High-Speed Electrical Backplane Transmission Using Duobinary Signaling

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***Abstract*—High-speed electrical data transmission through low-cost backplanes is a particularly challenging problem. We present for the first time a very effective approach that uses the concept of duobinary signaling to accomplish this task. Using a finite-impulse-response filter, we are able to compensate for the phase and amplitude response of the backplane such that the resulting frequency response of the channel is that of an ideal duobinary filter. At the output of the backplane, we use an innovative pseudodigital circuit to convert the electrical duobinary to binary. For 10-Gb/s data transmission, we demonstrate a bit error rate $< 10^{-13}$ through electrical backplane traces up to 34 in in length on FR4. A full discussion of the concept, system architecture, and measured results are presented. Analysis is presented that compares and contrasts this approach to PAM-4 and standard nonreturn-to-zero signaling.**

***Index Terms*—Backplanes, correlative coding, duobinary, equalization, high-speed data, PAM-4, pre-emphasis, signal integrity.**

I. Introduction

GIGAHERTZ-SPEED data transmission is required in high-capacity multiservice switches (MSSs) such as core routers and cross-connect switches [1]. Currently, state-of-the-art MSSs use 2.5-Gb/s data transmission to route traffic between line and switch fabric cards over passive electrical backplanes. The need for higher data capacity in these MSSs and the support of 40-Gb/s line rates to the optical network are driving system development toward higher line rates over electrical backplanes. As a result, the next-generation MSSs will employ 10-Gb/s backplane signaling; however, current low-cost backplane materials and connectors do not provide adequate bandwidth to support this transmission rate. Backplane ethernet is another application that is pushing the required backplane transmission speed by providing solutions for routers, switches, and blade server platforms. As a result of the growing demand in many markets for high-speed backplane transmission, the design of electrical backplane systems operating at 10 Gb/s and beyond has emerged as an important field of study. There are several approaches being pursued by many vendors to increase backplane transmission speed. These techniques fall into basically two categories, i.e., passive and active.

Passive solutions incorporate the use of high-quality microwave substrate materials, innovative via-hole techniques, and new connector technology. These techniques can help to solve the transmission problem; however, the use of costly microwave substrates and special high-bandwidth backplane connectors are often required, while very long trace lengths may still result in unacceptable transmission characteristics.

Active solutions strive to increase the backplane throughput by using signal processing to overcome the poor transmission properties of the physical channel. The claim is that modification of electronics should ultimately be more cost effective than modifying the backplane itself. Additionally, the bandwidth of backplane systems can be increased by changing only the line cards, not the whole system. Such an approach provides a way to incrementally upgrade systems that have already been deployed in the field. There are currently several active solutions being pursued by the industry, including adaptive equalization [2], pre-emphasis, PAM-4, or some combination thereof [3], [4]. Typically these solutions can provide good transmission performance even for long trace lengths; however, power consumption and system complexity are issues here. Typically equalization and pre-emphasis techniques must provide correction for the entire NRZ data bandwidth. The problem is that, for numerous low-quality backplane transmission systems, the frequency-response rolloff is severe, and the use of via-holes on thick backplanes results in nulls in the frequency range of interest. Clearly, equalization or pre-emphasis through nulls requires the use of higher order networks, and the resulting correction will be very sensitive to temperature and parameter variations.

A solution to the problem of poor high-frequency response is to compress the bandwidth using multilevel coding. PAM-4 with equalization is currently being used by some vendors [3], [4] to solve this problem. With multilevel signaling, signal-to-noise ratio (SNR) performance is sacrificed in exchange for a narrower bandwidth requirement. This technique has been shown to provide very good performance even over long traces; however, these circuits are typically complex, leading to difficulty providing dense integration and significantly increased power consumption relative to standard NRZ signaling.

The use of electrical duobinary signaling [5] for high-speed backplane transmission to accomplish the two tasks of bandwidth reduction and simplification of implementation relative to other approaches was proposed for the first time in [6]. Duobinary takes advantage of the inherent rolloff of the backplane channel, which results in an architecture that is very simple to build. Additionally, duobinary does not incur nearly as much SNR penalty [(relative to nonreturn to zero (NRZ)] as PAM-4, as will be discussed. In [6], the basic system concept, implementation, and illustration of performance at 10 Gb/s through traces up to 34 in ($\sim$86 cm) using low-cost backplane materials and established backplane connectors was demonstrated for the duobinary signaling scheme. Here, we expand upon this

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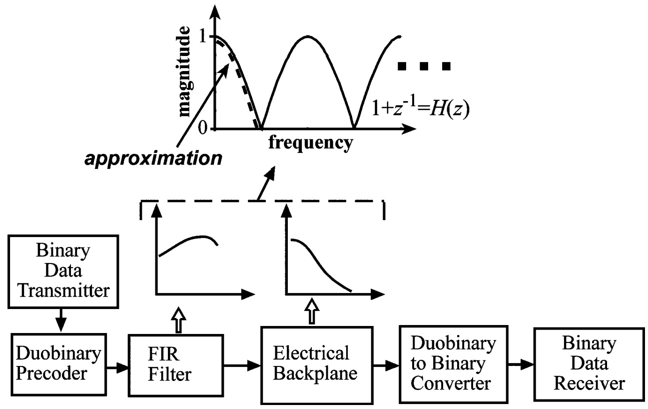

Fig. 1. System architecture (from [6]).

previous study by demonstrating the advantage of using fractional finite-impulse-response (FIR) pre-emphasis, describing more advanced FIR synthesis techniques, and providing an analytic comparison of performance of NRZ, PAM-4, and duobinary signaling. Additionally, we provide simulation results that illustrate eye diagram performance for NRZ, PAM-4, and duobinary through 6- and 34-in Tyco Quadroute backplanes. Finally, we discuss the asymmetric nature of the duobinary eye and its impact on the optimal decision threshold.

