ELECTRICAL SIGNAL PROCESSING TECHNIQUES IN LONG-HAUL, FIBER-OPTIC SYSTEMS

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ABSTRACT

In this paper we demonstrate the potential for electrical signal processing to mitigate the effect of intersymbol interference in long-haul, fiber-optic systems. Intersymbol interference in long-haul fiber-optic systems can severely degrade performance and consequently limit both the maximum distance and data rate. The sources of intersymbol interference include nonlinearity in the transmit laser, chromatic dispersion in systems operated at wavelengths other than the dispersion minimum of the fiber, polarization dispersion, and bandwidth limitations in the receiver. We discuss several techniques for reducing intersymbol interference in single-mode fiber systems with single-frequency lasers and show which techniques are appropriate at high data rates in direct and coherent detection systems. In particular, we analyze the performance of linear equalization, nonlinear cancellation, and maximum likelihood detection.

I. INTRODUCTION

In this paper we demonstrate the potential for electrical signal processing techniques at the receiver to mitigate the effect of intersymbol interference (ISI) in long-haul, fiber-optic systems. Intersymbol interference is a major impairment in long-haul, fiber-optic systems that limits both the transmission distance and data rate. Intersymbol interference can have many sources, including laser nonlinearity, nonideal receiver frequency response characteristics, and chromatic and polarization dispersion.

We consider linear equalization, nonlinear cancellation, and maximum likelihood detection to reduce intersymbol interference. All of these techniques can be made adaptive, and thus capable of optimizing system performance over a wide range of impairments and device characteristics. We analyze in detail these receiver signal processing techniques in long-haul, fiber-optic systems with single-mode fibers and single-frequency lasers[1]. We propose practical methods for implementing these techniques in high data-rate systems and show examples of the substantial performance improvement obtainable with these techniques.

II. SYSTEM

Figure 1 shows a block diagram of a long-haul, fiber-optic system. The nonreturn-to-zero (NRZ) input data stream, $x_1(t)$, is filtered (by the transmit filter[2] with frequency response $H_T(f)$) and the filtered signal, $x_2(t)$, modulates a single-frequency (distributed-feedback, DFB) laser, producing the signal $x(t)$ which modulates the electrical field. Alternatively, the data stream can be used to externally modulate the laser (to avoid laser nonlinearity). The transmitted optical signal passes through a fiber with frequency response $H_C(f)$, which may have chromatic and polarization dispersion. Chromatic dispersion will have a significant impact on system performance if the laser frequency is different from the dispersion minimum of the fiber (which is 1.3 μm in a standard fiber and 1.55 μm in a dispersion-shifted fiber). In this case, the dominant chromatic dispersion is linear delay distortion. At the receiver, the optical signal, $s_o(t)$, is converted to an electrical signal by a photodetector (usually an avalanche photodiode APD).

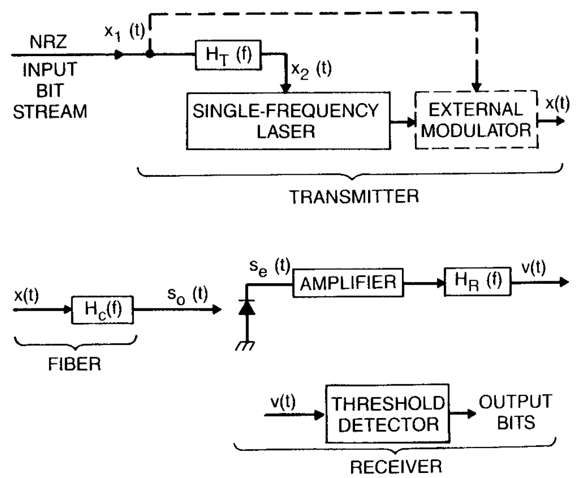

Figure 1 Block diagram of a long-haul, direct detection fiber-optic system.

With direct detection, as shown in Figure 1, the electrical signal, $s_e(t)$, is proportional to $|s_o(t)|^2$. Alternatively, coherent detection can be used (where the received optical signal is mixed with a local oscillator optical signal (at approximately the same frequency as the received signal) to generate an IF electrical signal whose *envelope* is proportional to $s_o(t)$.

