minimum phase channel $H_2(z)$. Hardware offsets and quantisation errors may cause performance degradation.

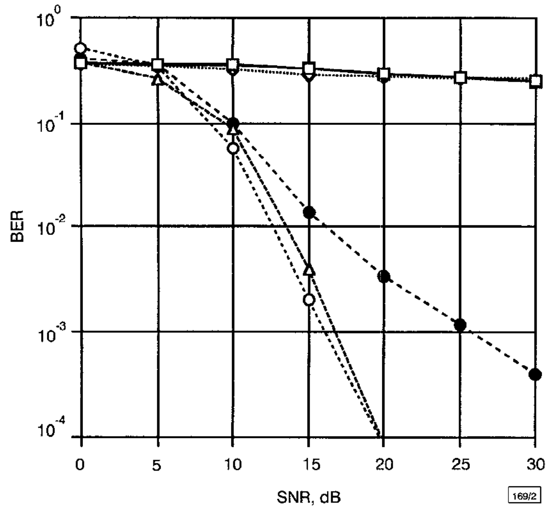

Fig. 2 *BER comparison for nonminimum phase channel*

u linear adaptive filter
e 2-layer perceptron
O 3-layer perceptron
n 2-layer perceptron with decision feedback
● 2-layer perceptron with decision feedback using neuro-chip sets

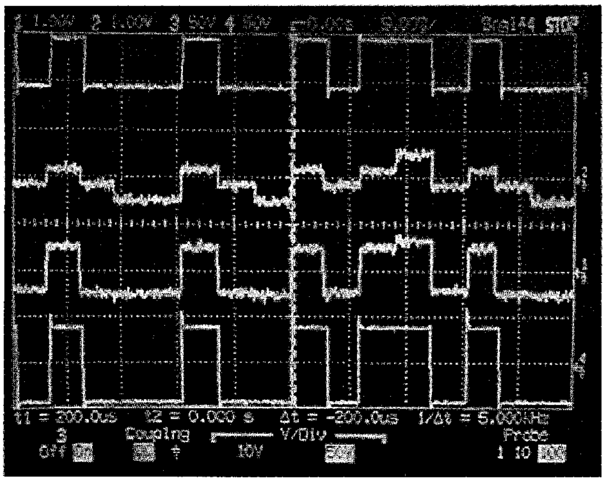

Fig. 3 *Equalisation in experiments for nonminimum phase channel*

Fig. 3 shows actual signals which are measured using an oscilloscope in the case of the nonminimum phase channel, $H_2(z)$. The first signal is an input generated in the host computer, and the second signal is the output of the channel that is also the input of the analogue neuro-chip. The third signal is the output of the analogue neuro-chip, and the last one is the result of thresholding.

Conclusions: In this Letter, a new modular analogue neuro-chip set with an on-chip learning capability is presented for adaptive nonlinear equalisers. Using the analogue neural equaliser, digital telecommunication receivers do not need, and can eliminate, supplementary devices such as analogue-to-digital converters and digital equalisers. Although several chips are used for experiments, one-chip integration is readily possible. The chips operate in real time, even with nonminimum phase channels.

22 August 1997
Electronics Letters Online No: 19971277

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Simple dynamic model of fibre amplifiers and equivalent electrical circuit

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Indexing terms: Fibre amplifiers, Erbium-doped fibre amplifiers, Equivalent circuits, Optical fibre networks

Building on the results in [1], the authors have determined that the gain dynamics of a doped-fibre amplifier are completely specified by its total number of excited ions whose time behaviour is described by a simple first-order differential equation. An equivalent electrical circuit is derived, so that any circuit analysis package can be used to solve for complex networks of optical fibre amplifiers.

Introduction: Transparent wavelength routed optical networks (WRON) will form large meshes of interconnected fibre links and cross-connects. Doped-fibre amplifiers will become key components of such networks, and a study of their gain dynamics in a networking scenario is essential for the WRON design.

In this Letter, we build on the results of [1], and show that the gain dynamics of a doped-fibre amplifier are completely specified by its total number of excited ions whose time behaviour is described by a simple first-order differential equation. Such an equation can also describe the time behaviour of an equivalent capacitive electrical circuit. Any circuit analysis package can then be used to solve for complex networks of optical fibre amplifiers.