II. Duobinary Signaling Architecture for High-Speed Backplanes—System Concept

The idea behind the duobinary signaling architecture is to take advantage of the rolloff response of the backplane channel and use it to help shape the data waveform that is to be sent to the receiver. In order to understand why this makes sense, one must first understand how duobinary data is generated from an NRZ data waveform.

As described by Lender [7], duobinary data can be generated by sending NRZ data through a "delay and add" filter. As shown in Fig. 1, this filter has a Z-transform of $1 + z^{-1}$, and a frequency response as illustrated. Since all physical channels are band-limited at some point, we are unable to reproduce the infinite number of passbands illustrated in Fig. 1 so we use a reasonable approximation to the ideal response as shown. As illustrated by the dashed line in Fig. 1, the approximation is that of a low-pass filter. It is important to note that typically the roll response of a backplane is *also* that of a low-pass filter; however, one that is too steep. If we provide some additional filtering, either before or after the backplane, we can generate the required response (illustrated in Fig. 1) from the cascade of the filter(s) and the backplane. The advantage of such a technique over more complex multilevel signaling schemes is that duobinary only requires a low-pass filter to convert the input two-level NRZ signaling to a three-level signal. In the microwave world, filter design is one of the most mature fields and, thus, there are many techniques, both digital and analog, available to the designer for providing the necessary pre- or post-emphasis required to turn the backplane channel into the required approximation of the delay-and-add response.

The proposed transmission system has several main components, as illustrated in Fig. 1. An NRZ data source, a duobinary precoder, a data spectrum reshaping filter, the backplane, a duobinary-to-binary data converter, and an NRZ receiver comprise the entire system. It will be shown that, from a hardware standpoint, it is simple to reshape the spectrum with a filter at the transmitter; however, certain conditions may dictate the use of a receive filter in addition to or instead of a transmit filter.

As mentioned earlier, the typical low-cost backplane has a frequency rolloff that is much steeper than that of a 10-Gb/s duobinary signal. As a result, the reshaping filter will have the job of emphasizing the higher frequency components, as well as flattening the group-delay response across the band. It is important to note that since the duobinary data spectrum has a null at half the bit rate, the amount of high-frequency emphasis is greatly reduced when compared to uncoded NRZ data. Additionally, nulls that occur in the transfer function of the backplane as a result of via-hole resonances typically become more predominant toward the higher end of the frequency spectrum; therefore, the reduced data bandwidth provides a distinct advantage.

III. Designing the Reshaping Filter

The frequency response of an N-tap integer-delay FIR filter will have the form

$$H_{\mathrm{FIR}}(\mathbf{C},\omega) = \sum_{q=0}^{N-1} c_q e^{-jq\omega T} \tag{1}$$

where c_q = the qth tap coefficient, $\mathbf{C}$ = vector of coefficients, T = bit period, and ω = angular frequency. We will show that we can achieve less pattern dependence than previously reported [6] with a low-order filter that contains *fractional* delays between taps and has the following form:

$$H_{\mathrm{FIR}}(\mathbf{C},\boldsymbol{\tau},\omega) = \sum_{q=0}^{N-1} c_q e^{-j\omega\tau_q} \tag{2}$$

where τ_q = the qth tap delay and $\boldsymbol{\tau}$ = vector of tap delays.

If we define the complex transmission frequency response (S_{21}) of the backplane as $H_{\mathrm{CH}}(\omega)$, then the required filter response will be

$$H_{\mathrm{FIR}}(\omega) = \frac{H_D(\omega)}{H_{\mathrm{CH}}(\omega)} \tag{3}$$

where $H_D(\omega) = 1 + \exp(-j\omega T)$, which is the ideal frequency response of a duobinary filter, or simply a delay and add filter, as described in Section II. In general, the use of (3) may result in a filter that has numerous coefficients. For a high-speed discrete time implementation, this is not desirable. Instead, we can limit the number of coefficients and use the L_p norm to carry out an optimization to obtain the required filter response, as described in [6].

It is important to note that the frequency-domain approach to optimization is not the only technique that will yield optimal

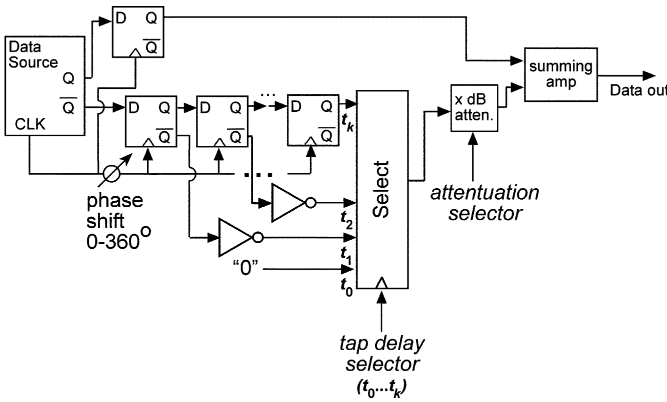

Fig. 2. Proposed integrated-circuit implementation of the fractional FIR pre-emphasis filter—general two-tap solution.

taps. A time-domain approach to optimization is also possible. Furthermore, more advanced optimization techniques can be used that take advantage of bit error rate (BER) calculations. Such a technique will be illustrated in Section V.