Polarization dispersion can be characterized mainly in terms of first and second order (in frequency) effects [2]. The first order effect is a delay in the signal in one polarization relative to the delay in the signal in the other polarization. Thus,

1. Equalization techniques for multimode fiber systems have been studied previously for multimode [1] lasers.
2. This filter is used to reduce the high bandwidth components of the modulating signal and thereby reduce laser nonlinearity (chirp).

with direct detection and first-order polarization dispersion effects, since signals in orthogonal polarizations add powerwise at the receiver, the electrical signal is given by the *linear* combination of the individually detected signals,

$$s_e(t) = \alpha_1 \Big(|s_o(t)|^2 + \alpha |s_o(t+\tau)|^2 \Big) \quad , \qquad (1)$$

where α_1 is the conversion constant between the optical and electrical signals, α is the ratio of the signal strengths in the two polarizations, and τ is the time delay between propagation in the two polarizations.

Next, the electrical signal is amplified and filtered, and this signal is detected by comparing the signal level, during a short period of time at the peak opening of the eye of the signal to a decision threshold.

Intersymbol interference in the detected signal can be caused by laser nonlinearity, chromatic dispersion, polarization dispersion, and nonideal receiver frequency response. The laser nonlinearity and the receiver frequency response will vary among devices, and polarization dispersion will vary slowly over time (e.g., on the order of hours). Chromatic dispersion is reasonably fixed for a given length of fiber, but its effect on system performance depends on laser nonlinearity and receiver frequency response, which can vary. This suggests that adaptive signal processing structures may be required.

A key issue in the effectiveness of intersymbol interference compensation techniques, is the linearity[3] of the intersymbol interference. The linearity of the intersymbol interference determines whether linear or nonlinear equalization techniques are more appropriate.

III. COMPENSATION TECHNIQUES

A. Linear Equalization

To compensate for linear distortion, a linear equalizer (transversal filter) can be used between the receiver filter and the detector. In particular, we will consider the (analog) tapped delay line implementation of the equalizer[4] with N taps with tap weights c_i, $i=1,N$, as shown in Figure 2. Note that the

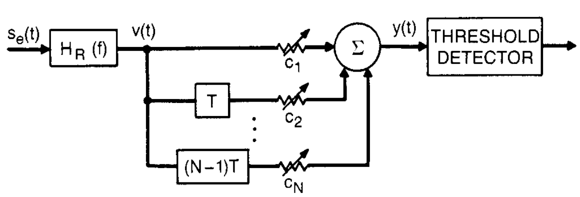

Figure 2 Tapped delay line for equalization with N taps.

equalizer output signal $y(t)$ is given by

$$y(t) = \sum_{j=1}^{N} c_j \, v\Big(t-(j-1)T\Big) . \qquad (2)$$

3. That is, whether or not the intersymbol interference in the electrical signal at the symbol detector can be considered as a superposition of intersymbol interference from symbols other than the symbol to be detected.

4. Such an equalizer has been implemented at 8 Gbps [3].

Since the receiver filter bandwidth is usually much less than the signal bandwidth (to reduce receiver noise), the tap spacing (T) need only be the symbol period, i.e., the equalizer is a synchronous linear equalizer (a fractionally-spaced equalizer is not needed in this case to reduce the ISI, although such an equalizer can reduce the equalizer noise enhancement). At high data rates, symbol delays can be implemented by a short transmission line (e.g., less than 4 cm at 8 Gbps), and the weights can be implemented by a variable-gain amplifier.

Since, in many cases (e.g., with laser nonlinearities such as chirp and relaxation oscillation), most of the intersymbol interference is due the symbols preceding the detected symbol, there may be more taps for the precursor symbols than for the future symbols. Other performance issues include weight adjustment techniques, the noise enhancement of the equalizer, and the linearity of the distortion. These issues are discussed below.

1. Weight Adjustment Techniques

The weights can be preset in the factory or set by craftsman during installation. With manual adjustment, the weights can be set to minimize the bit error rate or maximize the eye opening (system margin). However, manual adjustment to optimize these parameters may be difficult when there are more than a few weights, or when the optimum weights vary significantly among receivers (i.e., with variations in device characteristics). In addition, manual adjustment cannot be used to compensate for variations in devices over time (e.g., with temperature) or to track polarization dispersion. In these cases, some type of adaptive algorithm must be used.

