3Gbit/s Transmission over Plastic Optical Fiber with Adaptive Tomlinson-Harashima Precoded Systems

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***Abstract*—This paper presents a 3 Gbit/s transmission scheme for an automobile optical physical layer, which is based on plastic optical fibers (POF) and low cost light emitting diodes (LED). Tomlinson Harashima Precoding (THP) and an adaptive feedforward equalizer (FFE) are used to cope with the strong inter-symbol-interference (ISI). By computer simulations, the system is proven to be a cost-effective and reliable solution for high-speed in-car communications.**

I. Introduction

The up-to-date in-car infotainment backbone MOST150 (the Media Oriented Systems Transport 150) offers a data-rate of 150 Mbit/s in its optical physical layer. The future MOST systems, however, are expected for a 2 $\sim$ 3 Gbit/s transmission capability to serve various automobile applications, like side and back cameras or driver assist applications. Meanwhile, a smooth upgrade from the current MOST150 is required due to cost reasons, i.e., the car-makers can still use plastic optical fibers (POF) and low cost light emitting diodes (LED) in the MOST physical layer [1]. Since the limited bandwidths of POF and LED cause very strong inter-symbol-interferences (ISI) during multi-Gbit/s transmissions, advanced signal processing techniques are necessary for compensating the ISI.

There are several papers on this topic proposing solutions utilizing either an analog prefilter/peaking, or an advanced modulation scheme like discrete multitone modulation (DMT). In [2], a 1.25 Gbit/s transmission with pre-filtered four-level pulse amplitude modulation (4PAM) signaling and fractionally spaced (FS) equalization is presented. However, the prefilter scheme is quite impractical to meet the demand for ever-higher data rates, because the dynamic range of a pre-distorted input signal increases along with the data-rate. So a large DC offset is needed for the LED, which leads to more power consumption and a decrease in the receiver sensitivity. In [3], a 1 Gbit/s transmission is demonstrated using DMT. As is well known, DMT requires high linear transceivers over a wide range of the input optical power, which is difficult for the circuit design. Moreover, the increased cost from using QAM modulator/demodulator and FFT/IFFT components in DMT is a sensitive factor for the car-makers. Until now, most research works targeting at 3 Gbit/s used costly laser diode (LD) or complex modulation schemes. Our work, on the other hand, reports on attempts to achieve the 3 Gbit/s data-rate with a low cost LED and a practical cost-effective system layout. In [4], a 3 Gbit/s transmission is first achieved with a decision feedback equalizer (DFE) and a cheap LED.

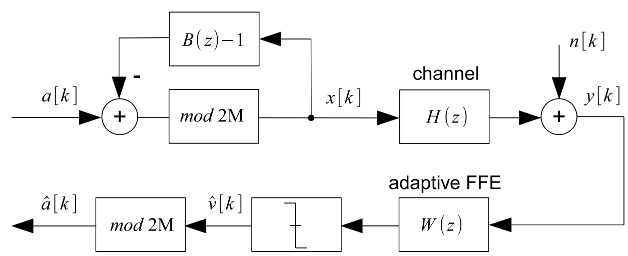

Fig. 1. System block diagram

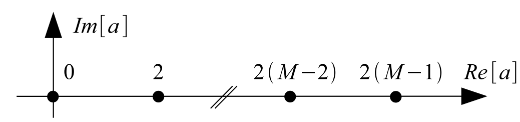

Fig. 2. Modified signal constellation diagram

To reach better performance than [4], while avoiding drawbacks in [2] and [3], this paper combines the Tomlinson Harashima Precoding (THP), which is a nonlinear form of pre-equalization, with a linear feedforward equalizer (FFE) at the receiver, in order to mitigate the error propagation problem of the DFE as well as to restrict the transmit power. The complexity of THP depends linearly on the channel length. The FFE taps are either optimized by the minimum mean squared error (MMSE) criterion or updated by the least mean square (LMS) algorithm. It will be shown that THP-FFE is superior to DFE at 3 Gbit/s, and is robust against variations in a transmission.

II. Overview of the system design

The system block diagram at the symbol level is depicted in Fig. 1. The information bits in the bit level are first encoded by Reed Solomon (RS) coding and then mapped into M-level unipolar pulse amplitude modulation (MPAM) symbols $a[k]$. By taking the bandwidth efficiency, cost and system linearity into account, 8PAM is chosen as the mapping scheme.