Fig. 2 illustrates a proposed architecture for a two-tap reshaping filter that can be used for both integer and fractional-delay implementations. This is not the most general reshaping filter; however, it is believed to be adequate for most cases while providing simple implementation. A delayed, scaled, and when required, inverted replica of the original data stream is added to the original data stream so as to realize the response from (2) with $N = 2$. The available delays for this second tap are multiples of the bit time plus a time offset ($t_n = (T \times n) + \tau$, where $t_n \in \{t_0, \ldots, t_k\}$ and $0 < \tau < T$), and are selected using the tap delay selector in Fig. 2. In other words, a delay from 0 to $(k+1)T$ is possible. It is important to note that a linear inverting amplifier is not needed to realize a minus sign in (2) since the input data is purely digital at this point in the system.

IV. Duobinary-to-Binary Data Conversion

To implement the proposed signaling scheme, the ability to convert high-speed duobinary data to binary data is required. When the concept of duobinary was first invented [7], [8], processing data rates as high as gigabits per second was not a possibility. The original implementation for the duobinary-to-binary data converter was merely a full-wave rectifier as explained by Lender [7], [8]. Unfortunately, for high-speed data transmission, such as data rates ranging from 10 to 40 Gb/s, the implementation of a broad-band full-wave rectifier with a flat response over the entire data bandwidth would be very difficult to build, even today. The use of broad-band microwave matching circuits would be required, resulting in a circuit that would not lend itself well to dense circuit integration, as desired here. It is important to note that the duobinary-to-binary conversion discussed here is achieved in a completely different way than that discussed in recent literature describing the use of duobinary signaling in the optical domain [9].

We have come up with an innovative approach for converting duobinary to binary using a pseudoanalog/pseudodigital approach that can be implemented in any high-speed logic family. This data converter is illustrated in Fig. 3. This circuit can

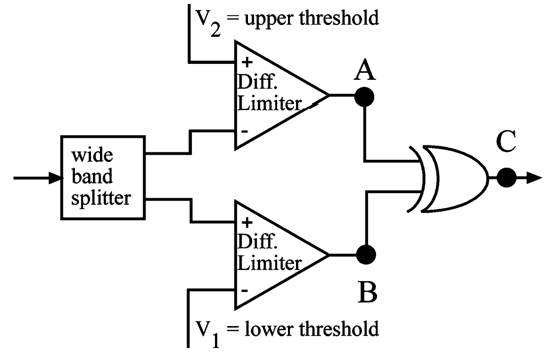

Fig. 3. Duobinary-to-binary data converter (from [6]).

(a)

Condition	A	B	C
$s(t)>V_2$	0	1	1
$V_2>s(t)>V_1$	1	1	0
$V_1>s(t)$	1	0	1

(b)

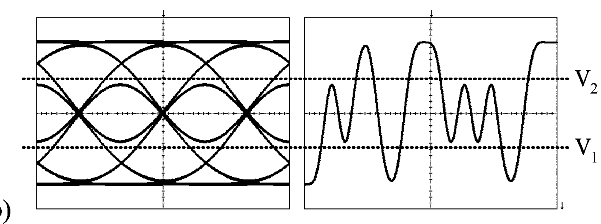

Fig. 4. Data converter (a) circuit truth table, where nodes **A**–**C** are illustrated in Fig. 3 and (b) duobinary data signal $s(t)$ plotted as an eye diagram (left-hand side) and as a pattern (right-hand side).

be realized in hardware using a balanced input exclusive-OR gate where the thresholds at the inputs are set appropriately. The duobinary input signal is presented to a wide-band splitter with a bandwidth of at least half the data rate, and then sent to two comparators (differential limiters), where the signal is simultaneously compared to two different thresholds. The results of these comparisons are sent to an X-OR gate whose output will now be a binary signal. Fig. 4 illustrates the truth table and logic levels that explain the functionality of this converter completely. We have measured the performance of this duobinary-to-binary conversion circuit and achieved error rates $< 10^{-15}$ up to 10 Gb/s. Using higher speed logic families, we expect that conversion at 40 Gb/s should be achievable.

V. Experimental Results

As a demonstration of our proposed system architecture, we have assembled a laboratory experiment that illustrates the important aspects of this methodology. Fig. 5 illustrates our setup. This setup is intentionally the same as that used in [6] so as to provide a fair comparison of our new performance results with our previously reported results [6]. A bit error rate tester (BERT) operating at 10 Gb/s is used to generate an NRZ data waveform, which then travels through a FIR filter, an experimental backplane, a duobinary-to-binary data converter, and finally into the receiver of the BERT. The duobinary precoder of Fig. 1 is absent from the test setup because the use of a precoded pseudorandom

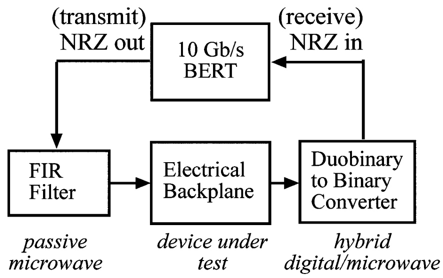

Fig. 5. Experimental test setup (from [6]).