Two celebrated adaptive algorithms are the least-mean-square (LMS) algorithm of Widrow [4] and the zero-forcing algorithm of Lucky [5,6]. Here we will only consider the zero-forcing algorithm. Figure 3 shows an adaptive equalizer using

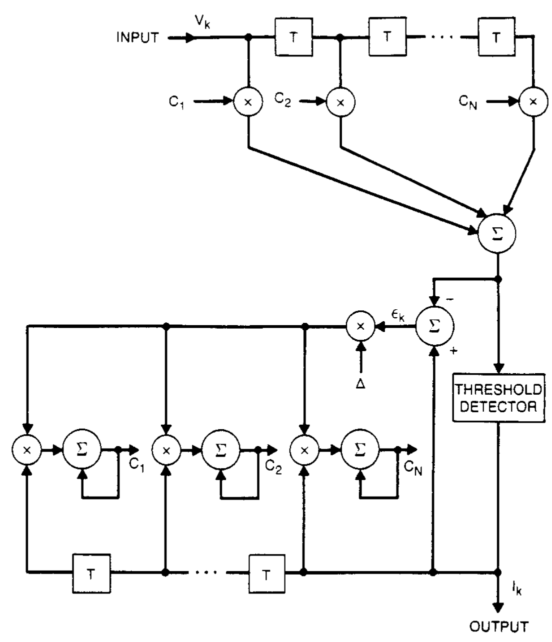

Figure 3 Adaptive linear equalizer using the zero-forcing algorithm.

this algorithm. For the zero-forcing algorithm, the weight update equation is given by

$$c_j^{k+1} = c_j^k + \Delta\ \epsilon_k\ I_{k-j} \quad , \quad k=1,2,... \tag{3}$$

where c_j^k is the j^{th} tap weight during the k^{th} symbol period, Δ is a small constant that controls the magnitude of the weight adjustment, and ϵ_k is the error in the equalizer output signal y_k given by

$$\epsilon_k = \left(I_k - y_k\right) \quad , \tag{4}$$

where I_k is the k^{th} output symbol. This algorithm adjusts the weights to minimize the peak distortion, but does not necessarily minimize the bit error rate or maximize system margin. The zero-forcing algorithm is only effective at minimizing peak distortion when the unequalized eye is open. With modest amounts of linear ISI and thermal noise, this equalizer produces distortion-free outputs and provides performance close to the optimum.

At the high data rates of long-haul systems, generating analog samples of signals is very costly and, hence, may not be practical[5]. Thus, single bit accuracy samples (± 1) must be used whenever possible. An algorithm that provides single-bit accuracy is the modified zero-forcing algorithm [5,7]

$$c_j^{k+1} = c_j^k + \Delta\ \mathrm{sgn}(\epsilon_k)\ I_{k-j} \quad . \tag{5}$$

Quantizing the signal samples reduces the rate of convergence of the algorithm (which is not a major concern, since channel impairments should change very slowly with time - on the order of hours or longer). The algorithm converges to the same weights as the continuous version (3) (when the eye of the received signal is open [neither algorithm works when the eye is closed]).

Figure 4 shows a possible implementation of the quantized (discrete) zero-forcing adaptive algorithm. The detected bits, the I_k's, are used to adjust the threshold of a second detector that compares the received signal samples to the predicted levels for ones and zeroes (i.e., determines the signum of the error). An even simpler technique (as suggested by D. G. Duff) is to set the decision threshold in the second detector to that for the predicted level for a one, and to only use the signum of the error when a one is detected. This eliminates the need to vary the decision threshold in the second detector. Once the $\mathrm{sgn}(\epsilon_k)$ and I_k values have been determined, the multiplications (by ± 1, i.e., sign inversion) and additions required for weight adaptation can be done at a much slower rate (e.g., by a microprocessor). Thus, the only costly hardware for the adaptive algorithm is the second detector.

The performance criterion we will consider in this paper is the optical signal power penalty due to intersymbol interference (i.e., the increase in received optical signal power required to maintain the same eye opening with intersymbol interference), which can be derived from the minimum eye opening over all input bit sequences. The minimum eye opening is the minimum sampled signal value for a "1" minus the maximum sampled signal value for a "0", with no noise at the receiver. Thus, if the difference between the signal levels for a "1" and a "0" without ISI is Y, the minimum eye opening (in percent) is given by

5. Since the analog sample of the error ϵ_k, gives the eye opening, it is also useful for determining the system margin and showing degradations before bit errors occur. Thus, analog sampling of the error may be worthwhile, even if it is not required for weight adaptation.

