Fig. 2 shows calculated sensitivity using Q factor estimation [1]. Indicated with lines are theoretically analyzed sensitivities as a function of bit rate without an SOA, with an SOA, and with an SOA and an optical band-pass filter, respectively. Our experimental result with SOA demonstrated later, a previous work with SOA [2], and the state-of-the-art with APD [3-7] are also plotted in the figure.

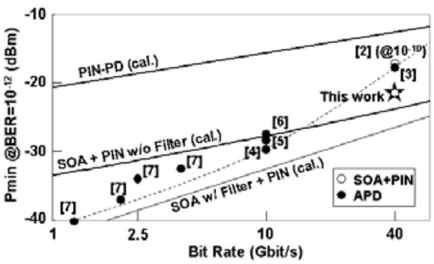

Fig. 2. Calculated minimun sensitivity Pmin of SOA. Reported measured sensitivity of SOA and APD is also plotted. Pmin is defined as the optical power at BER = 10^{-12} unless otherwise mentioned. Dashed line is a fitting of reported sensitivity of APD.

Up to 10 Gbit/s, APD receivers show the minimum sensitivity in inverse proportion to the bit rate, which would be reasonable given that in the APD receiver where the signal-dependent noise is dominant, the electrical SNR is proportional to the incident optical signal power and is inversely proportional to the electrical bandwidth. 40-Gbit/s APD receivers get out of this relation due to insufficient multiplication factor and, as a result, show lower sensitivity than SOA receivers without an optical band-pass filter.

Fig. 2 also indicates that the effect of ASE filtering upon the receiver sensitivity becomes smaller in higher bit rate region. At 40 Gbit/s, the sensitivity improves by 2.7 dB with a filter whereas at 10 Gbit/s it does by 4.8 dB and at 2.5 Gbit/s, by 7.3 dB. Optical-filterless 40 Gbit/s receivers provide operation in the entire C-band, in exchange for 2.7-dB sensitivity degradation.

3. Experiment and Result

The experimental setup of the receiver sensitivity measurement is depicted in the Fig. 3. An NRZ 40 Gbit/s optical signal is generated by external modulation of CW-laser (1550 nm wavelength), with a $LiNbO_3$ modulator. The modulator is driven with 2^{31}-1 PRBS signal from 4 x 10G PPG and a 4:1 multiplexer. The receiver consists of an SOA, a PIN-photodiode, a transimpedance amplifier (TIA), and a post-amplifier. The output of the SOA was directly connected to the PIN-photodiode. The bit error rate was measured with a 1:4 demultiplexer and 4 x 10G BER testset.

The optical signal waveform at the output of the modulator and the SOA are shown in the Fig. 4. The degradation of eye opening caused by the SOA was not significant. At the modulator output, the extinction ratio was 11.1 dB and at SOA output 10.8 dB. 1.2 ps of rms jitter at the modulator output did not increase at the SOA output.

Fig. 5 shows the measured bit error rate curves. The measured minimum sensitivity with 2^{31}-1 PRBS was -21.3 dBm with an SOA and -11.4 dBm at BER = 10^{-12} without an SOA. The sensitivity improved by 9.9 dB using 14-dB gain SOA without an optical band-pass filter. This is the highest sensitivity ever reported. The difference of sensitivity between 2^{31}-1 and 2^7-1 PRBS was less than 0.2 dB. The polarization dependent gain of the SOA used in this experiment was 0.5 dB. BER was stable during the experiment.

Considering the finite gain of the SOA and the circuit noise of the TIA, we calculated theoretical sensitivity -23.1 dBm with an SOA and -12.6 dBm without an SOA. We suppose that the discrepancy between measured and calculated sensitivity is chiefly because of the inter-symbol interference and the ambiguity in decision, which were not included in the calculation model.

We confirmed that the receiver worked within the wavelength range of 1530 nm to 1570 nm, which covers the full C-band. The SOA in this experiment has 70-nm ASE bandwidth, which suggests that it has the potential for the operation around 70 nm, broader than the EDFA.

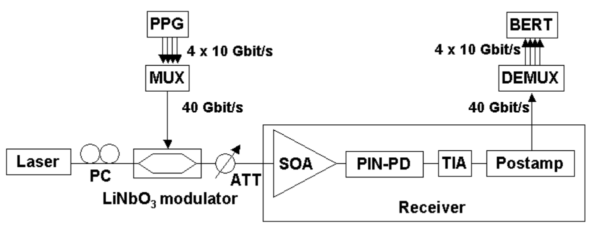

Fig. 3. Experimental setup. PPG: Pulse Pattern Generator, BERT: Bit Error Rate Testset, PC: Polarization Controller, ATT: Optical Attenuator, SOA: Semiconductor Optical Amplifier, TIA: Transimpedence Amplifier.