Analysis: We start from the rate and photon equations used in [1]), derived assuming a two-level system for the dopant ions, an homogeneously broadened gain spectrum, no background loss, and no self-saturation by spontaneous emission (ASE). The rate equation for the fraction of excited ions N_2, $0 \le N_2 \le 1$, is

$$\frac{\partial N_2(z,t)}{\partial t} = -\frac{N_2(z,t)}{\tau} - \frac{1}{\rho A}\sum_{j=0}^{N} u_j \frac{\partial Q_j(z,t)}{\partial z} \qquad (1)$$

and the equations describing the propagation along z of the photon fluxes Q_k [photons/s] of channel k, $k = 0, ..., N$, are:

$$\frac{\partial Q_k(z,t)}{\partial z} = \rho u_k \Gamma_k [\sigma_k^T N_2(z,t) - \sigma_k^a] Q_k(z,t) \qquad (2)$$

where τ [s] is the fluorescence time, ρ [m^{-3}] is the ion density in the doped fibre core of effective area A [m^2]; Γ_k, σ_k^e [m^2] and σ_k^a [m^2] are the confinement factor, emission and absorption cross-sections of channel k, respectively, and $\sigma_k^T \triangleq \sigma_k^e + \sigma_k^a$. The length of the amplifier is L [m]. Channels entering at $z = 0$ have $u_k = 1$ while those at $z = L$ have $u_k = -1$. The pump is on channel 0.

Dividing both sides of eqn. 2 by $Q_k \neq 0$, multiplying by dz and integrating from $z = 0$ to L yields

$$G_k(t) = B_k r(t) - A_k \qquad k = 0, ..., N \qquad (3)$$

where $G_k(t) \triangleq \int_0^L u_k \partial Q_k / Q_k = \ln(Q_k^{out}(t)/Q_k^{in}(t))$ is the logarithmic gain, $A_k \triangleq \rho \Gamma_k \sigma_k^a L$ and $B_k \triangleq \Gamma_k \sigma_k^T / A$ are non-dimensional parameters, and $r(t) \triangleq \rho A \int_0^L N_2(z, t) dz$ is the total number of excited ions in the amplifier, which we call the 'reservoir'; it is a number between 0 and ρAL.

From eqn. 3, it is clear that the log-gains of all channels are linear functions of the reservoir r only. We can also rewrite eqn. 3 in a more familiar form as $G_k(t) = \rho L\Gamma_k(\sigma_k^T x(t) - \sigma_k^a)$, where $x(t) \triangleq (1/L)\int_0^L N_2(z, t)dz$ is the mean fraction of excited ions [2].

Multiplying both sides of eqn. 1 by dz and integrating from 0 to L yields

$$\frac{\partial r(t)}{\partial t} = -\frac{r(t)}{\tau} - \sum_{j=0}^{N}(Q_j^{out} - Q_j^{in}) \qquad (4)$$

Using eqn. 3 in eqn. 4, we finally obtain a first-order differential equation describing the dynamic time behaviour of the system's state, i.e. the reservoir $r(t)$:

$$\dot{r}(t) = -\frac{r(t)}{\tau} + \sum_{j=0}^{N} Q_j^{in}\left(1 - e^{B_j r(t) - A_j}\right) \qquad (5)$$

Once the initial condition $r(0) = r_0$ is specified, the solution of eqn. 5 is unique. r_0 can be any number in the allowed range $[0, \rho AL]$. However, if at time $t = 0^-$, i.e. one instant before the start of the observation period, the amplifier is at equilibrium, then r_0 must satisfy eqn. 5 with $\dot{r}(0^-) = 0$:

$$r_0 = \tau \sum_{j=0}^{N} Q_j^{in}(0^-)\left(1 - e^{B_j r_0 - A_j}\right) \qquad (6)$$

which is the well-known Saleh steady state equation [3]. For a starting guess in its numerical solution, the upper bound $\tau\Sigma_{j=0}^{N} Q_j^{in}(0^-)$ can be used.

Note that, since by eqn. 5 the derivative of r exists, r is a continuous function of t, even when the inputs $\{Q_j^{in}\}$ are discontinuous. In exactly the same way as for charge on a capacitor, the reservoir cannot 'jump' instantaneously. Note also that eqn. 5 has a nice physical meaning: the variation of the reservoir (the 'charge' on the amplifier) is given by the total input flux $\Sigma_{j=0}^{N} Q_j^{in}$, minus the output flux $\Sigma_{j=0}^{N} Q_j^{out}$, minus the spontaneous decay from the excited level r/τ.