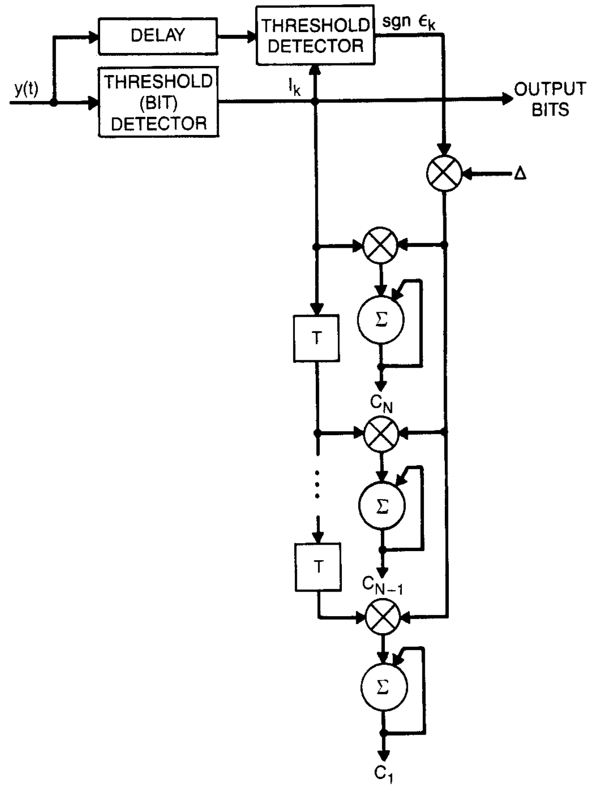

Figure 4 Implementation of the zero-forcing adaptive algorithm.

$$eye\ opening = \frac{\underset{k}{\mathrm{Min}}\ (y_k/I_k{=}1) - \underset{k}{\mathrm{Max}}(y_k/I_k{=}0)}{Y} \cdot 100 \tag{6}$$

or

$$eye = \underset{\substack{k,i \\ I_k=1 \\ I_i=0}}{\mathrm{Min}} \left(\frac{y_k - y_i}{Y} \right) \cdot 100 \quad . \tag{7}$$

The optical power penalty is given by

$$penalty = \begin{cases} 10\ \log_{10}(eye/100)\ \mathrm{dB} & \text{for direct detection} \\ 20\ \log_{10}(eye/100)\ \mathrm{dB} & \text{for coherent detection} \end{cases} \tag{8}$$

since the received current is proportional to the optical power with direct detection and the received current is proportional to the magnitude of the optical field with coherent detection.

2. Noise Enhancement

Since the linear equalizer combines the weighted received signals, the noise level in the output signal may be increased if the signal is amplified over a range of frequencies (to compensate for attenuation over the frequency range). In this paper, we will not consider the absolute level of the noise at the receiver, but only consider the relative increase in the noise level due to equalization (referred to as noise enhancement). If we assume that, at the output of the receiver filter, samples of the noise taken every T seconds are independent, zero-mean, Gaussian random variables with variance σ^2, then the noise enhancement of the equalizer is just $\sum_{j=1}^{N} c_j^2$. A measure of the relative received signal-to-noise ratio (SNR), reflecting both ISI and noise, is given by (in percent), from (7),

$$\rho = \underset{\substack{k,i \\ I_k=1 \\ I_i=0}}{\text{Min}} \left[\frac{\left(y_k - y_i\right)}{Y\sqrt{\sum_{j=1}^{N} c_j^2}} \right] \cdot 100 \quad , \qquad (9)$$

with the optical power penalty being 10 $\log_{10}(\rho/100)$ dB for direct detection and 20 $\log_{10}(\rho/100)$ dB for coherent detection.

The noise enhancement can be reduced or eliminated by using a fractionally-spaced equalizer (decision-feedback equalization or linear cancellation can also be used). In general, a tap spacing of $T/2$ is adequate for reducing the noise enhancement as much as possible. However, with chirp, tap spacings may need to be closer than the reciprocal of the chirp bandwidth, i.e., tap spacings of less than $T/5$ may be required to significantly reduce the noise enhancement. Therefore, a five-fold increase in the number of taps could be required to eliminate ISI without noise enhancement. To simplify our analysis and keep the number of taps small, we will only study synchronous (tap spacings of T) equalizers.

B. Nonlinear Cancellation

Since there are cases when the distortion is nonlinear (e.g., direct detection with chromatic dispersion), nonlinear techniques must be used to significantly reduce distortions. Here, we consider the use of nonlinear cancellation (see [8]). Nonlinear cancellation can be implemented as follows. Using knowledge of previously detected bits and, perhaps, estimates of bits to be detected, the decision threshold in the detector is adjusted up or down to be halfway between the expected signal levels for each bit to be detected. If the adjustment is a linear sum of the previous and estimated bits, then the technique is just linear cancellation [9,10] (or decision-feedback equalization if only previously detected bits are used). Otherwise, a lookup table (or explicit computation[6]) may be used, with 2^{N-1} entries for $N-1$ previously detected and estimated bits (a total of N bits are used to determine the data bit) to provide nonlinear cancellation. At the high data rates of the long-haul systems, the size of the lookup table may limit $N-1$ to just a few bits (fortunately, this is typically the extent of the ISI in high-speed lightwave systems).