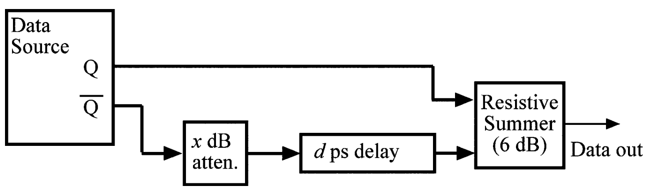

Fig. 6. Microwave realization of NRZ data + FIR filter used in laboratory experiments (34-in trace on Quadroute backplane) (from [6]).

TABLE I
BACKPLANE CHARACTERISTICS

Manufacturer/ Board Name	Nelco Dielectric	Trace Geometry (width, space, width)	Connector Type
Tyco/Quadroute	4000-6	4,6,4 (*mils*)	HM-ZD
Tyco/XAUI	4000-2	10,14,10 (*mils*)	HM-ZD
Tyco/XAUI	4000-6	10,14,10 (*mils*)	HM-ZD

bit sequence (PRBS) sequence would result in exactly the same sequence shifted in time.

Fig. 6 illustrates the circuit used to realized the FIR reshaping filter for both integer and fractional delays. We used two fixed delay/attenuator settings, one for 34-in traces and another for 20-in traces. Table I illustrated the characteristics of the different backplanes used. They are all FR4 boards designed by Tyco and are all modifications of their standard release boards, except for the Tyco XAUI made with Nelco 4000-2, which is a commercially available unit. All measurements made were single ended, as it is expected that the single-ended transmission performance will be very similar to the differential-mode transmission performance based on the loose coupling between traces, as described in Table I. The unused trace in the differential pair was terminated with 50-Ω loads on either end. Fig. 7 illustrates the insertion loss for three typical traces on the Tyco Quadroute backplane for total lengths of 6, 20, and 34 in, respectively. By inspection, it is clear that these responses can be approximated, to first order, as lines on a log-magnitude versus frequency plot.

A. System Performance Using a Two-Tap Integer-Delay FIR Pre-Emphasis Filter—Review and Techniques for Improvement

System performance using a two-tap integer-delay FIR filter was first presented and explained in detail in [6]. Fig. 8 illustrates the 10-Gb/s waveforms at various points so as to illustrate the evolution of the waveform through the system. It is important to note that the duobinary nature of the waveform in Fig. 8(b) is apparent, however, it appears to be less than ideal when compared to illustrations of duobinary eyes, such as those shown in Fig. 4(b) (Section IV) and by Lender [7]. Although the eye appears to be open and free from dispersion, the opening is somewhat smaller than it could be and appears to be suffering from not quite enough pre-emphasis. Clearly one solution to the problem would be to add additional taps to the FIR filter so as to better match the response of the channel, however, in the interest of simplicity, we will show that we can improve the performance using the same number of taps with fractional delays.

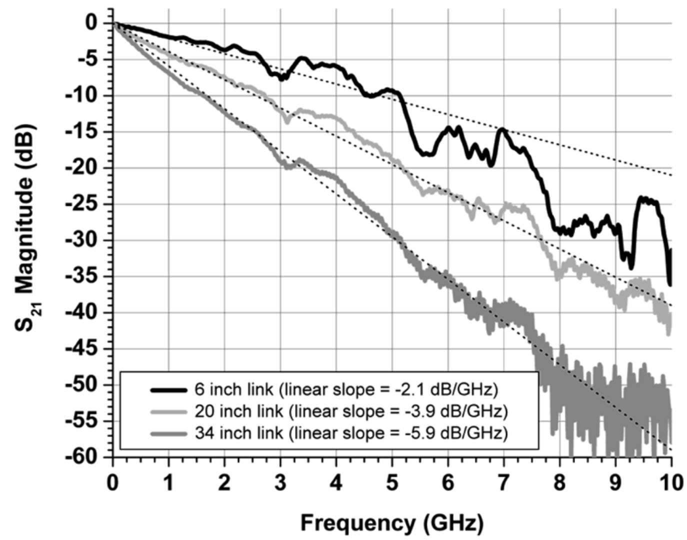

Fig. 7. Representative insertion-loss measurements of 6-, 20-, and 34-in Tyco Quadroute traces.

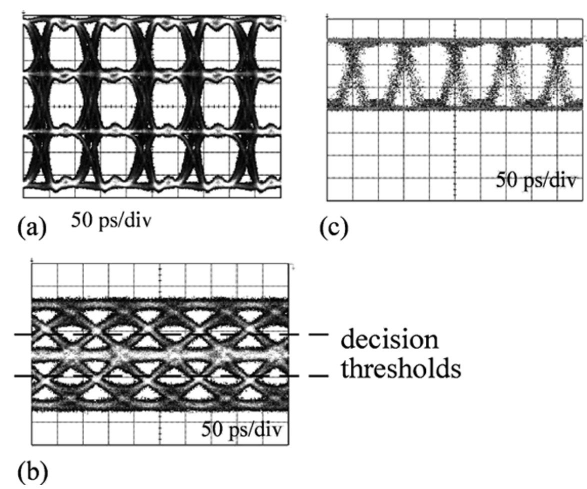

Fig. 8. Waveform evolution through the system with an integer-delay FIR filter (from [6]). Output from: (a) FIR filter, (b) 34-in trace on Tyco Quadroute backplane, and (c) data converter.

