slope over the wavelength range 1520-1600nm [8]. Fig. 3 shows the dispersion accumulated through the concatenation of a 75km singlemode fibre (SMF) (16.9ps/nm/km) and the DCF. In the figure, the dispersion of a 75km DSF is shown for comparison.

Fig. 4 Gain profile of dispersion compensating Raman amplifier

Fig. 4 shows the gain profile of the DCRA. In the experiment, eight channel WDM signals with -20dBm/ch are launched into the DCRA. The maximum gain is 19dB when the pump unit operates at an output of 800mW. By adjusting the output powers of each laser, we are able to tune the gain level and its flatness simultaneously. The loss of the DCF can be adjusted arbitrarily between 10dB and the lossless region over a 50nm wavelength range within a flatness of 1.0dB. For lossless operation, the total pump power was 939mW.

Fig. 5 Noise figure of dispersion compensating Raman amplifier for lossless operation

Fig. 5 shows the noise figure (NF) of the DCRA under lossless operation. Since double Rayleigh backscattering increases the optical noise in a Raman amplifier, the NF at shorter wavelengths with larger Rayleigh scattering tends to be larger [9]. The increase in NF at 1585nm is caused by the wavelength dependence of the isolator loss at the input side.

Fig. 6 shows the gain profiles corresponding to different input powers, where the pump powers are maintained under the same conditions. The gain is saturated as the input power increases. This saturation has almost no wavelength dependence, in contrast to EDFAs.

In conclusion, we have demonstrated broadband Raman amplification in a ~1300ps/nm DCF using polarisation-combined four channel WDM 1440nm laser diodes for the pumping light source. Losses in the DCF are completely compensated for by means of Raman gain over 50nm bandwidth within ±0.5dB variation.

© IEE 1998

Electronics Letters Online No: 19985109

Y. Emori, Y. Akasaki and S. Namiki (Opto-Technology Laboratory, Furukawa Electric Co., Ltd., 6 Yasutaka-Koigan-Doi, Iruhara, Chiba 290-8555, Japan)

E-mail: yemori@ch.furukawa.co.jp

References


Burst-mode optical receiver with two preamplifiers having different bandwidths

Jang-Won Park, Y.C. Chung and Chang-Hee Lee

The authors propose a new burst-mode optical receiver with two preamplifiers, each with a different bandwidth, to suppress the burst-mode penalty without preamble bits. The receiver, designed with 0.6um CMOS, has a sensitivity of -40.8dBm with a dynamic range of 20.8dB at 155MHz.

Introduction: The PON (passive optical network) has been considered as a cost-effective solution for the broadband access network. The burst-mode optical receiver is a basic building block of the PON that utilizes time division multiple access. To receive burst-
mode signals with different amplitudes, both feedback type and feed-forward type burst-mode receivers have been reported [1, 2]. However, there is a burst-mode penalty of 1.5dB, since the decision threshold is corrupted with noise [3]. This penalty can be reduced by using preamble bits and a complex control scheme [3, 4]. In this Letter, we propose a new burst-mode optical receiver to suppress the burst-mode penalty without preamble bits. The receiver, designed with 0.6μm CMOS model parameters, has a sensitivity of –40.8dBm with a 20.8dB dynamic range at 155Mbit/s. The minimum required guard time is < 3 bit periods, when the power difference between successive packets is 20dB.

![Fig. 1 Schematic diagram of proposed burst-mode receiver](image)

**Operation principles:** A schematic diagram of the proposed burst-mode receiver is shown in Fig. 1. It consists of an optical power splitter, an optical delay line, two preamplifiers, a threshold control circuit, and a limiting amplifier. The upper part of Fig. 1 represents a well-known optical receiver. The delayed optical packets are received by preamplifier A which has a bandwidth of ~0.65 times the bit rate of the incoming signal. The delay time is about half of the packet duration. A small portion of the input power is coupled to preamplifier B which has a much narrower bandwidth compared with preamplifier A. Then, the output of preamplifier B becomes the average power of the incoming input packet. It can be used as the threshold voltage of the limiting amplifier. Therefore, the burst-mode penalty induced by the noise-corrupted decision threshold can be suppressed.

The threshold control circuit in Fig. 1 selects a threshold voltage for each packet. Initially, the output of preamplifier B is used as a threshold voltage of the limiting amplifier. After detection of the start of the packet, the sample and hold (S/H) output of preamplifier B is used as a threshold. The analogue multiplexer is used to select one of the two output states. When the end of the packet is detected, the multiplexer is switched back to the preamplifier B output. The employed multiplexer reduces the required minimum guard time which was limited by the charging and discharging time of the sample and hold capacitor [3, 4].

![Fig. 2 Sensitivity and dynamic range of designed receiver at bandwidth of 110MHz](image)

**Design results:** We designed the proposed burst-mode receiver at 155Mbit/s using 0.6μm CMOS model parameters. The bandwidth of preamplifier A was set to 110MHz when the equivalent input capacitance of preamplifier A was 0.6pF. We obtain a sensitivity of –40.8dBm with a dynamic range of 20.8dB and a feedback resistance of 100kΩ. Both the sensitivity and dynamic range increase as the feedback resistance decreases, as shown in Fig. 2.