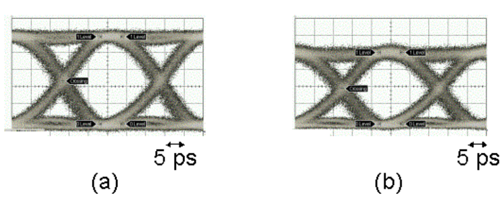

Fig. 4. Eye pattern of (a) modulator output and (b) SOA output. The vertical axes are in arbitrary unit.

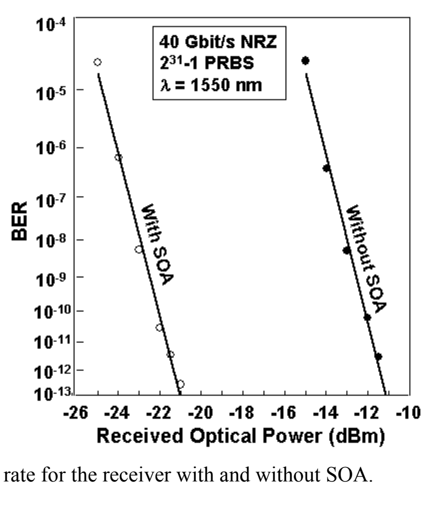

Fig. 5. Measured bit error rate for the receiver with and without SOA.

4. Conclusion

We have presented the possibility of the SOA equipped high sensitivity 40 Gbit/s receiver both theoretically and experimentally. Unlike the APD, this device is not restricted by the gain bandwidth product essentially. The measurement has proved a record-high -21.3 dBm of the sensitivity at BER = 10^{-12}. We believe that this type of receiver is one of the prospective candidates for optical receivers in the short-reach, high-speed communication systems.

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ThG4 **9:30 AM**

Improved Performance of 10 Gb/s Multimode Fiber Optic Links Using Equalization

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A differential 7-tap transverse filter was designed and used to equalize an 850nm multimode link. The ISI penalty was reduced from 11dB to 2.2dB on a 10Gb/s link with 600m of 50µm next-generation multimode fiber.

Introduction

Multimode fiber (MMF) is dominant in the LAN and SAN environment (Ethernet, Fibre Channel). At speeds of 10 Gb/s and beyond, the achievable distances in this environment become impractically short. For example, the recently adopted IEEE 802.3ae standard (10 Gb/s Ethernet) includes a short wavelength (850nm) physical media dependent layer. The achievable distances using 50 µm multimode fibers (MMF) are up to 86m for standard FDDI grade fiber and 300m using the next generation MMF. The achievable distances on 62.5 µm fiber are much shorter. For comparison, for Gigabit Ethernet these distances were 550m on standard FDDI grade 50 µm MMF. Many customers with existing infrastructure that supported up to 550m on Gigabit Ethernet want to reuse the fiber for 10 Gb/s transmission. Similarly, on a standard FDDI 50 µm grade fiber, cus-

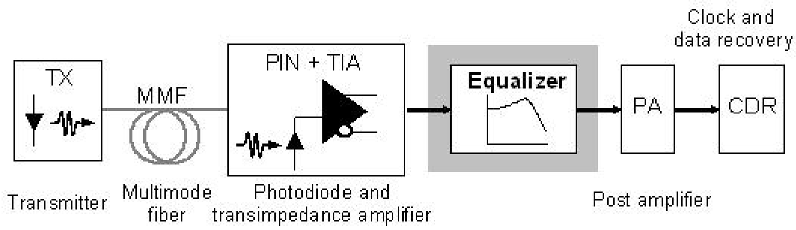

Figure 1: Typical multimode link for LAN applications, with an equalizer inserted after the trans-impedance amplifier.

tomers want to achieve at least 100m. Using next generation fiber, the desired reach for very short reach applications (VSR) is at least 600m. While it is possible to reach those distances by simultaneously implementing more stringent requirements on the fiber DMD profile and the laser encircled flux [1], as well as the receiver bandwidth, these solutions become very costly. Furthermore, for existing installations, the limitation is due to the fiber bandwidth, which may not be possible to be overcome by improvement in the receiver (bandwidth, sensitivity), laser transmitter rise time and/or a better-controlled fiber launch.