Since estimating bits to be detected requires an additional detector (or even additional interference reduction techniques if the eye is closed), the most practical technique is to adjust the decision threshold based on previously detected ("decided") bits only, and use an analog tapped delay line (which has been implemented at 8 Gbps [3]) to reduce distortion caused by "future" bits. This canceler is shown in Figure 5, where N_1

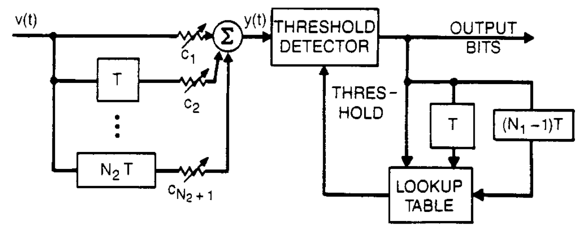

Figure 5 Nonlinear canceler using N_1 previous bits and a tapped-delay line for reduction of intersymbol interference from N_2 future bits.

decided bits are used for nonlinear cancellation and a tapped delay line with N_2+1 taps is used for the N_2 future symbols plus the data bit to be detected. Thus, at the detector,

$$y(t) = \sum_{i=1}^{N_2+1} c_i v(t-(i-1)T) \quad , \qquad (10)$$

and the decision rule is

$$I_k = \begin{cases} 1 & \text{if } y(t_0) > y_0 + f(I_{k-1},\cdots,I_{k-N_1}) \\ 0 & \text{otherwise} \end{cases} \quad , \qquad (11)$$

where y_0 is the unadjusted decision threshold, t_0 is the sampling time, and $f(\cdot)$ is the output of the look-up table which provides an estimate of the (nonlinear) intersymbol interference. This type of nonlinear canceler can only reduce nonlinear distortion that is caused by the N_1 decided bits. However, this is the main form of the distortion with chirp (since laser chirp produces an increase in the fall time of the pulse).

The adaptive algorithms for nonlinear cancellation are a generalization of the algorithms for linear equalizer updating.

In the remainder of the paper, we will consider nonlinear cancellation using both future as well as decided bits, rather than using a tapped delay line for the future bits. This simplifies our analysis and results, and gives an upper bound on the performance improvement with a tapped delay line for future bits.

The performance of nonlinear cancellation is given by the minimum distance between sampled signal values for a "1" and a "0", given the same N_1 decided and N_2 future bits. That is, for nonoverlapping levels for "1"'s and "0"'s for each set of N_2 future and N_1 decided bits, the minimum percent eye in the detected signal is given by,

6. In practice, the lookup table could be implemented by a switch with the previously detected and estimated bits controlling which one of 2^{N-1} voltage levels is connected to the threshold.

$$eye = \min_{\substack{k,i \\ I_k=1 \\ I_i=0 \\ \{z_k\}=\{z_i\}}} \left[\frac{|y_k - y_i|}{Y} \right] \cdot 100 \quad , \qquad (12)$$

where $\{z_j\}$ is the set $\{I_{j-N_1}, \ldots I_{j-1}, I_{j+1}, \ldots I_{j+N_2}\}$ which is the set of the $N-1$ $(=N_1+N_2)$ bits used in the lookup table.

Nonlinear cancellation will fail (i.e., the error rate will be nonzero even without noise at the receiver) if both $\min_{\substack{k,i \\ I_k=1 \\ I_i=0 \\ \{z_k\}=\{z_i\}}} \left[y_k - y_i\right] \leq 0$ and $\max_{\substack{k,i \\ I_k=1 \\ I_i=0 \\ \{z_k\}=\{z_i\}}} \left[y_k - y_i\right] \geq 0$, i.e., the levels for "1"'s and "0"'s overlap for some set of N_1 decided and N_2 future bits. To see where this could occur, consider the simple example of direct detection, where the sampled values of the optical signal with intersymbol interference are given by (using baseband notation)

$$s_o(kT) = I_k - .5\, I_{k-1} \; . \qquad (13)$$

Thus, neglecting the receiver filter and any noise,

$$y_k = s_e(kT) = |s_o(kT)|^2 = I_k^2 - I_k I_{k-1} + .25\, I_{k-1}^2 \qquad (14)$$

or

$$y_k = \begin{cases} 0, & I_{k-1}=0,\ I_k=0 \\ 1, & I_{k-1}=0,\ I_k=1 \\ .25, & I_{k-1}=1,\ I_k=0,1 \end{cases} \; . \qquad (15)$$

