, N_1) - 1]h\nu L_{AR} \tag{47c}$$

The signal current and noise variance are determined using these equations, and are given by:

$$I_{sig}(b_c, b_r, N_1) = R_0[P_{sig}(b_c, b_r, N_1) + P_{echo}(b_c, b_r, N_1)] \tag{48a}$$

$$\begin{aligned}\sigma^2(b_c, b_r, N_1) &= 2e[I_{sig} + R_0S_{sp}B_o]B_e + 4R_0^2P_{sig}S_{sp}B_e \\ &\quad + 4R_0^2P_{echo}S_{sp}B_e + R_0^2S_{sp}^2[2B_oB_e - B_e^2] \\ &\quad + 2R_0^2P_{sig}P_{echo} + \frac{4kTB_e}{R_L}\end{aligned} \tag{48b}$$

As before, there are many signal levels for both bit 1 and bit 0 and the threshold is computed using the lowest level for bit 1 and the highest level of bit 0, so that the probability of error for the chosen levels is equalised. The probability of error $P_e(b_r, N_1)$ for a given b_r and N_1 will be

$$\begin{aligned}P_e(b_r, N_1) = \frac{1}{2}\Bigg[&\frac{1}{2}erfc\left(\frac{I_{sig}(1, b_r, N_1) - D_{th}}{\sqrt{2}\sigma(1, b_r, N_1)}\right) \\ &+ \frac{1}{2}erfc\left(\frac{D_{th} - I_{sig}(0, b_r, N_1)}{\sqrt{2}\sigma(0, b_r, N_1)}\right)\Bigg]\end{aligned} \tag{49}$$

The average P_e will be

$$P_e = \sum_{b_r=0}^{1}\left[\frac{1}{2}\sum_{N_1=0}^{N-1} P'_{N1}P_e(b_r, N_1)\right] \tag{50}$$

where P'_{N1} is given by eqn. 36.

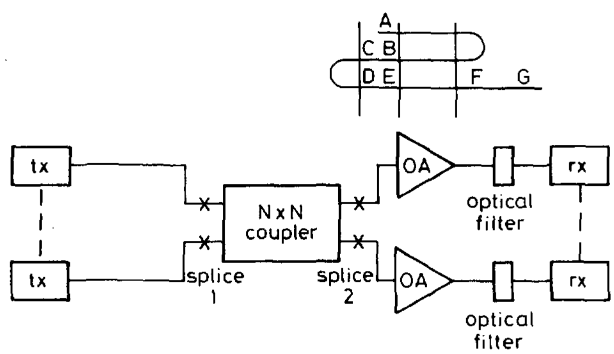

Fig.3 *Star network with preamplifiers*
A, B, ... G is echo path

8 Example

Consider a N-user star network (N = 2, 4, 8, 16, ...) and data rate for each channel as 1 Gbit/s. Typical values of parameters in the network are as follows [6, 2]:

- quantum efficiency of photodetector 0.95
- length of fibre from user to star coupler 1km
- attenuation coefficient of fibre 0.2dB/km
- unsaturated amplifier gain 1800
- saturation power level of amplifier –6dBm
- optical filter bandwidth 10GHz
- insertion loss of each 2 × 2 coupler 0.5dB
- insertion loss of splice 0.5dB
- coupler variability 0.5dB
- amplifier facet reflectivities 10^{-5} (–50.25dB)
- amplifier coupling loss 3dB
- temperature 300K
- load resistance 100ohms
- spontaneous emission factor 2.0
- electrical bandwidth of receiver 1GHz
- operating wavelength 1.55μm
- filter insertion loss 0.5dB

For this network, numerical computations have been made and the required transmitted power for a given number of users N and bit error rate of 10^{-9} is determined without OA, with optical postamplifier and preamplifier. In the analysis with OA, effect of gain saturation and reflection noise have also been considered. These results are shown in Figs. 4–13. In the following, results and inferences based on the preceding figures are discussed.