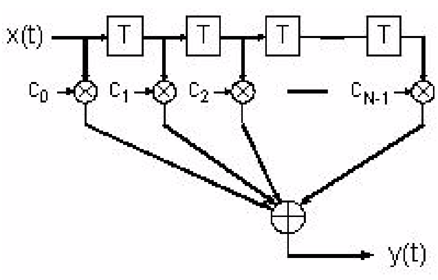

Figure 2: High-level schematic of the transverse filter.

These extended distances can be achieved by using equalization techniques in the link. In the past, equalization has been reserved for long distance telecommunication links. However, their use in LAN links has been considered by the IEEE 802.3ae and its ad hoc committee on equalization. Furthermore, equalization for multimode links has been proposed in [2]. In this paper we report on the design and performance of a 7-tap transversal filter equalizer built in 0.18 µm SiGe-BICMOS technology to reduce the ISI penalty due to inter-modal dispersion.

Equalizer Design

At low speeds, equalization is often achieved through a digital decision feedback equalizer (DFE) using memory or shift registers as the delay elements. As data rates approach 10Gb/s, digital techniques are difficult to implement and analog implementations or RF techniques are required. In analog designs, sampling is done after equalization, as shown in Figure 1. This has the advantage that signal delay in the equalizer does not affect the performance and stability of the clock and data recovery circuits.

In the design stage we investigated several different equalization schemes. Simulations to assess the performance of the equalizer were done using data from 6 transmitters and twelve 300m long fibers distributed by the Telecommunications Industry Association for round robin testing to finalize the specifications for the next generation MMF. The data set was enlarged by changing the launch conditions for each transmitter by introducing offset and measuring the signal at the link output. Measurements for 21 offsets for 6 lasers and 12 fibers (or more than 1500 links) were available to simulate the behavior of each equalizer type and assess its performance. We looked into the distance improvement factor versus complexity, number of stages and delay per stage (Figure 3) and concluded that a 7 tap filter with a 50 ps delay per stage would be adequate. In Figure 4 we show the percentage of links meeting the link power budget vs. distance. Due to the variety of laser launches and fibers, the distance improvement factor is between 1.5 and 2 for the analog forward equalizer, and between 2 and 3 for a digital DFE equalizer. More than 80% of all links would work with the forward equalizer at 600m, and nearly all links with a DFE equalizer, compared to only 6% without equalization.

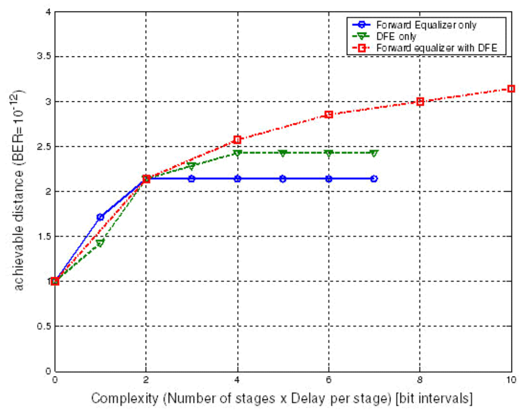

Figure 3. Comparison of tradeoff between equalizer complexity (number of stages, delay per stage) and achievable distance improvement for forward equalizer, DFE only and forward equalizer with DFE.

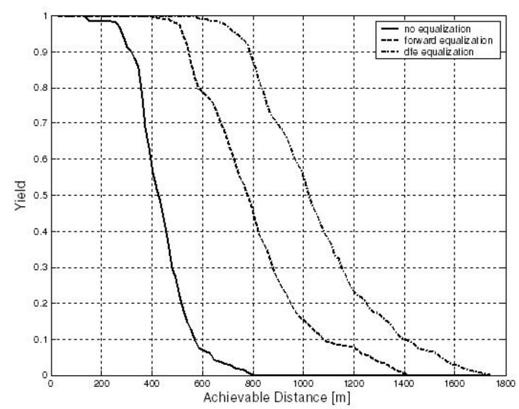

Figure 4. Percentage of links meeting link power budget versus achievable distance for links without equalization, with forward equalization, and DFE.

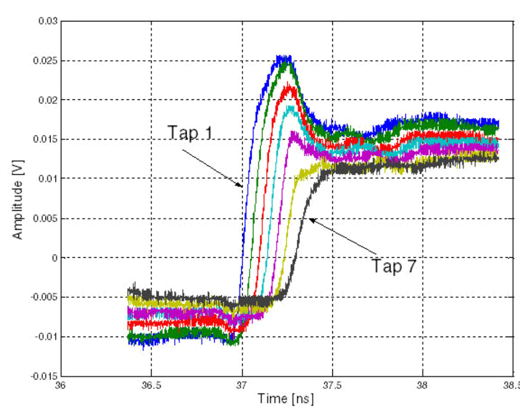

Figure 5. Step response of each stage in the 7-tap filter. The delay between stages is approximately 50 ps for each stage.