$H(z)$ is the discrete form of the electrical equivalent base-band channel of the MOST optical physical layer. The con-

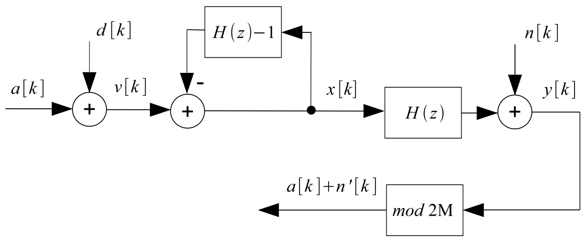

Fig. 3. Linearized THP model

sidered optical layer is a cascade of an LED, a 10 m POF and a photo diode (PD). According to the MOST150 optical physical layer specification [5], its electrical equivalent transfer function can be modeled by a Gaussian low-pass filter:

$$H_a(f) = Ae^{-2(\pi\sigma f)^2} e^{-j2\pi f \tau L_{pof}}; \qquad (1)$$

where L_{pof} is the fiber length, A is the linear fiber loss, $\sigma = \frac{0.132}{B}$ is the standard deviation, $B = 1009 \cdot (\frac{L_{pof}}{m})^{-0.8747}$ MHz is the 3 dB bandwidth, and $\tau = 4.97 \cdot 10^{-9}$ s/m. For a 10 m POF, the 3 dB channel bandwidth is about 134 MHz.

The information symbols $a[k]$ are first encoded by the THP, which consists of a modulo 2M (mod 2M) device and a feedback loop. Here, a change has been made to the classical THP and bipolar MPAM combined scheme. That is, we used the unipolar MPAM and shifted the output range of the mod 2M device from $[-M, M)$ to $[0, 2M)$, as shown in Fig. 2. The reason is that a photo diode is only capable of detecting the power of the received signal, and a loss of phase information is inevitable. Note that this modification does not change the properties of THP.

The mod 2M device is the key element which makes THP outperform the linear prefilter in [2]. Its algorithm can be better explained by a linearized THP model as shown in Fig. 3, where we assume $H(z)$ is a monic channel and no $W(z)$ is required at the receiver. The mod 2M operation is equivalent to adding an unique value $d[k] \in 2M \cdot N$, where M is the order of MPAM and N is an arbitrary integer number, to the symbol $a[k]$ at each time instant k, so that the channel input signal $x[k]$ lies in the interval $[0, 2M)$. By operating non-linearly, the amplitude of $x[k]$ is bounded, thus an amplification of the transmit power in [2] is avoided. To be more precise, THP also introduces certain amplification to the transmit power, but it is small and limited. We will prove this in the later section. Without the presence of noise, $y[k] = v[k] = a[k] + d[k]$ is received after the channel, because the feedback loop perfectly equalizes the channel $H(z)$. So the receiver can simply repeat the mod 2M operation to remove the additive sequence $d[k]$ in order to recover $a[k]$.

To initialize the system, the channel estimation is done within a training block before the data transmission. Based on the channel estimate $\hat{H}(z)$, the feed forward equalizer $W(z)$ is calculated at the receiver by forcing a $B(z) = \hat{H}(z) \cdot W(z) \cdot z^{k_0}$ as close as possible to a causal and monic channel, where k_0 is the index of the main tap. The post-cursors of $B(z)$ are then fed back to the transmitter to determine the coefficients of THP, i.e., the feedback loop of THP is set to $B(z)-1$. Once the THP coefficients are determined, they are fixed. While $W(z)$ serves as the adaptive part in the system, which is constantly updated to handle transmission errors like mismatch of the THP coefficients, channel variation or channel estimation errors. Note that the pre-cursors of $B(z)$ can not be handled by THP, which will severely decrease the performance of THP when they come into play.