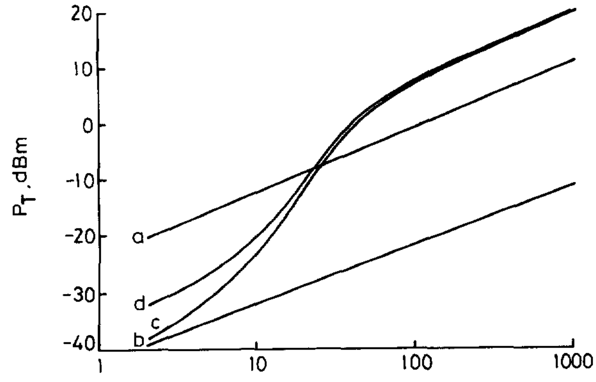

Fig.4 *Variation of minimum required average transmitter power P_T with number of users N in postamplifier scheme for ε = 0.0*
a Without Amplifier
b Ideal amplifier
c Amplifier with gain saturation
d As *c* but with reflection $R_{sp} = 10^{-2}$

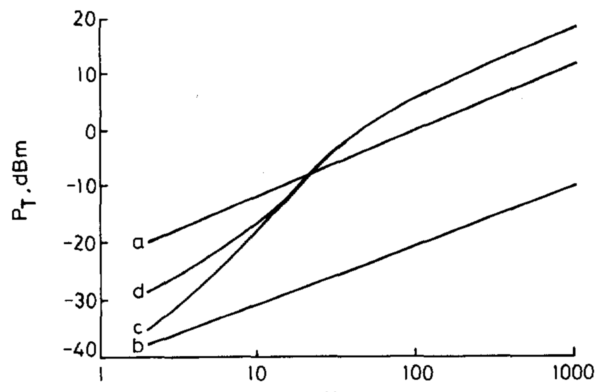

Fig.5 *Variation of minimum required average transmitter power P_T with number of users N in postamplifier scheme for ε = 0.05*
a Without Amplifier
b Ideal amplifier
c Amplifier with gain saturation
d As *c* but with reflection $R_{sp} = 10^{-2}$

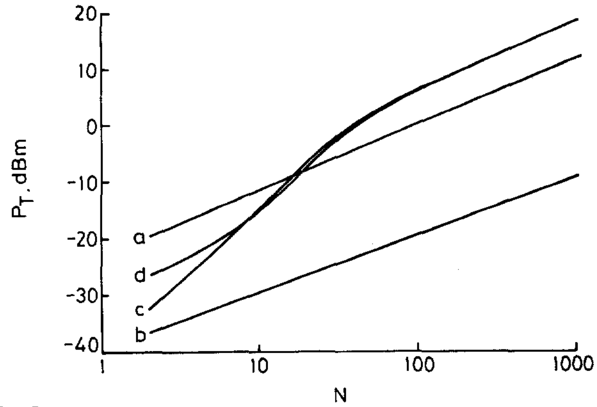

Fig.6 *Variation of minimum required average transmitter power P_T with number of users N in postamplifier scheme for ε = 0.10*
a Without Amplifier
b Ideal amplifier
c Amplifier with gain saturation
d As *c* but with reflection $R_{sp} = 10^{-2}$

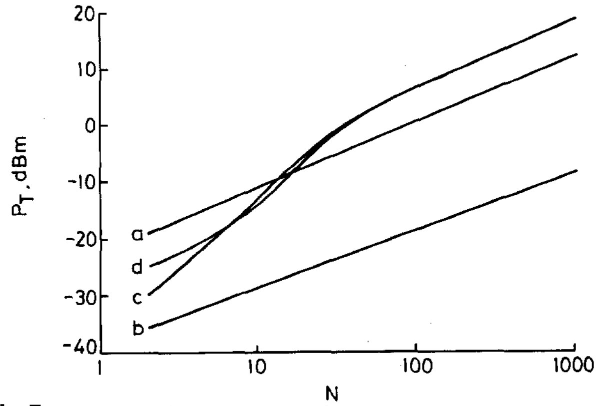

Fig.7 *Variation of minimum required average transmitter power P_T with number of users N in postamplifier scheme for ε = 0.15*
a Without Amplifier
b Ideal amplifier
c Amplifier with gain saturation
d As *c* but with reflection $R_{sp} = 10^{-2}$

(i) Use of an ideal amplifier reduces the required minimum transmitter power level for both postamplifier and preamplifier cases. The reduction is greater in the postamplifier than in the preamplifier because ASE noise undergoes more attenuation.