B. System Performance Using a Fractional-Delay FIR Pre-Emphasis Filter

As previously reported [6], we found some pattern dependence when measuring BER performance for the 34-in trace lengths (pattern length $= 2^{15} - 1$) using integer-delay FIR filters. Furthermore, the eye opening shown in Fig. 8(b) appears to be less than optimal. If we were to use a high-order filter, these issues could probably be alleviated; however, our goal is to maintain a simple filter architecture. We will now show

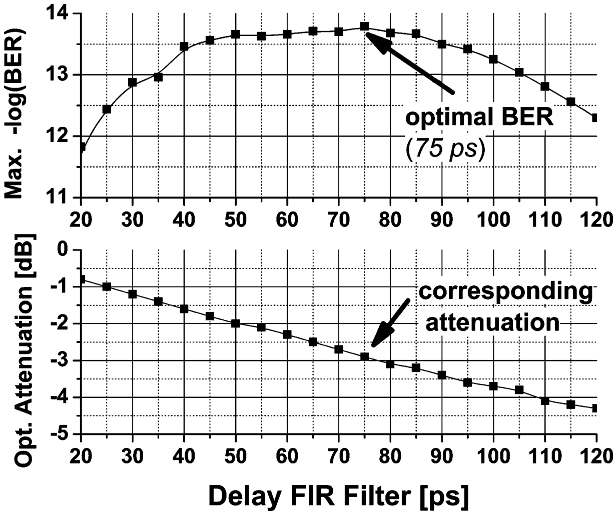

Fig. 9. Optimal tap selection for two-tap fractional FIR synthesis on the Tyco Quadroute 34-in trace. The BER data corresponds to an arbitrarily chosen noise level at the transmitter and does not imply a limit on BER performance of the system.

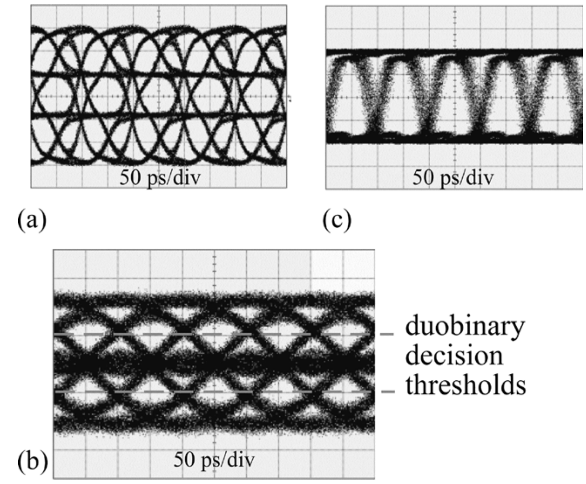

Fig. 10. Waveform evolution through the system with a fractional-delay FIR filter. Output from: (a) FIR filter, (b) 34-in trace on Tyco Quadroute backplane, and (c) data converter.

TABLE II
MEASURED RESULTS

Trace Length	20 in. (~51cm)	34 in. (~86 cm)
Data Rate	10 Gb/s	10 Gb/s
Sequence Length	2^{31}-1	2^{23}-1
BER	$<10^{-13}$	$<10^{-13}$
Backplanes Tested	QUADROUTE, XAUI	QUADROUTE, XAUI
FIR Architecture	Fractional 2-tap	Fractional 2-tap

that the pattern dependence and eye opening measured using the two-tap integer-delay FIR filter can be improved by the use of a fractional-delay two-tap FIR filter [10]–[12]. This result is somewhat intuitive if we consider the function of the FIR filter in the time domain. The purpose of the FIR filter is to remove as many pre- and post-cursors in the impulse response of the channel as possible while generating a pulse width that approaches the inverse Fourier transform of the first lobe of the ideal delay-and-add FIR filter illustrated in Fig. 1. If we are trying to maintain a minimum number of taps for the required FIR filter, there is no reason to assume that the required tap delays should fall on integer multiples of the bit time. As such, it is quite reasonable to assume that, in general, the best second tap delay will not fall on an integer multiple of the bit period.

We have used a simulation approach that computes the expected BER of the link including a two-tap fractional-delay FIR filter, and varies the tap delay and weight to minimize this quantity. The simulation uses a technique know as semianalytic analysis [13] to determine the BER at the output of the link. Fig. 9 illustrates the optimization approach used on the 34-in trace. This simple graphical approach is possible because with a two-tap FIR filter, only one weight and delay need to be determined. This approach is applied to the other trace lengths as well.

Table II summarizes the BER performance results and corresponding system parameters for duobinary signaling using fractional FIR filter pre-emphasis. We obtain performance with longer sequence lengths that previously reported in [6] due to the improved performance of the fractional FIR filter used for pre-emphasis. Furthermore, the eye opening in Fig. 10(b) appears much improved over that of Fig. 8(b).

For each trace length, a single set of FIR settings is used for all three different boards. This illustrates that the backplane properties over the lower half of the frequency spectrum (<5 GHz) are very similar even for different boards. Although in a final commercial implementation of this technology, the FIR filter may need to be adaptive, in this experiment, fixed tap delays and amplitudes were maintained throughout for a given trace length.

VI. COMPARING DUOBINARY WITH OTHER DATA FORMATS

To the best of the authors' knowledge, the first discussion of duobinary signaling for high-speed electrical backplanes is in [6]. As such, the current market has not addressed the differences between this technique and others commonly used. Currently, the two most common data formats used for electrical high-speed data transmission are NRZ and PAM-4 signaling [14]–[22]. The high-speed backplane community is currently embracing both of these techniques. A full and comprehensive comparison of these formats including backplane effects, crosstalk, and signal-to-noise issues is complex and important enough that we plan to follow this study with an entire paper dedicated solely to this topic. Here, we will provide an overview and comparison between NRZ, PAM-4, and duobinary in such a way as to provide the reader with some intuition into the problem.