Note that when $I_{k-1} = 1$, the signal level y_k is independent of I_k, and, therefore, the output levels cannot be separated and nonlinear cancellation will not work. However, we can see that I_k can be determined from y_{k+1} with a 3-level detector, i.e., a reasonable decision rule is

$$I_k = \begin{cases} 1 & \text{if } .125 \leq y_{k+1} \leq .625 \\ 0 & \text{otherwise} \end{cases} \quad , \qquad (16)$$

This leads us to consider the optimum detection scheme, maximum likelihood detection, in the next section.

C. Maximum Likelihood Detection

Maximum likelihood detection (MLD) is the optimum detection technique in that it minimizes the error probability for determining a bit (or bit sequence), given N received signal samples. It can be complex to implement for large N, although it is useful as a bound on performance. There are techniques to implement simplified versions of maximum likelihood detection, however, that may be practical at high data rates, if N is small. Here we present one such technique. As before, we assume that the received signal levels and intersymbol interference are deterministic (given the bit pattern), and that the only source of noise is additive Gaussian noise (thermal noise with direct detection and high-intensity shot noise with coherent detection).

The technique uses a sliding window, with N_1 previously detected bits used to determine the state (i.e., one of the 2^{N_1} possible cases that exist prior to transmission of the next N_2+1 signal samples), and N_2+1 signal samples used to determine the detected bit $(N=N_1+N_2+1)$. Specifically, the detector calculates the Euclidean distance between the received signal vector of length N_2+1 and each of the 2^{N_2+1} stored signal vectors to determine the stored signal vector that is closest to the received signal vector and outputs the bit corresponding to the first bit in that stored signal vector. There is a separate set of 2^{N_2+1} stored signal vectors for each of the possible 2^{N_1} states, for a total of $2^{N_1+N_2+1}$ (or 2^N) signal vectors. Figure 6 shows a block diagram of this technique.

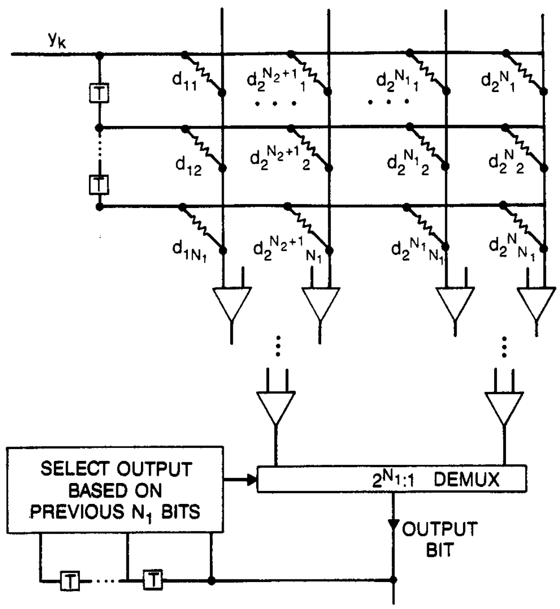

Figure 6 Maximum likelihood detector using N_1 previous bits and an N_2+1 bit signal vector to determine a single bit.

The performance of the sliding window detector can be determined from the minimum Euclidean distance between received signal vectors with the same N_1 decided bits and different first bits, followed by any combination of N_2 future bits. That is, the minimum eye opening is

$$eye = \min_{\substack{k,i \\ I_k \neq I_i \\ \{I_{k-N_1} \cdots I_{k-1}\} \neq \{I_{i-N_1} \cdots I_{i-1}\}}} \left[\frac{\sqrt{\sum_{j=0}^{N_2+1} (y_{k-j} - y_{i-j})^2}}{Y} \right] \cdot 100 \qquad (17)$$

The detector of Figure 6 can be made adaptive by channel estimation (when the detected bits are virtually error free). In addition, the maximum likelihood detector can be reduced in size and made adaptive through the use of neural network techniques, in particular, with a 3-layer feedforward network using backpropagation for adaptation [11].