(ii) Gain saturation in the amplifier results in an increase in the required transmitter power. The difference in power level is referred to as power penalty. In the postamplifier, the penalty becomes more severe with increasing N. It can be explained as follows. As N increases, the star coupler's split loss also increases. Therefore more transmitted power is required to compensate for this loss. It results in a decrease in gain of the amplifier owing to an increase in gain saturation. The penalty becomes so severe after $N = 32$ that it overcomes the amplifier gain resulting in degradation of system performance. The reason for the gain saturation penalty exceeding unsaturated amplifier gain is that for large N, the gain for bit 1 is almost unity due to saturation effect and hence ASE noise PSD is negligible, but for bit 0, the gain is much greater than unity. This results in substantial ASE noise for bit 0. Therefore on average, the noise level is greater than without the amplifier. It degrades the system performance for large N. The same trend is observed for all values of extinction ratio (ε). The gain saturation effect for small N (i.e. N = 2, 4, 8, 16) increases with the increase in ε. At higher N, it does not change significantly with change in ε.

In the preamplifier, the effect of average gain saturation is small for low values of N. It increases with an increase in N. The increase is much faster for higher values of ε because with an increase in ε, more average transmitter power is required to maintain the given BER. This implies more power at the input of amplifier and hence more gain saturation.

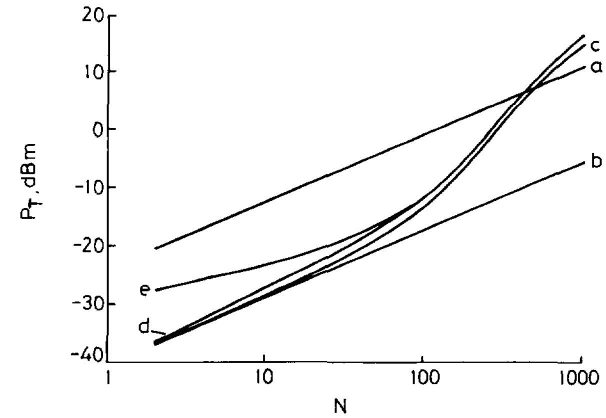

Fig. 8 *Variation of minimum required average transmitter power P_T with number of users N in preamplifier scheme for $\varepsilon = 0.0$*
a Without Amplifier
b Ideal amplifier
c Amplifier with average gain saturation
d Amplifier with average gain saturation and gain fluctuation
e As *d* but with reflection $R_{sp} = 10^{-2}$

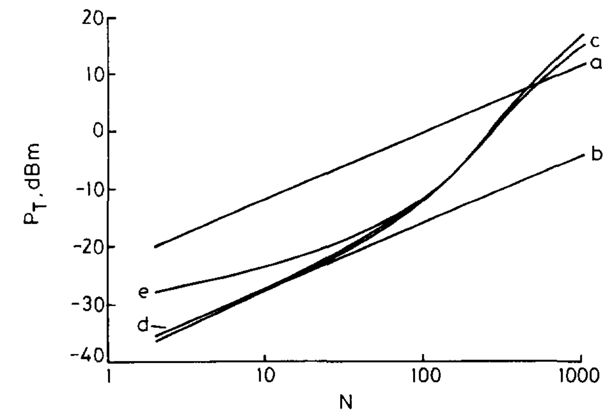

Fig. 9 *Variation of minimum required average transmitter power P_T with number of users N in preamplifier scheme for $\varepsilon = 0.05$*
a Without Amplifier
b Ideal amplifier
c Amplifier with average gain saturation
d Amplifier with average gain saturation and gain fluctuation
e As *d* but with reflection $R_{sp} = 10^{-2}$

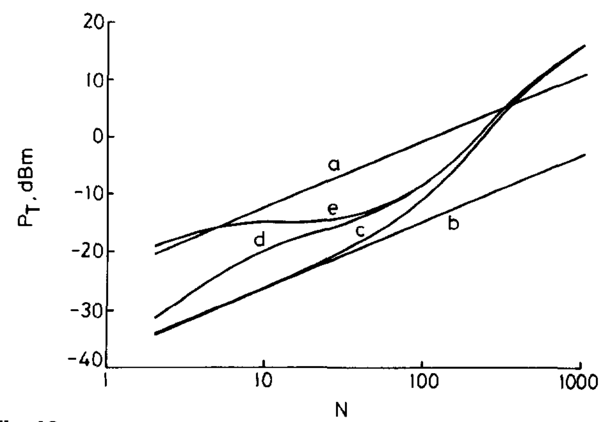

Fig. 10 *Variation of minimum required average transmitter power P_T with number of users N in preamplifier scheme for $\varepsilon = 0.10$*
a Without Amplifier
b Ideal amplifier
c Amplifier with average gain saturation
d Amplifier with average gain saturation and gain fluctuation
e As *d* but with reflection $R_{sp} = 10^{-2}$

