A 100 Gbps True Full-Duplex Link with Interference Cancellation in the Background

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Abstract—In this paper, we introduce the concept of true full-duplexing (TFD) for high-speed interconnects. Full-duplex transceivers at the two ends of an interconnect can support simultaneous bidirectional data transfer. Full-duplexing can help in increasing the throughput per lane and also enable higher data transfer bandwidths without an increase in the routing density. The proposed TFD approach allows for the use of independent modulation schemes as well as independent baud-rates by the transceivers on either side of the interconnect. The approach also obviates the requirement of a dedicated start-up protocol or a training sequence for cancellation of near-end interference in the TFD transceivers. In this work, we have used a correlation-based cancellation technique which can support background cancellation of the interference. Here, we demonstrate a 100 Gbps TFD link over a 1 m long coaxial cable with off-the-shelf components, which is the first demonstration of a full-duplex communication link at such a high data rate.

Index Terms—Adaptive echo cancellation, correlation, full-duplex, high-speed serial links, self-interference, simultaneous bidirectional, true full-duplex (TFD).

I. INTRODUCTION

High-performance computation engines are required for artificial intelligence, big data, and deep-learning architectures. These architectures use high-speed data interconnects to interface many high-performance modules like multi-core processors, high-bandwidth memories, graphics processors, network interfacing cards, and machine learning accelerators [1]. The performance of digital circuitry in these modules continuously improves with device scaling. The modules can process a large amount of data at a rapid pace and require very fast data exchanges. A significant effort is being made to further enhance the performance of the engines consisting of multiple such modules using heterogeneous integration (HI) [2]. With HI, modules with highly optimized individual performances, termed as chiplets, are integrated either over a high-density package or on a common substrate to achieve data exchange throughputs as high as (0.1-1) Tbps/mm (die-edge) [3], [4]. To match this increase in the data exchange rate among the system modules, interconnects with better throughputs and lower latencies are required.

To meet such throughputs, multi-Gbps simplex and duplex interconnect standards are being developed. Among the simplex solutions, AIB-Plus, BoW, BoW-T, PCIe 6.0, and Kandou bus are recent short reach standards [5]–[8], whereas, USB 3.2, SATA 3.5, and Thunderbolt 3 are the long-reach standards [9]–[11]. These simplex solutions use either a large number of slow

lanes [5], [7], a small number of high-speed lanes [6], [9]–[11], re-configurable transceivers [12]–[14], chord signalling (bit-wire encoding) [8], or multi-level signalling schemes [15]. In the simplex standards, except for the Kandou bus where a chord signalling scheme is used, crosstalk is a critical issue affecting signal integrity. The crosstalk or coupling from the nearby aggressors becomes even more severe with an increase in the routing density or operating baud rate [16]–[22]. The reduction of the crosstalk in the simplex standards demands either an increased lane spacing or a decreased operating baud rate, thus lowering the throughput/die-edge.

An alternative solution to reduce the crosstalk without affecting the throughput is to support simultaneous bidirectional (SBD) transfer of data on the same interconnect. Full-duplex (FD) or SBD standards support simultaneous data transmission and reception with a transceiver on either side of the electrical interconnect. These bidirectional standards, such as Ethernet [23] and BoW-BiDi [7] utilize a channel's bilateral nature to reduce the lane density by 50%, and thus lowering the effect of the crosstalk. The SBD transceivers suffer from the presence of interference from their own transmitters. The transceivers receive a composite signal $S_{TR}(t)$ at the input-output (IO) port. $S_{TR}(t)$ contains (i) a far-end (FE) signal S_F and (ii) an interference signal S_{NI} - due to the near-end (NE) transceiver. The S_{NI} includes a strong self-interference (SI) and echoes due to reflections from locations in the link with impedance discontinuities. The presence of S_{NI} significantly affects the bit-error-ratio (BER) at the transceiver, and therefore, it must be suppressed to faithfully extract the FE information from $S_{TR}(t)$ without much penalty on the link performance.

Most interference cancellation approaches [7], [23]–[38] use hybrids for suppressing the SI, and digital domain cancellers for minimizing the echoes and any remnants of the SI (post suppression). The use of the hybrid relaxes the dynamic range requirements of the analog-to-digital-converter (ADC) used by the transceiver. The ADC performs the required digitization, to further minimize the interference digitally. The digital interference cancellers use adaptation engines based on the least-mean-square (LMS) or sign-sign LMS (SSLMS) algorithms. The engines are used for suppressing the time-dependent variations in the amplitude, baud-rate, and pulse shape of the NE transmitted signal. The digital cancellers need synchronization or clock forwarding between the two transceivers in a master-slave configuration [35]. The dependence of these cancellation approaches on the FE transceiver makes calibrating the hybrid and training the canceller challenging, especially when the FE transmitter is active. This limits the utility of these approaches to systems that allow frequent training.

Fig. 1: A general approach for interference cancellation in high-speed SBD links to extract the information transmitted by the far-end (FE) transceiver. Here, L_i - i^{th} reflecting location of the link; P_m - the main signal path; P_c - the interference cancellation path; LD - line driver. Various components of the NE interference i.e. the SI, SSI, and echoes are shown using dashed lines.

