Adaptive channel precoding for personal communications

W. Zhuang, W.A. Krzymien and P.A. Goud

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A novel adaptive channel precoder is proposed to improve the bit error rate performance of a personal wireless communication system, without increasing the complexity of the portable unit, over a slowly fading channel with severe intersymbol interference.

Introduction: In high-bit-rate personal communication systems, the intersymbol interference (ISI) due to multipath can dramatically increase the transmission bit error rate (BER). The decision-feedback equaliser (DFE) is the most popular equaliser for combating ISI. When coded modulation is used to improve the system performance, the combination of coding with the DFE does not perform well because zero delay decisions in a coded system are not sufficiently reliable for decision feedback. The problem can be solved without increasing the complexity of the portable unit by moving the equalisation function from the receiver of the portable unit to the transmitter of the base station. Haraslima-Thomlinson (HT) precoding is a transmitter equalisation technique which is very suitable for QAM modulation [1]. However, the precoding technique is not very suitable for time-varying fading channels, for which amplitude fading causes difficulty in retrieving the original amplitude information in the receiver; in particular when an amplitude limiter is used at the receiver with constant-envelope phase modulation, the use of modulo-arithmetic reduction to recover the original transmitted signal is impossible. This Letter proposes a new adaptive channel precoding technique for phase modulation, which predistorts both the carrier phase and amplitude of the transmitted signal in such a way that the signal arriving at the receiver is ISI free.

Adaptive channel precoding: Fig. 2 shows the structure of the adaptive channel precoder. The switchable delay device (SDD) is controlled in such a way that the precoder always introduces a constant time delay to the received signal. The precoding filter (PF) consists of feedforward (FF) and feedback (FB) linear filters. Two examples are studied here: The first is a fading channel (at time t) with a minimum-phase transfer function

\[ H(z) = 0.661 + 0.523 z^{-1} + 0.523 z^{-2} \]

The PF filter function \( F(z) = 1 \) can be simply set as the inverse of the channel transfer function, i.e., \( F(z) = 1/H(z) \). The PF filter allows only the first path signal (ISI free) to arrive at the receiver. The second is a fading channel (at time t) with a non-minimum-phase transfer function

\[ H(z) = 0.41 + 0.523 z^{-1} + 20z^{-2} \]

We use an FB filter with transfer function \( F(z) = 1/(0.41 + 0.523 z^{-1} + 20 z^{-2}) \) to equalise the minimum-phase component, and a (four-tap) FF filter with transfer function \( F(z) = 0.0625 + 0.125 z^{-1} + 0.125 z^{-2} + 0.5 z^{-3} \) to equalise the non-minimum-phase component [2]. The PF filter \( F(z) = F(z)/F(z) \) introduces a time delay of \( 4T \) to the received signal (here \( T \) is the unit time interval, which can be smaller than or equal to a symbol interval). To reduce the difference in time delays introduced by the filters \( F(z) \) and \( F(z) \), we modify the PF filter \( F(z) \) to \( F(z) = F(z)/F(z) \) so that the delay introduced by \( F(z) \) is \( 3T \). As a result, the delay difference of the precoders corresponding respectively, to \( H(z) \) and \( H(z) \) is reduced to \( T \). The delay of the SDD is set at \( T \) for \( H(z) \), but is reduced to zero for \( H(z) \); therefore, the precoder introduces a constant time delay to the transmitted signal. Generally, for a non-minimum-phase channel, the tap coefficients of the PF filter, \( c_1, \ldots, c_4 \), are zero except \( c_1 \) of the last tap; the delay introduced by the PF is \( (n - 1)T \). For a non-minimum-phase channel, the equivalent delay introduced by the precoder will be equal to or larger than \( 4T \) depending on the non-minimum-phase component.

The precoded signal may have amplitude fluctuations depending on the fading channel. The automatic gain control (AGC) unit is used to suppress the fluctuations and to keep the transmitted power constant, which may slightly degrade the performance improvement achieved by the PF filter. All the precoding functions are implemented in the transmitter only, and no additional functions need to be performed in the receiver for equalisation purposes, which is different from the HT precoding technique. Consequently, using an amplitude limiter in the receiver does not degrade the performance improvement achieved by the precoding technique proposed here. Based on the impulse response estimated in the reverse link over the previous data frame, an adaptive algorithm should be applied to obtain the complex tap coefficient arrays of the FF and PB filters for the current data frame, because the channel impulse response over two adjacent data frames can be considered as invariant for a slowly fading channel and the radio channel is reciprocal in a TDD system. The channel precoding is not subject to error propagation, which is an advantage of the precoder over conventional DFE.
Analysis and dimensioning of a unidirectional WDM optically amplified open ring network using channel tap-and-add devices

E. Gay, M.J. Chawki, D. Hui Bon Hoa, M. Le Ligné and V. Tholey

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A unidirectional, WDM optically amplified open ring network is analysed using a commercial software package. Optimisation of the ring configuration is performed under the constraint of high and nearly equal SNRs for all channels. For eight nodes, the maximum ring length is shown to be greater than 500km.

Introduction: Recently, many WDM ring architectures have been proposed for implementing uni- and bidirectional self-healing SDH rings [2, 3]. The WDM rings use an optical amplifier and a wavelength add drop multiplexer (ADM) at each node which perform dynamic wavelength routing and allow the network to be transparent to the signal format and the bit rate. We analyse a unidirectional open WDM ring network linking a main station to a small number of secondary nodes (eight) suited for the junction or regional network layer. This configuration (Fig. 1) permits wavelength sharing between different stations because the wavelengths circulating along the open ring are not removed. The bit rates under consideration are 155Mbit/s, 622Mbit/s and 2.5Gbit/s. The number of wavelengths sent by the main station is equal to the number N of secondary stations. The main station receives 2N wavelengths.

Fig. 1 Configuration of unidirectional open ring network

Optical signal spectra at the output of the first, fourth and eighth stations are shown

Our simulation package [1] is able to determine the optimal values of the link loss between amplifiers, input power at the nodes, wavelength arrangement, and unsaturated small signal gain of the EDFA, that are needed to maintain a high and nearly equal SNR for all channels [4].

Simulation description: The signal and ASE power accumulation along the EDFA chain are computed using a spectrally resolved model [5]. We use the Er:Al-Oxide-silicate fibre described in [6]. The erbium-fibre length is chosen to provide 24dB small signal gain with a 30mW, 1.48μm pump. The link losses between the amplifiers include fibre and passive coupler attenuations. The optical booster amplifier of the main station allows us to obtain an optical power of +6dBm for each channel. The link loss between this amplifier and the input to the first in-line amplifier (at station 1) is 21dB (17dB fibre attenuation 0.2dB/km × 85km, and 4dB insertion loss in the coupler). The power per channel at the input to the