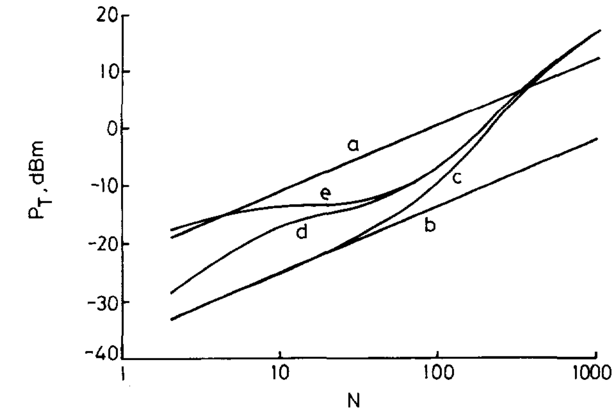

Fig. 11 *Variation of minimum required average transmitter power P_T with number of users N in preamplifier scheme for $\varepsilon = 0.15$*
a Without Amplifier
b Ideal amplifier
c Amplifier with average gain saturation
d Amplifier with average gain saturation and gain fluctuation
e As *d* but with reflection $R_{sp} = 10^{-2}$

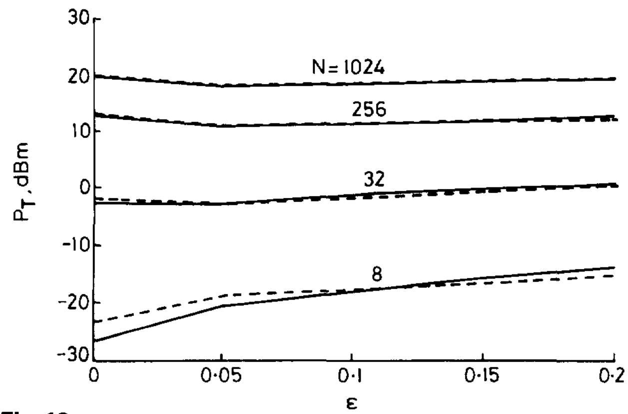

Fig. 12 *Variation of P_T with ε for different N in postamplifier scheme*

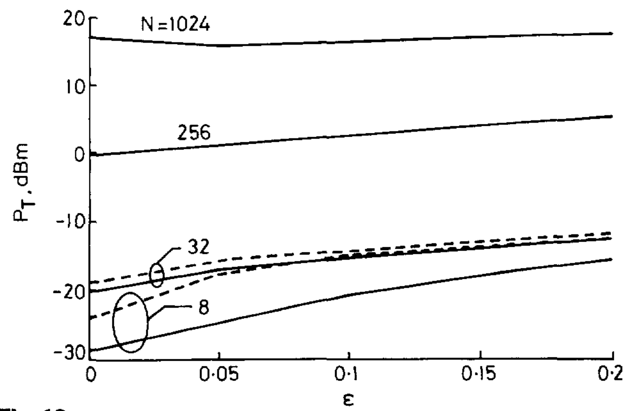

Fig. 13 *Variation of P_T with ε for different N in preamplifier scheme*

The gain fluctuation effect in the preamplifier first increases with N and after attaining maximum value decreases with increase in N (Figs. 8–11). Since the gain fluctuation effect depends on the variances of the probability density function of gain, for lower value of N, variance increases with N. However, for large N average gain shifts towards the lower extreme of the range (1 to G_0) and the levels are cluttered together. This results in a decrease in variance with N at large N. For some intermediate value of N, the variance will be maximum. This explains the variation of the gain fluctuation effect with N.

(iii) The effect of reflection noise reduces with the increase in N for both post and preamplifiers. As N increases, the coupler's split loss also increases which

Table 1: Number of users supported in postamplifier scheme for $\varepsilon = 0.10$

P_T, dBm	Number of users			
	Without amplifier	Ideal amplifier	OA with gain saturation	OA with gain saturation and reflection ($R_{sp} = -20$dB)
0	64	>1024	32	32
−10	8	512	8	16
−20	0	64	4	4
−30	0	8	2	0

results in increased required transmitter power. In the postamplifier case it will give rise to gain saturation and hence a reduction in amplifier gain, while in the preamplifier, the total power at the amplifier input increases with the increase in number of channels. Therefore amplifier gain reduces with an increase in N in the preamplifier also. A reduction in amplifier gain means a reduction in amplifier reflectance, which reduces the degrading effect of reflection. It is also observed that preamplifier scheme is more severely affected by reflection than the postamplifier scheme. This can be attributed to more gain saturation in the postamplifier.