A. Simple Overview of NRZ Versus Bandwidth Compressed Formats—Duobinary and PAM-4

The first comparison that will be done is between techniques that compress data bandwidth versus standard NRZ. We will derive a very simple expression that compares the performance of an NRZ, PAM-4, and duobinary based solely on occupied data bandwidth. We will define the required data bandwidth for NRZ as f_0 and the required data bandwidth for PAM-4 and duobinary as 1/2 f_0. What follows is a high-level discussion that will provide an intuitive understanding of when the use of bandwidth compression will prove beneficial when compared to standard

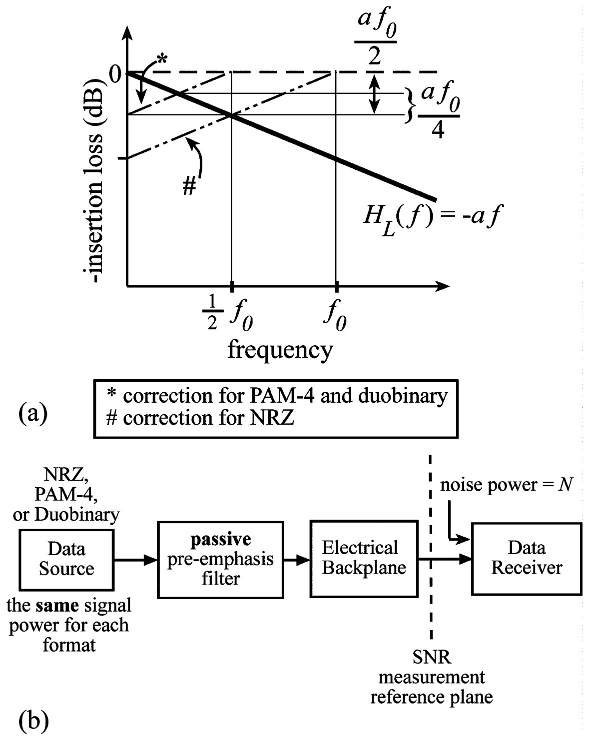

Fig. 11. (a) Required correction for data with and without 2× data compression using passive pre-emphasis. (b) Simple schematic illustrating assumptions in SNR analysis.

NRZ signaling. The premise of the argument to follow is that we can completely correct for the degradation of the channel by flattening the channel response over the frequency band of interest. In reality, this is not exactly true; however, for the purpose of providing intuition as to when higher level signaling is beneficial, it is a worthwhile exercise.

Fig. 11 illustrates a simplified view of the insertion loss of a backplane transmission line as a function of frequency. To a first order, this linear approximation on a log-magnitude scale is correct. We assume that the rolloff response can be expressed mathematically as

$$H_L(f) \cong -af \tag{4}$$

where $f =$ frequency, $H_L(f) = 20\log_{10}[|H(f)|]$, $-H_L(f) =$ insertion loss (dB), $a \in \mathbf{R}$, and $a > 0$.

For duobinary and PAM-4, we have a 2× bandwidth compression and, thus, according to Fig. 11, we must correct for the degradation up to a frequency of 1/2 f_0. For NRZ, twice the data bandwidth is required, thus, correction must be used up to f_0. We will assume the same transmitted power (P_{IN}) for each of the three signal formats, i.e., NRZ, PAM-4, and duobinary. Fig. 11 illustrates the reasoning behind our simple derivation of the receiver SNR. For the purpose of adding no additional power sources in our analysis, a passive pre-emphasis circuit is used to flatten out the required bandwidth (1/2 f_0 for duobinary and PAM-4 and f_0 for NRZ). When we look at the cascade of the pre-emphasis filter and the backplane, we have a flat response over the corresponding bandwidth, as illustrated in Fig. 11. The resulting insertion losses are $af_0/4$ dB for the illustrated multilevel formats and $af_0/2$ dB for NRZ. We will now derive the resulting SNR at the receiver. As illustrated in Fig. 11, we will assume a noise power of N, where N is white Gaussian noise, and the following signal levels at the output of the electrical backplane (or, equivalently, the input of the receiver) for multilevel and NRZ formats, respectively:

$$\begin{aligned} P_{\text{out-multi}} &= P_{\text{IN}} - \frac{af_0}{4}\ (\text{dB}) \\ P_{\text{out-NRZ}} &= P_{\text{IN}} - \frac{af_0}{2}\ (\text{dB}). \end{aligned} \tag{5}$$

It is now very simple to calculate the SNR at the receiver input as follows for multilevel and NRZ formats, respectively:

$$\begin{aligned} P_{\text{out-multi}} - N &= \text{SNR}_{\text{multi}}\ (\text{dB}) \\ P_{\text{out-NRZ}} - N &= \text{SNR}_{\text{NRZ}}\ (\text{dB}). \end{aligned} \tag{6}$$

By substituting (5) into (6) and subtracting the resulting expressions, we can obtain an expression for the multilevel SNR *relative* to the NRZ SNR at the receiver input

$$\begin{aligned} \text{SNR}_{\text{multi}} - \text{SNR}_{\text{NRZ}} &= P_{\text{out-multi}} - P_{\text{out-NRZ}} \\ \Delta\text{SNR}_{\text{multi/NRZ}} &= \frac{af_0}{4}\ (\text{dB}). \end{aligned} \tag{7}$$

Equation (7) is an estimate for the relative SNR of both multilevel formats relative to NRZ measured at the receiver given that the input signal powers in Fig. 11 are equal. Clearly, as the slope gets steeper (a increases), multilevel signaling's SNR advantage over NRZ increases. This makes sense since the wider band NRZ signal is affected more by the backplane rolloff (see Fig. 7), while the noise at the receiver is constant.