IV. RESULTS

The results in this section were generated, via computer simulation, as follows. The data rate was 8 Gbps, and the transmit filter was a low pass RC filter with a 3 dB bandwidth of 4 GHz for the cases with direct modulation and 32 GHz (i.e., the transmit filter can be neglected) in the other cases. Laser characteristics were obtained from [12], with the transmitted waveform generated by programs as described in [13]. The data rate and transmit filter bandwidth are similar to those studied in [13], and the laser parameters are measured data from the most recently available DFB lasers. The programs in [13] use a repetitive pseudorandom data stream of length 64, which contains all bit sequences of length 6. Thus, the results should be accurate as long as the intersymbol interference extends over only a few bit periods. As stated in Section I, the results generated using these parameters demonstrate the typical improvement that can be obtained with the techniques of Section III. However, the improvement in other systems (e.g., with different data rates, laser characteristics, etc.) could vary.

A. Chromatic Dispersion

Let us consider the effect of the various impairment-reducing techniques on chromatic dispersion. Figure 7 shows the effect of equalization techniques with chromatic dispersion and laser nonlinearity. Results are shown for LE, NLC, and MLD with $N=6$, and a 3-pole Butterworth (nearly ideal) receiver filter. Note that the combined effect of chromatic dispersion and laser nonlinearity produces a dip in penalty above 40 km, similar to that shown in [13], where different DFB laser characteristics were used. However, the power penalty in this region is much higher than in [13], showing that the more recently available DFB lasers are more affected by chromatic dispersion (although other characteristics of these lasers are much improved). Although chromatic dispersion produces nonlinear intersymbol interference at the receiver, LE still reduces the power penalty somewhat - by at least 1.5 dB for distances above 40 km (or equivalently increases the maximum distance for a given penalty by about 20%)[7]. NLC and MLD, however, decrease the penalty by more than 3 dB for distances above 40 km and substantially increase the dispersion-limited transmission distance (to beyond 300 km)[8]. Here, the best N consecutive samples included at most 1 future bit sample, which simplifies implementation of NLC.

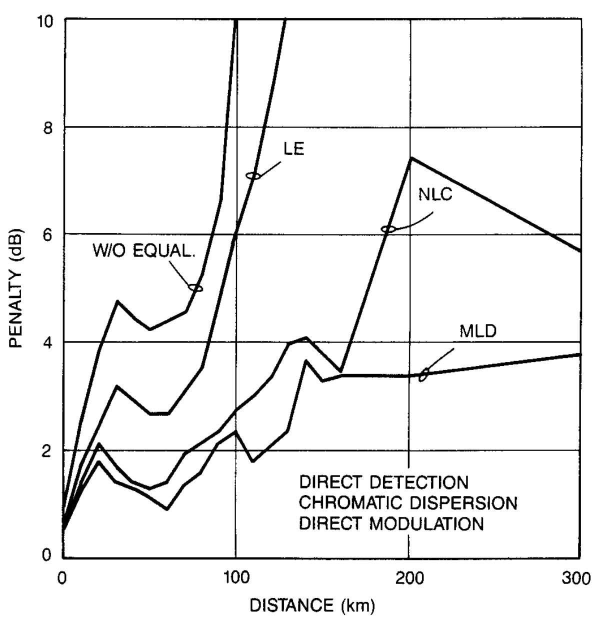

Figure 7 Effect of equalization techniques with chromatic dispersion and laser nonlinearity - optical power penalty versus distance with LE, NLC, and MLD for $N=6$.

Figure 8 shows the effect of equalization techniques with chromatic dispersion when laser nonlinearity is not present (i.e., with external modulation). As before, LE decreases the penalty (by more than 1.5 dB above 160 km) or, alternatively, increases the dispersion-limited distance (by 25% for a 3 dB penalty)[9]. Although with LE the penalty dips above 240 km, this effect is highly dependent on the transmit pulse shape and receiver frequency characteristics and, therefore, may not necessarily be present in systems with slightly different characteristics. On the other hand, NLC and MLD do not have such variations. These techniques greatly reduce the penalty above 200 km, increasing the dispersion-limited distance for a 3 dB penalty to 270 and 400 km with NLC and MLD, respectively.