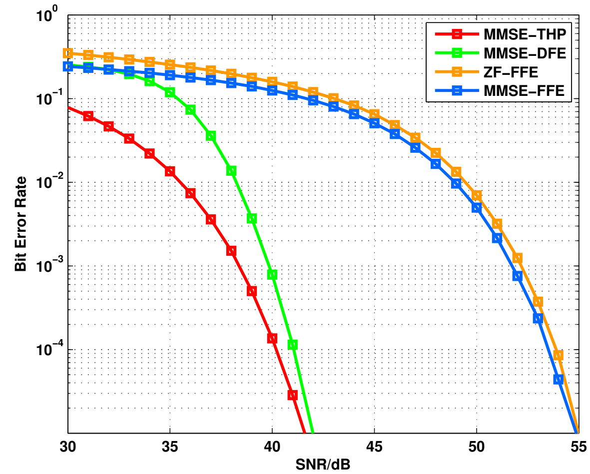

Fig. 4. Comparison of BER performances for THP, DFE, ZF-FFE and MMSE-FFE transmission schemes as a function of SNR at 3Gbit/s

III. Simulation Results

The system is evaluated by simulations of bit error rate (BER) as a function of signal-to-noise ratio (SNR). The area-of-interest for SNR is $20.4 \sim 59.5$ dB based on [4] according to the MOST optical physical layer link budget [5]. The area-of-interest for BER is 10^{-4} to 10^{-3}, so the RS coding used in the bit level can decrease the final BER to 10^{-9} and below.

The FFE $W(z)$ at the receiver is fractionally-spaced with twice the symbol rate. Its taps are either optimized in terms of mean squared error for a specific SNR value, or updated by the least mean squares (LMS) algorithm if $W(z)$ is adaptive. Because of the slow fading nature of the optical channel, a small step-size of LMS is sufficient for the variance tracking.

Simulations are first run to identify the improvement of MMSE THP-FFE in comparison to linear FFE and MMSE-DFE, respectively. Then, the robustness of the adaptive THP-FFE as well as its performance under reduced channel bandwidths are investigated. Finally, THP-FFE is compared to an ideal DFE without error propagation.

A. Comparison of THP-FFE, FFE and DFE at 3Gbit/s

Fig. 4 shows a 3 Gbit/s transmission with different system layouts, where taps of THP-FFE and DFE are calculated with the MMSE criterion. Taps of FFE are calculated with zero forcing (ZF) and MMSE criterion, respectively. It can be seen that the THP-FFE outperforms other schemes by reaching the target BER $10^{-4} \sim 10^{-3}$ with a SNR less than 40 dB.

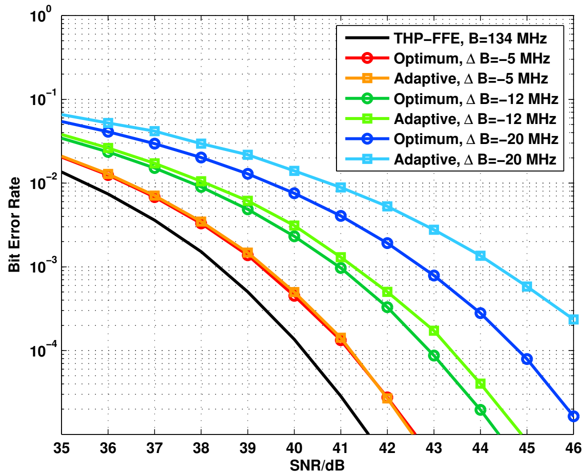

Fig. 5. BER performance of the adaptive THP-FFE transmission scheme as a function of SNR for various bandwidth reductions ΔB

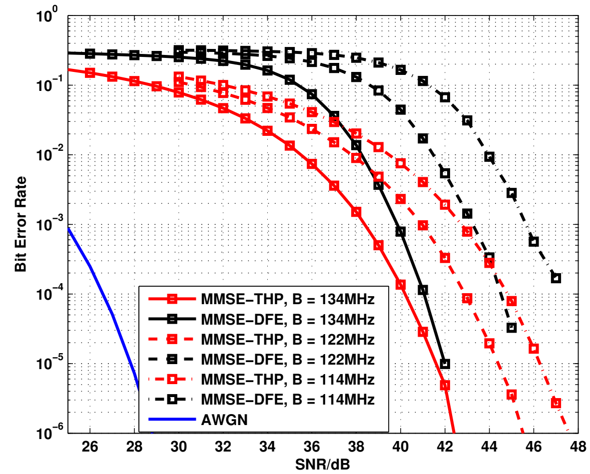

Fig. 6. Comparison of BER performances for MMSE THP-FFE and MMSE DFE transmission schemes with different channel bandwidths B

According to the power budget, a power margin of nearly 20 dB is reserved. Comparing to the DFE solution in [4], THP-FFE presents a performance gain of $1 \sim 2$ dB.