It is important to note that even when the transmission channel is perfectly flat, in order to achieve the same BER, the required SNR at the receiver for duobinary is ~2 dB more than NRZ while PAM-4 requires ~7 dB more than NRZ [23]. This effect works against the advantage illustrated in (7). To compare the formats fairly for the same BER, we must subtract the appropriate SNR penalty (relative to NRZ) for both duobinary and PAM-4. The result is an expression that relatives the relative SNR of duobinary to NRZ and PAM-4 to NRZ, respectively,

$$\begin{aligned} \Delta\text{SNR}_{\text{duo/NRZ}} &\cong \frac{af_0}{4} - 2\ \text{dB} \\ \Delta\text{SNR}_{[\text{PAM-4}]/\text{NRZ}} &\cong \frac{af_0}{4} - 7\ \text{dB}. \end{aligned} \tag{8}$$

These are very approximate expressions intended only to give insight into the impact of the backplane slope on the relationship between NRZ SNR and duobinary and PAM-4 SNR, as clearly there are numerous implementation details that may have significant effects on this comparison. For example, in our current implementation for duobinary transmission, we launch NRZ data and let the pre-emphasis filter and backplane shape the signal. As a result, our transmitted data occupies a bandwidth of f_0, not 1/2 f_0, although the upper half of the data spectrum is not necessary and will have no impact on the resulting BER performance. A higher order FIR filter could be used to remove this upper part of the spectrum to improve SNR performance and approach the result shown in (8).

When the slope $a = 0$, we have a relative SNR for duobinary of -2 dB with respect to NRZ, while for PAM-4, the relative

TABLE III
OPTIMUM TWO-TAP FIR SETTINGS FOR EYE-OPENING COMPARISONS

Tyco Quadroute Traces	NRZ	Duobinary	PAM-4
6" Link (#1)	20 ps, -2.6 dB	***200 ps, -18.6 dB***	180 ps, -16.0 dB
20" Link (#5)	25 ps, -1.4 dB	***160 ps, -9.8 dB***	150 ps, -8.6 dB
34"Link (#8)	30 ps, -1.1 dB	***75 ps, -2.9 dB***	65 ps, -2.4 dB

SNR is -7 dB with respect to NRZ. As the slope gets steeper, the multilevel formats will perform better than the NRZ.

In addition to the bandwidth discussion, a very important effect not addressed above must be considered when comparing multilevel signaling with NRZ signaling; specifically, the stub effect. Many legacy backplane systems suffer from resonances in the passband, which result from the resonant behavior of unterminated via-holes [16], [17], [21]. This problem is commonly referred to as the stub effect. In the above argument, we assumed that the backplane rolloff is monotonic in nature. In a typical legacy system, resonances will occur over the passband and will, in general, cause significant degradation in the eye performance. The advantage of multilevel signaling is that, by compressing the data bandwidth, we try to contain most of the significant signal energy below the location of these resonances. Since the 3-dB bandwidth of PAM-4 and duobinary are half that of NRZ, in many cases this advantage can be achieved. A detailed analysis of this effect will be the subject of future study.

B. A Simulation-Based Comparison of Eye Openings Between NRZ, PAM-4, and Duobinary Given Amplitude Limited Inputs

Here, we will compare the eye diagrams of NRZ, PAM-4, and duobinary resulting from transmission through a Tyco Quadroute backplane given that the available peak-to-peak excitation amplitude from the transmitter is limited to a constant value. We have chosen a peak-to-peak amplitude of 600 mV to represent a typical integrated circuit available output swing. Importantly, we have determined the optimum two-tap FIR filter to pre-emphasize each format in such a way as to minimize BER performance using a semianalytic analysis technique, as described earlier (see Table III). The optimum filter is determined for each of the three formats and included as part of the transmitter. We then compare the output eye diagrams from the resulting simulations (see Fig. 12) and compare the minimum eye openings for each format (see Table IV). Note that, in Fig. 12, the duobinary eye is larger for both the 34- and 6-in traces; however, the benefit with the 34-in trace is much more significant. In summary, we see that, for the case where the transmitter outputs are limited to identical peak-to-peak output voltage swings, the duobinary format has the largest eye openings for all three trace lengths.

C. Comments on the Asymmetric Properties in the Duobinary Eye

Compared to NRZ and PAM-4, the duobinary eye has two noticeable asymmetries that are worth discussing. The first asymmetry is with respect to the eye shape. While the PAM-4 and NRZ eyes appear more like an oval, the duobinary eyes are somewhat triangular in shape. In general, the optimum decision thresholds will not occur at the point of the maximum horizontal eye openings and, thus, as the decision thresholds approach the "point" of the triangular opening, this signaling technique becomes more sensitive to severe timing jitter, possibly more so than NRZ or PAM-4. As such, it is important to determine the noise dependence of the decision threshold. The second asymmetry is in the probability distribution of the three signal levels. Random data are distributed in an NRZ format with a probability of 0.5 for each bit level, while it is 0.25 for each bit level in the case of PAM-4. For duobinary signaling, random data are distributed with a probability of 0.25, 0.5, and 0.25 among the three bit levels or, in other words, the sum of probabilities of the two outer levels is equal to the probability for the midlevel. When detecting a three-level duobinary signal, two independent decisions are made, one for each the lower and upper eye. The total bit error probability is the sum of bit error probabilities from these two decisions. Due to the asymmetric probability distribution, the optimum decision threshold levels for the upper and lower eye of the duobinary signal become a function of the noise level. We have investigated this by means of simulations to determine how much the decision thresholds vary. To accomplish this, we have modeled an NRZ signal that was transmitted through a 3.0-GHz eighth-order Bessel filter, as shown in