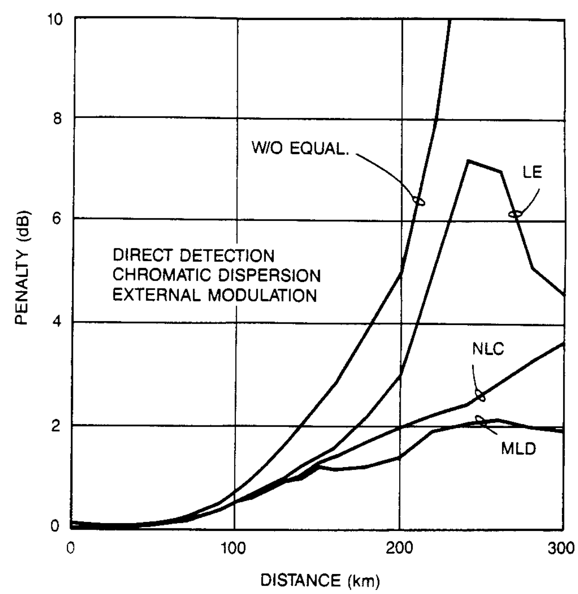

Figure 8 Effect of equalization techniques with chromatic dispersion without laser nonlinearity in direct detection systems.

B. Polarization Dispersion

Finally, let us consider the effect of the various impairment reducing techniques on first-order polarization dispersion effects. Because of its simple form, first-order

7. A fractionally-spaced equalizer may do even better.
8. Distances above 200 km are only feasible because of the advent of optical amplifiers.
9. Again, a fractionally-spaced equalizer may do even better.

polarization dispersion effects can easily be eliminated by NLC or MLD using at most 3 signal samples, with the available adaptive algorithms for NLC cancellation [8]. However, since linear equalization has been shown to be effective against nonideal receiver frequency response and chromatic dispersion, and, unlike NLC and MLD, has been implemented at Gbps rates [3], let us consider linear equalization, which, as shown in Section III, can easily be adaptive.

The effect of polarization dispersion depends on α (the ratio of signal powers in the two polarizations) and τ (the time delay between the two polarizations). The worst case for α occurs when $\alpha = 1$. In this case the eye is closed when τ is a multiple of the bit period (125 psec), and a LE will not open the eye. However, for all other values of α and τ, LE reduces the penalty. Figure 9 shows the effect of LE (with 6 taps) for

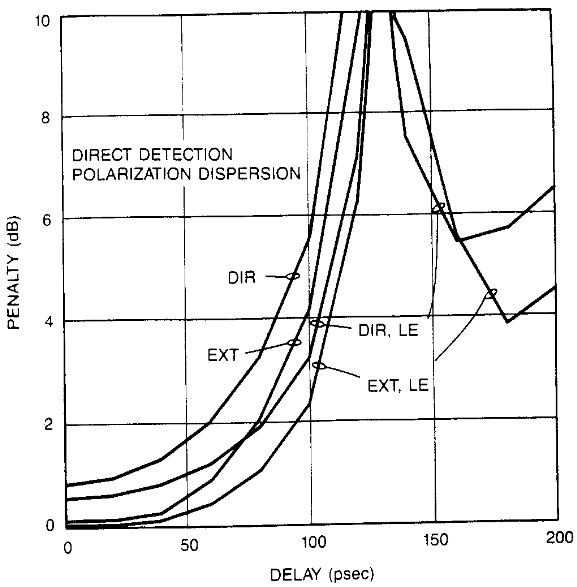

Figure 9 Effect of linear equalization with polarization dispersion (worst case) in direct detection systems.

$\alpha = 1$, with direct and external laser modulation. LE reduces the penalty by more than 1 dB for $\tau > 70$ psec and increases the tolerable delay for penalties greater than 3 dB by about 10% (this corresponds to a 20% distance increase since the delay varies as the square root of the distance). Note also that, for delays greater than the bit period, LE reduces the penalty to as low as 4 dB. Of course, for $\alpha \neq 1$, LE reduces the penalty to even lower values[10].

V. SUMMARY AND CONCLUSIONS

In this paper, we have studied the various impairments in high speed lightwave systems, presented techniques to reduce these impairments, and analyzed their performance in a typical system. The impairments include laser nonlinearity, chromatic and polarization dispersion, and nonideal receiver filtering. The techniques include linear equalization, nonlinear cancellation, and maximum likelihood detection. Methods for implementing these techniques, including adaptive linear equalization, were presented. Computer simulation results for an 8 Gbps system using measured laser and receiver characteristics showed that in direct detection systems a 6-tap linear equalizer can reduce the penalty due to chromatic and polarization dispersion by more than 1 dB (or increase the dispersion-limited distance by more than 20%). Nonlinear cancellation and maximum likelihood detection can reduce the penalty even further, more than doubling the dispersion-limited distance in some cases.

10. In addition, a fractionally-spaced equalizer may reduce the penalty even further.

VI. ACKNOWLEDGEMENT

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