At the same time, a huge improvement of the nonlinear equalizers (THP-FFE and DFE) in comparison to the linear equalizers (ZF-FFE and MMSE-FFE) can be observed. In the BER range of interest, the nonlinear equalizers have 10 dB SNR gain over the linear ones. Because both DFE and THP manage to avoid the large noise enhancement of a linear FFE.

B. Performance of the adaptive THP-FFE under a temperature change

In a practical environment, a channel bandwidth decrease might happen due to the increase of the operating temperature during the transmission. Robustness of the adaptive THP-FFE against such a situation is examined in this section.

According to measurements in [6] under a temperature change in the automobile environment from -40°C to 105°C, the maximum bandwidth decrease is $\Delta B_{max} \approx -20$ MHz. Hence, we test the adaptive system with three different bandwidth drops respectively: a small one $\Delta B = -5$ MHz, a middle one $\Delta B = -12$ MHz, and a big one $\Delta B = -20$ MHz. During the transmission, ΔB is reached gradually.

The results are depicted in Fig. 5. The black curve is the BER performance of MMSE THP-FFE under room temperature with an initial channel bandwidth $B_{\text{-3 dB}} = 134$ MHz. The colored curves demonstrate the performances with bandwidth decreases. For each value of ΔB, two set-ups are compared: an optimum set-up where the coefficients of MMSE THP-FFE are re-calculated once the bandwidth decreases, and an adaptive set-up where LMS algorithm is used at the receiver FFE to follow the channel variation, while the THP coefficients at the transmitter remain fixed.

As expected, performances for both set-ups get worse with the decreasing bandwidth. However, comparing the adaptive set-up with the optimum one, we notice that for a small bandwidth decrease of 5 MHz, the difference between them is negligibly small. In such a situation, a simple adaptive filter at the receiver side is completely sufficient. For a bandwidth drop of 12 MHz, they still have similar performances. Only under the situation that the bandwidth is decreased by 20 MHz, the difference is noticeable. But even so, the adaptive filter performs very stable.

C. Effects of a reduced channel bandwidth

Based on the consideration that an aging degradation in hardware or use of a longer fiber will lead to a permanent decrease of the channel bandwidth, this section investigates the THP performances with a channel bandwidth that is smaller than the theoretical value. Fig. 6 compares the MMSE THP-FFE with MMSE DFE for different channel bandwidths at 3 Gbit/s. It clearly shows that THP-FFE outperform DFE. The reason is that a smaller channel bandwidth produces stronger ISI for a fixed data-rate. In order to combat the ISI, DFE enlarges correspondingly its tap length and values, which increases the probability of an erroneous decision and prolongs the propagation of an error. Contrarily, THP does not suffer from the error propagation, because the encoder uses the actual past symbols instead of the decided past symbols for getting rid of the ISI. Therefore, THP is more advantageous in applications with e.g., a longer fiber, a slower LED or a higher data-rate. Note that the reference BER curve for AWGN channel belongs to the biased MPAM.

D. Discussion of THP-FFE losses

It is well known that THP-FFE suffers from modulo loss and precoding loss, whereas DFE suffers from error propagation. Whether THP-FFE or DFE is more beneficial depends on which loss is dominant. Therefore, we first calculate numerically the THP-FFE losses, and then compare THP-FFE with an ideal DFE without error propagation. The ideal DFE can be considered as the upper bound for THP-FFE, because THP-FFE is somehow equivalent to a DFE by moving the feedback filter in DFE to the transmitter [7], it can never be superior to an ideal DFE without error propagation.

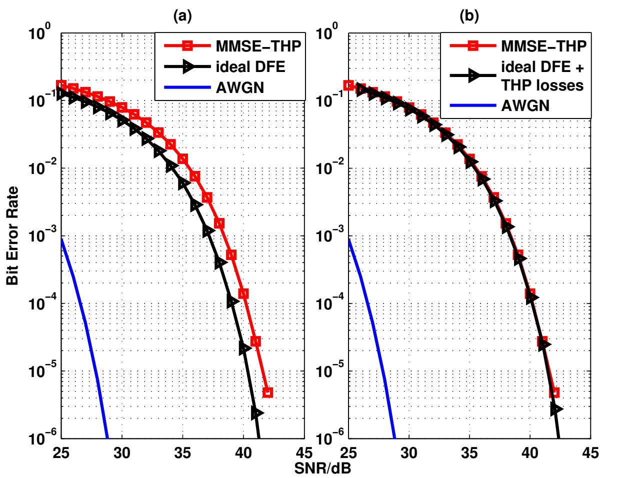

Fig. 7. a) Comparison of a MMSE THP-FFE and an ideal DFE; b) Comparison of a MMSE THP-FFE and an ideal DFE added by THP losses