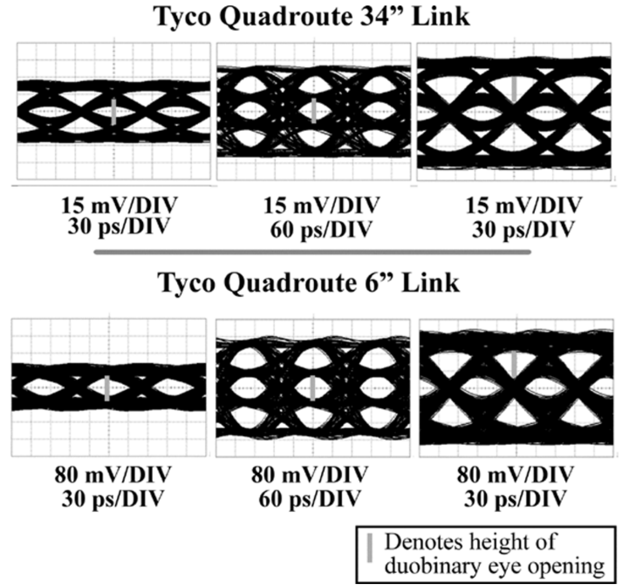

Fig. 12. Simulated eye openings (PRBS 10) resulting from noise- and jitter-free 600-mVpp signals (optimally pre-emphasized with a two-tap fractional FIR filter) through a Tyco Quadroute backplane using NRZ (left-hand side), PAM-4 (center), and duobinary (right-hand side) signaling.

TABLE IV
MINIMUM EYE OPENINGS THROUGH A TYCO QUADROUTE BACKPLANET

	6" Link (#1)	20" Link (#5)	34" Link (#8)
NRZ	78.3 mV_{pp}	37.4 mV_{pp}	10.2 mV_{pp}
Duobinary	***125.6 mV_{pp}***	***82.9 mV_{pp}***	***24.9 mV_{pp}***
PAM-4	100.8 mV_{pp}	53.1 mV_{pp}	15.3 mV_{pp}

The assumptions stated in Fig. 12 are assumed here as well.

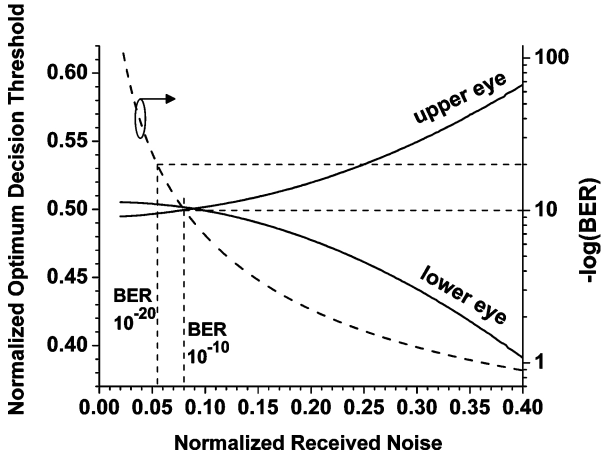

Fig. 13. Optimum threshold levels for duobinary and corresponding BER both as function of normalized receiver input noise level.

Fig. 4. We have optimized the decision threshold and measured the BER as a function of the standard deviation of the colored Gaussian noise that was presented to the receiver. These results are plotted in Fig. 13. The noise voltage was normalized by dividing its rms amplitude, or standard deviation, by the vertical eye opening (horizontal axis in Fig. 13), as were the two decision threshold levels (vertical axis in Fig. 13). We can see that, for a reasonable range of BER values (10^{-10}–10^{-20}), the two decision levels are both close to the expected vertical center of each eye. Increasing the noise level in this range changes the normalized decision thresholds by only $+0.2\%$ for the upper and -0.2% for the lower eye, respectively. Only very high BER values lead to a significant change of decision levels due to the asymmetric probability distribution of the bit levels and, thus, for typical communication links running at low BERs, this phenomenon should not have a major impact.

VII. Conclusion

We have demonstrated a new approach for providing high-speed data transmission through electrical backplanes using a duobinary signaling approach and illustrated the performance with experiments at 10 Gb/s. We have demonstrated BER $< 10^{-13}$ on FR4 backplane traces up to 34 in. We reshape the data spectrum using a FIR filter such that the data at the receiver is duobinary. Additionally, we have developed an innovative duobinary-to-binary data converter that has been demonstrated at 10 Gb/s and whose operation can be extended to 40 Gb/s using higher speed logic.

We have compared the currently popular approaches, NRZ and PAM-4, with our proposed duobinary technique using a two-tap fractional FIR filter and have shown that, for a fixed amplitude eye opening, duobinary has the larger eye at the output of a backplane. Additionally, we have provided an intuitive explanation that gives the reader insight as to when duobinary will perform favorably when compared to NRZ.

The demonstrated system architecture has the potential for providing the advantages of active compensation techniques with less power and less circuit complexity than required by PAM-4. As a result, the duobinary signaling approach may provide a path to obtaining performance at 10 Gb/s and beyond for long-trace backplanes in highly integrated environments such as those required in routers and cross-connect switches.

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