![Fig. 3 Penalty of designed receiver against preamplifier B bandwidth for 2'th PRBS input pattern](image)

- total penalty (nominal)
- ambiguity penalty (nominal)
- total penalty (worst case)
- ambiguity penalty (worst case)

The proposed receiver can suppress the burst-mode penalty induced by the noise corrupted decision threshold. However, there exists a fluctuation in the decision threshold voltage when we average random input data, e.g. PRBS (pseudo random binary sequence). The effects of this fluctuation on the power penalty were investigated as a function of the preamplifier B bandwidth. The results are shown in Fig. 3 for a 2'th PRBS input pattern, which is the scrambled output of the ATM-PON upstream signal. The worst case penalty is shown for the peak-to-peak value and the nominal one for the root-mean-square value of the fluctuation. The total penalty is the sum of the penalty induced by the fluctuation, the burst-mode penalty [3], and the splitting loss of the power splitter. When the bandwidth of preamplifier B is 500kHz, the total penalties are 0.76dB and 10 dB for the worst case and the nominal case, respectively.

![Fig. 4 Transient response of designed burst mode receiver at limiting amplifier input, when successive packets have 20dB power difference](image)

- required minimum guard time = 9ns
- required guard delay time = 8.5ns

When the low power packet follows the high power packet, the low power packet must be delayed until the amplitude of the last bit in the previous packet is sufficiently small, otherwise we have interference between these packets. To find the minimum guard time for the reception of successive packets without interference, we investigated the transient response of the designed receiver at the input of the limiting amplifier, as shown in Fig. 4. When the power difference of the packets (~20dBm and ~40dBm) is the same as the targeted dynamic range of 20dB, the fall time (from 10 to 90%) of the high power packet is 9ns. Thus, the minimum required guard time is 9ns. If we include one bit margin for the guard time, the required guard time is < 3 bit periods (6.5ns x 3 = 19.5ns). This satisfies the value recommended by the ATM-PON standard, which defines 4 bit periods as the minimum guard time.

**Conclusion:** In conclusion, we proposed a new burst-mode receiver with two preamplifiers, each having a different bandwidth. The
Crosstalk reduction by carrier suppression in an analogue WDM optical communication system

F.S. Yang, M.E. Marhic and L.G. Kazovsky

Crosstalk reduction is demonstrated in a WDM analogue optical communication system using optical carrier suppression. A crosstalk reduction by 20 dB over 2 GHz is achieved while maintaining the same received RF power.

Introduction: Stimulated Raman scattering (SRS) crosstalk has been identified as a major obstacle to the implementation of analogue WDM optical communication systems [1, 2]. A crosstalk cancellation technique using parallel fibres has been proposed in [3], but the bandwidth is limited by fibre nonlinearities and dispersion to 200 MHz. In this Letter, we propose a new technique for crosstalk reduction by suppressing the optical carrier. 20 dB of crosstalk reduction is achieved over a 2 GHz bandwidth.

Background: Most analogue optical communication links operate with modulation depth of a few percent in order to maintain high linearity. As a result, most of the transmitted optical power resides in the optical carrier power $P_c$, which contains no useful information. Suppressing this optical carrier has been shown to effectively increase the modulation depth without sacrificing linearity, thus increasing the linear dynamic range of analogue optical communication systems [4, 5].

Crosstalk in analogue WDM optical communication systems arises from fibre nonlinearities, most notably SRS [6] and cross-phase modulation combined with group velocity dispersion (XPM/ GVD) [7] in standard singlemode fibre (SMF) operating at ~1550 nm. It can be shown that both SRS and XPM/GVD efficiencies increase approximately as $P_c^2$; thus if $P_c$ can be reduced by a factor $r$, without sacrificing the received RF power, then the crosstalk will be reduced by $r^2$.

Experiment and discussion: The experimental setup is shown in Fig. 1. A DBR laser operating at 1550 nm, which we called the pump laser, is externally modulated using an LNO: Mach-Zehnder (MZ) modulator, at frequency $f_c$, that can be tuned from 0 to 2 GHz. An unmodulated external cavity laser, tuned 10 nm above the pump laser, is used as the probe. External phase modulation by a 1 GHz pseudorandom bitstream is applied to suppress stimulated Brillouin scattering (SBS). At the output of the EDFA, both wavelengths have equal power. After 25 km of SMF, an optical filter selects the wavelength to be received by the photodetector. Crosstalk is measured by dividing the received RF power on the probe wavelength by the received RF power on the pump wavelength and expressing the result in dBc.

We first operate the MZ modulator with normal bias, i.e., no carrier suppression. The modulation depth is 4% corresponding to typical CATV operation. The transmitted power into the fibre is 18 dBm/λ. Fig. 2a shows the measured crosstalk. The solid curve is the theoretical calculation of the SRS crosstalk based on [6]. At low $P_c$, SRS is the dominant crosstalk mechanism, and the measured data agree well with the theoretical prediction. However, because SRS efficiency decreases with increasing $P_c$ while XPM/GVD increases with increasing $P_c$, crosstalk beyond 500 MHz includes a significant amount of XPM/GVD effect and begins to deviate from the theoretical SRS prediction.

Next we suppress the optical carrier by low-biasing the MZ modulator [5]. By adjusting the bias of the MZ modulator and the gain of the EDFA, we reduce $P_c$ by 10 dB. The residual optical carrier is strong enough so that direct detection can still be employed. Because of the increase in modulation depth arising from the suppressed carrier, the received RF power remains the same. We maintain equal optical power for both wavelengths at the output of the EDFA. The resulting measured crosstalk is shown in Fig. 2b. As predicted, a 10 dB reduction in $P_c$ results in a 20 dB reduction in crosstalk, at both low and high $f_c$. The residual crosstalk is below the CATV industry standard of -50 dBc for CNR. The 2 GHz limit is imposed by the RF sources available in our laboratory (theory predicts no bandwidth limitation in crosstalk reduction using carrier suppression).