The losses of THP-FFE are calculated as follows:

- the modulo loss:
  When introducing the congruent signal constellation for THP-FFE, the MPAM signal constellation is periodically repeated, hence the symbol error probability of the edge symbols are now doubled. The increased symbol error rate from $P_{S,MPAM}$ to $P_{S,THP}$ can be calculated as:

$$\gamma_{modulo} = \frac{P_{S,MPAM}}{P_{S,THP}} = \frac{2\frac{M-1}{M}Q(\frac{b}{\sigma_n})}{2Q(\frac{b}{\sigma_n})} = \frac{M-1}{M} \quad (2)$$

  with b being the distance from a signal point to its neighborhood decision threshold, σ_n being the noise standard deviation, and $Q(x)$ being the Q-function. For $M = 8$, we get $\gamma_{modulo} = 1.1429$. In the logarithmic plot this factor corresponds to a shift of $\log(1.1492) = 0.058$ upwards.
- the precoding loss:
  The precoding loss comes from an increased transmit power of THP-FFE comparing to DFE. However, we use the received SNR as the figure of merit. So the precoding loss at the transmitter has to be transferred to the receiver. Unlike an unbiased (zero mean) transmission, the simple shift of the constellation diagram at the transmitter changes the distribution of the modulo output. Therefore, calculation of the power penalty at the receiver, which is the ratio of the expected received signal power for THP-FFE to the expected received signal power for DEF, has to be done explicitly. For the fractionally spaced system, we have:

$$\mathrm{E}[y^2[k]] = \frac{1}{2}\sigma_s^2 \sum_{i=0}^{P-1} h^2[i] + \frac{1}{2}\mu_s^2 (\sum_{\substack{i=0 \\ i \text{ even}}}^{P-1} \sum_{\substack{j=0 \\ j \text{ even}}}^{P-1} h[i]h[j] + \sum_{\substack{i=0 \\ i \text{ odd}}}^{P-1} \sum_{\substack{j=0 \\ j \text{ odd}}}^{P-1} h[i]h[j]) \quad (3)$$

  with h being the fractionally sampled channel, μ_s and σ_s^2 being the mean value and the variance of the transmit signal $s[k]$, respectively. So the power penalty from the precoding loss is:

$$\gamma_{power-penalty} = \frac{\mathrm{E}[y_{DFE}^2[k]]}{\mathrm{E}[y_{THP}^2[k]]} \quad (4)$$

  For the biased DFE, $\mu_s = \mathrm{E}[a[k]] = M - 1$, and $\sigma_s^2 = \sigma_a^2 = (M^2 - 1)/3$. For the biased THP-FFE, $\mu_s = \mathrm{E}[x[k]] = M$, and $\sigma_s^2 = \sigma_x^2 = M^2/3$. According to (4), we get $\gamma_{power-penalty} = 1.2744$ for our channel. This power penalty can be regarded as a shift of 1.05 dB to the right on the SNR axis.

When we bring in the calculated losses on the BER plots in Fig. 7, we see: in Fig. 7 (a), an ideal DFE without error propagation performs always better than THP-FFE. However, by adding the two expected losses of THP-FFE to its BER curve, we get almost the curve of THP-FFE, as shown in Fig. 7 (b). Therefore our analysis and calculation of the expected losses are very reliable. When an actual DFE suffers from severe error propagation under strong ISI, THP-FFE has only a certain loss to the ideal DFE. This also well explains the increasing gap between the BER curves of THP-FFE and DFE in Fig. 6 for a decreasing channel bandwidth.

IV. Conclusion

As a conclusion, THP-FFE offers a larger SNR margin than DFE and linear FFE for the sake of data transmission at low error rates. THP-FFE is also more energy-efficient than a pure prefilter. In addition, the adaptive THP system is robust against transmission errors and channel bandwidth reductions caused by temperature variations or hardware degradation. It is a cost-effective, simple and reliable solution for high speed in-car communications based on the MOST optical physical layer.

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