Hence, by identifying photon fluxes with currents, and the reservoir with the charge on a capacitor, or better yet with the voltage across a capacitor of capacity $C = 1$, we obtain the equivalent electric circuit depicted in Fig. 1*a*. The circuit is composed of N input current sources (the channels) and of a pump channel 0, hidden inside the amplifier. The currents feed the capacitor, whose voltage is the reservoir r. The currents out are voltage-controlled current generators, whose gain is $g_j(r) = e^{G_j(r)}$. The output pump current is shunted to ground and does not exit the amplifier. The RC constant of the capacitive circuit is τ. However, this is an active circuit, and it is clear that the actual time constants involved in the dynamics are essentially independent of τ.

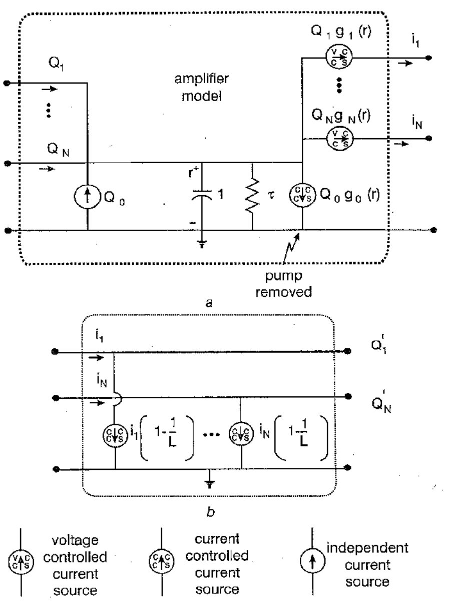

Fig. 1 *Fibre amplifier equivalent circuit and that for optical loss L*

a Fibre amplifier equivalent circuit
b For optical loss L

The circuit model for a subsequent loss L is shown in Fig. 1*b*. It is simply a bank of current-controlled current generators that shunt part of the output photons to ground, effectively wasting them.

Using these building blocks for the amplifier and the loss, any complex network of amplifiers is readily solved by any electric circuit simulator available on the market.

Acknowledgments: This work was supported by the European Community under INCO-DC project No 950959 'DAWRON', and by an NSERC Industrial Chair sponsored by Québec-Téléphone.

21 July 1997
Electronics Letters Online No: 19971273

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Suppression of optical beat interference using synchronised CDMA technique and in-band clipping carrier

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Indexing terms: Code division multiple access, Multi-access systems, Optical communication

The authors propose adding a coherent in-band clipping carrier to a previously demonstrated Walsh-code-based synchronised CDMA (S-CDMA) technique to further suppress the optical beat interference (OBI). Experimental results showed that a negligible system power penalty can be achieved even when the OBI-induced intensity noise was as high as –90.9dB/Hz. In addition, a large system dynamic range of 10dB can be obtained to relax the tight power control requirement on S-CDMA signals.

Introduction: The feasibility of using a Walsh-code-based synchronised CDMA (S-CDMA) technique to achieve multiple access in the presence of optical-beat and co-channel interference was previously demonstrated [1]. While conventional OBI suppression techniques such as out-of-band clipping-tone [2, 3] or over-modulating multiple frequency-division-multiplexed channels [4, 5] could cause in-band nonlinear distortions, the S-CDMA technique can effectively suppress the OBI without incurring the same problem and at the same time enjoy the full bandwidth of an optical fibre system. However, owing to the fact that a broadband CDMA cannot suppress OBI as effectively as narrow-band clipping tones, we found that some residual system penalties remained due to incomplete suppression of the OBI [1]. Therefore, in this Letter, we propose adding an *in-band* clipping carrier to the S-CDMA signal, to further suppress the residual OBI without using additional system bandwidth. By implementing this technique, the power control tolerance of the optical transmitters can be relieved by ~8dB compared to that of the previous case [1].

Experiment: Our experimental setup is shown in Fig. 1. An S-CDMA signal and an in-band clipping carrier were coherently combined and used to modulate a 1.3µm DFB laser (laser A). The S-CDMA signal was generated by spreading a 1.5625Mbit/s, 2^{20}–1