(iv) Figs. 12 and 13 show that the average required transmitter power is a minimum at $\varepsilon = 0.05$ for high N (i.e. $N \geq 32$ for postamplifier and $N \geq 1024$ for preamplifier). This is contrary to the expected trend of an increase in required minimum transmitter power with an increase in ε. The physical reason for this is as follows. At $\varepsilon = 0.0$ there is no gain saturation for bit 0 and hence ASE noise will be greater. When ε becomes 0.05, input power for bit 0 also saturates the amplifier. This causes a reduction in ASE noise and hence a reduction in transmitter power level. But due to the increase in ε, power level will increase. For a high value of N, it happens that a reduction in power level owing to a reduction in amplifier noise is greater than the increase in power level due to an increase in ε.

(v) Figs. 4–7 show that at higher ε, reflection slightly improves the system performance for smaller N (i.e. $N \leq 32$). This is contradictory to the expected trend. In this case, the echo signal saturates the amplifier and leads to reduced ASE. The required transmitter power due to reflection will increase as expected. However, due to reduction in ASE noise, the required power level will decrease; for a smaller value of N, the latter effect is predominant, leading to an improvement in system performance.

Figs. 4–11 are used as source for the data in tabular form in Tables 1 and 2. The following observations are made from these Tables.

(vi) With the use of an ideal amplifier in the network, the number of users is increased for both the post and preamplifier configuration. When the effect of amplifier saturation and reflection noise is considered, there may or may not be any advantage. For example, in the postamplifier case there is no increase in the number of users when the transmitter power level is 0dBm. As the transmitter power decreases to −20dBm, the number of users increases from 0 to 4 ($R_{sp} = -20$dB). In general, the number of users must increase as the reflection coefficient of splice decreases for a fixed transmitter power level. But there are some exceptions e.g. number of users decreases from 16 to 8 when R_{sp} is reduced from −20dB to no reflection. The reason for this is as given in (v).

(vii) In the preamplifier case, a greater number of users are supported except when P_T is very low and R_{sp} is very high e.g. $P_T = -20$dBm and $R_{sp} = -20$dB. It is observed that the number of users reduces with decrease in P_T and increase in R_{sp}. However, networks are usually not designed for very low transmitter powers, and reflection coefficients due to splices are much lower than −20dB.

(viii) For a typical transmitter power level of 0dBm, the preamplifier scheme is always better compared with the postamplifier scheme in terms of an increase in the number of users. When the transmitter power level is low (less than −20dBm) and the reflection coefficient high, the postamplifier is better. However, networks are not implemented for such low transmitter power levels and high reflection coefficients.

9 Conclusions

In this paper, placement of optical amplifiers in a WDM star network has been investigated. The number of users supported by the topology for post and preamplifier placement has been determined. The effect of gain saturation and reflection noise in both these cases has been analysed. This effect has been evaluated in terms of number of users supported or the required transmitter power for a given BER. The analyses also consider the effect of extinction ratio. When there is no gain saturation and reflection noise in the network, the postamplifier scheme performs better. In the presence of gain saturation, the postamplifier performs better for low transmitter power levels. However, such low transmitter power levels are not used in practice. When the number of users and consequently transmitter power increases, performance in the postamplifier scheme

Table 2: Number of users supported in preamplifier scheme for $\varepsilon = 0.10$

P_T, dBm	Number of users				
	Without amplifier	Ideal amplifier	OA with average gain saturation	OA with average gain saturation and gain fluctuation	OA with average gain saturation, gain fluctuation and reflection ($R_{sp} = -20$dB)
0	64	>1024	128	128	128
−10	8	128	64	64	64
−20	0	32	16	8	0
−30	0	4	4	2	0

degrades, while in the preamplifier scheme it improves. The power penalty due to reflection noise reduces in both schemes with increase in the number of users.

Overall, the preamplifier is a better scheme than the postamplifier in WDM star topology networks.

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