Based on these simulations, we designed a forward equalizer consisting of a differential 7-tap transversal filter [3]. A signal flow diagram of the design is shown in Figure 2. Since transmission lines implemented as microstrip lines or coplanar waveguides would be unreasonably long for a 50ps delay, the delay elements are implemented as LC ladders consisting of spiral inductors and MIM capacitors. Careful design of these structures is of critical importance [4]. The filter tap coefficients have both gain and sign controls. The gain control changes the gain of each differential stage. The sign control is used to swap the signal currents in the differential stage. Care was taken in the design to minimize the loading and load variation on the input and output transmission lines. This chip was fabricated in 0.18µ BiCMOS SiGe with f_T = 120 GHz. The area of the 7-tap device is 1.0 mm X 2.5 mm.

Results

Figure 5 shows the measurement of the step response of the 7-tap filter, when only one of the

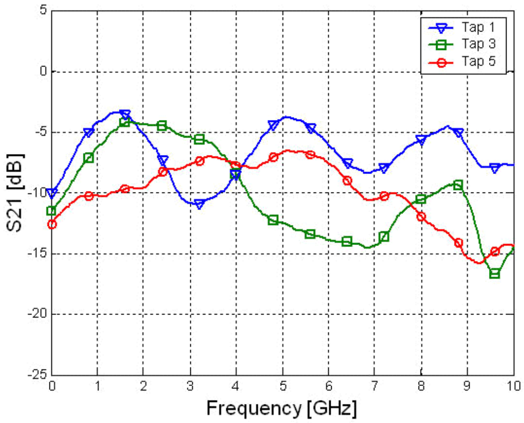

Figure 6. Frequency response of 3 taps of the filter.

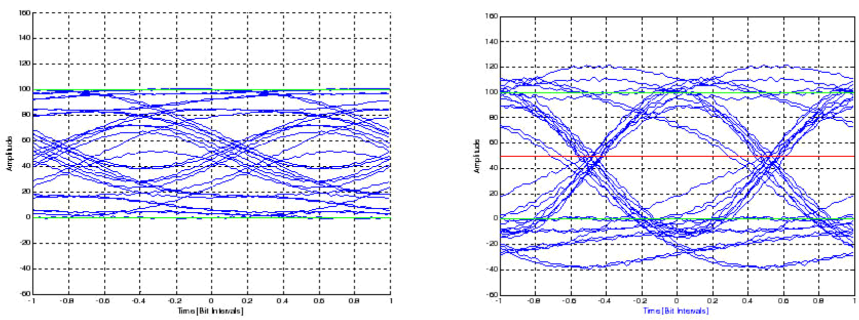

Figure 7. Averaged eye diagrams at the a) input and b) output of the equalization chip. The signal propagated through a 600 m long link and the ISI penalty at the link output was 11 dB. The ISI penalty was reduced to 2.2 dB after the equalization. At the same time the timing jitter was reduced from 60ps to 21 ps.

TF coefficients is non-zero. The plots show an approximate delay per stage of 50 ps. The transfer from the earlier stages also has more high-frequency peaking and less attenuation than the transfer from later stages. Figure 6 shows the frequency response of the 7-tap filter for each of the filter stages, obtained using a network analyzer. The bandwidth of the filter is adequate for bandwidth limited 10 Gb/s signals. Additional tests showed the output signal amplitude to be a linear function of the coefficient voltage. The signals through different taps were also shown to add linearly. The total power dissipation, including all biasing circuits, is 30 mW, plus 2 mW per active coefficient (typical max dissipation is 40 mW).

In order to test the actual equalization ability of the chip, a setup similar to the application environment shown in Figure 1 was used. A pattern generator was used to drive a vertical cavity surface emitting laser (VCSEL), and the light was launched into a 600 m long 50µm non-compliant next generation multimode fiber (not meeting the TIA-492AAAC specification). Since this was a non-compliant fiber, it wasn't able to achieve the 300m target distance under worst case conditions. The optical signal was detected using a commercially available optical receiver that had a 9GHz bandwidth. The pattern was set to be K28.5 8b/10b pattern and the data rate was 10 Gb/s. This generated a signal with 11 dB of ISI penalty (mostly due to intermodal dispersion) and 60ps of deterministic jitter at the input of the equalization chip, as shown in the eye diagram in Figure 7a.

To determine the proper coefficient weights and signs for the taps, we measured the responses of the individual stages as a function of the tap coefficient voltage. We then fitted a linear combination of the signals from each tap output to an idealized signal with a raised cosine pulse response. This process yielded the control voltages for each of the taps, and these were applied to the equalizer to produce the equalized signal shown in Figure 7b. The equalizer reduced the ISI penalty in the output signal from 11dB to 2.2 dB. This small ISI penalty is within the link power budget for 10 Gb/s Ethernet compliant multimode fiber links. Furthermore, the deterministic jitter was reduced from more than 60 ps to 21 ps after the equalization.

We checked the equalizer operation at 14 Gb/s, using the same 600m long multimode fibers as before, but with slightly different launch conditions, resulting in ISI penalty of 7.7 dB at the receiver output. This ISI penalty was reduced to 1.6 dB after the equalizer. The jitter remained the same after the equalization at 30 ps.

Conclusions

The design and performance of a 7 tap differential transverse filter equalizer are presented. The equalizer dramatically improved the performance of a 600m long multimode fiber link at 10Gb/s. The data shows that the equalization operation reduced both the vertical eye closure (ISI penalty) as well as the horizontal eye closure (timing jitter) in the signal. Significant ISI penalty reduction was achieved also at 14 Gb/s.

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ThG5 **9:45 AM**

Control of Combined Electrical Feed-Forward and Decision Feedback Equalization by Conditional Error Counts from FEC in the Presence of PMD

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Combined control of an adaptive 3-tap Feed Forward and 2-tap Decision Feedback Equalizer employing an adjustment of coefficients by means of conditional bit error ratio (BER) estimation based on Forward Error Correction (FEC) is proposed.

1. Introduction

Optical transmission at high data rates suffers from several degrading effects. Among those Polarization Mode Dispersion (PMD) has attracted much attention, due to its statistical nature as well as the dynamic behavior associated [1, 2]. Therefore adaptive equalization has been proposed like Adaptive Threshold (AT), Feed Forward Equalization (FFE), Decision Feedback Equalization (DFE) and Viterbi Equalization [3]. At the same time advanced Forward Error Correction (FEC) has become essential to reduce the required receiver optical input power or optical signal-to-noise ratio. A common method for getting proper FFE-coefficients is based on minimizing the mean square error between the signal before and after the DFE-threshold. This however requires at least the information about the sign of the analog signal on each tap, which is then combined with the current error-value at each sample and is therefore not feasible at very high data rates. One approach presented in [4] uses a separate eye monitor for optimizing the eye opening at the output of the FFE filter [5]. While at data rates of 10Gb/s those concepts appear to be close to commercial availability, implementation at 40Gb/s will suffer again from limitation of high-speed electronics The proposed method avoids any additional high-speed circuitry in the feedback control loop. Furthermore it intends to make use of upcoming improved FEC algorithms providing larger coding gain [6], from which higher dynamic range is available to support more complex control loop design. This paper presents a method to adapt the combination of a 3-tap FFE followed by a 2-tap DFE by means of estimated conditional BERs (i.e., error counts for different sequences of transmitted bits). Performance results for BER-based adaptations using conditional error rates are presented for simultaneous adjustment of FFE and DFE coefficients as well as AT, FFE or DFE individually.

2. BER-Based Adaptive FFE and DFE

Consider the received electrical noisy signal after the photo diode which is first feed into a *T*-spaced FFE filter, then by means of a proper clock and data recovery circuitry *T*-spaced sampled and finally equalized using a generalized DFE (Fig. 1) [7]. The FFE filter output is described by $r'(t) = \Sigma_i\, g_i\, r(t-iT)$, where T denotes the bit period, $r(t)$ the received input signal and $g_0,..., g_{L-1}$ the filter coefficients. Assuming a stable sample clock the sampled values in front of the DFE calculates to $r'_k = r'(kT+\varphi) = \Sigma_i\, g_i\, r(kT-iT+\varphi)$, where we have to introduce the sampling phase φ. Proper choice of the sampling phase significantly affects the resulting performance and can be fulfilled also by using BER-based control loops [8]. In the following, we assume a binary transmission with symbols '+1' and '-1' representing the data. Then one coefficient e.g. that one corresponding to the bit to be decided can be fixed. The remaining coefficients have to be adapted.

The DFE operation in its most general form can be done using a finite state machine which is then capable of equalizing nonlinear intersymbol interference too, see Fig. 1. The basic idea of a com-