TABLE I: List of signal abbreviations used

S_{TR}	Composite signal	rSI	Residual SI
S_F	Far-end signal	S_{NI}	Near-end interference
$T[n]$	Near-end digital data	S_C	Interference cancellation tap
S_{TA}	SI-suppression tap	S_T	Near-end transmitted signal
$F[n]$	Far-end digital data	S_{RIC}	Received far-end signal post interference cancellation
S_{FR}	Recovered and equalized far-end signal	S_{RI}	Received far-end signal with interference

To overcome the limitations in full-duplexing mentioned above, we propose the concept of true full-duplexing (TFD) for SBD links. In TFD links, cancellation for S_{NI} is independent of the baud rate and the signalling scheme used in S_F . The TFD link uses a correlation-based interference cancellation technique that allows for background cancellation of the SI and echoes present in $S_{TR}(t)$, even in the presence of S_F [39], [40]. The interference canceller, thus doesn't need a training sequence and can be trained with an active FE transmitter. Using this technique, we demonstrate a TFD link with an aggregate data rate of 100 Gbps over a 1 m long co-axial cable. To the best of our knowledge, the demonstrated link has the highest bidirectional throughput achieved over a single-lane.

The rest of the paper is organized as follows. The list of signal abbreviations used is provided in Table I. We review various SI and echo cancellation techniques and their limitations in Section II. Section III presents our correlation-based interference cancellation approach used in the TFD link. Section IV describes the measurement setup and the results obtained from a 100 Gbps TFD link. Section V concludes this work.

II. SI AND ECHO CANCELLATION IN CONVENTIONAL SBD LINKS

In SBD links as shown in Fig. 1, $S_{TR}(t)$ at the IO port of Transceiver-NE comprises of a strong SI and echoes due to the NE transmitter, and a weak but desired S_F from Transceiver-FE. Even though the SI and echoes are a form of the NE transmitted signal but have different origins, strengths, time-spans, and delays. These differences in the two interferers require different approaches for their cancellation.

The echoes are reflections of $S_T(t)$ from the channel and transceiver/package interfaces. The strength and instant of the echo occurrence depend on the operating baud-rate, channel

characteristic, channel length, reflection coefficient of the interface, and round-trip delay from the reflecting interface. With these dependencies, the echoes get dispersed in time and magnitude. This dispersive nature of the echoes makes an all analog echo cancellation approach power inefficient. The approach requires many high-speed amplifiers and delay elements, which consumes a lot of power [41]. Therefore, the echoes are cancelled using an adaptive digital echo cancellation approach along with an ADC and a set of digitally delayed transmitted taps. On the other hand, the SI is a strong and slightly delayed form of $S_T(t)$, and can be suppressed efficiently in the analog domain by using a hybrid and few delay elements.

Conventionally, to suppress the SI, the SBD links with low operating baud-rates use hybrids with passive components such as T-coils or magnetic bridges [42]–[44]. In contrast, hybrids for high baud-rate links use active components [7], [23]–[35]. These hybrids need tuning or calibration to matched delays and frequency responses of the main signal path P_m and the interference cancellation path P_c , as shown in Fig. 1. Mismatches in the spectral responses or delays in the two paths may arise due to variations in the interfacing conditions, load impedance, process, voltage, and temperature. These mismatches result in partial cancellation of the SI, leaving a “suppressed-SI” (SSI) in $S_{RI}(t)$ at the input of the transceiver ADC. In the SBD links, the hybrids enable the use of a low dynamic range ADC by the transceiver. Henceforth, a power efficient interference cancellation approach is to:

- (i) Suppress the SI present in $S_{TR}(t)$ to SSI in the analog domain - using a hybrid and a replica of the transmitted data $T[n]$.
- (ii) Digitize the SSI and echoes along with S_F present in the received signal $S_{RI}(t)$, using a low dynamic range ADC - to reduce power consumption of the SBD transceiver.
- (iii) Minimize the NE interference present in the $S_{RI}[n]$ using an adaptive digital SI and echo canceller (SEC) to obtain $S_{RIC}[n]$.

To recover the FE information, most digital cancellers used for minimizing the SSI and echoes, apart from initialization also need synchronization between Transceiver-NE and Transceiver-FE [7], [23]–[35]. Here, we propose the use of TFD in high-speed links as well as in high-density interconnects, where these requirements for minimization are obviated by the use of a correlation-based SEC.

Fig. 2: Block diagram of the proposed TFD transceiver. Here, Z_T - termination impedance; c_k - k^{th} coefficient of the cancellation filter; $T[n]$ - data input; $T_k[n]$ - k^{th} delayed copy of $T[n]$; Σ - accumulator; μ - loop update rate; and L - length of the cancellation filter used by the correlation-based SEC.

III. CORRELATION-BASED INTERFERENCE CANCELLATION IN THE PROPOSED TFD LINKS

The architecture of the TFD transceiver used on each side of the proposed TFD link is shown in Fig. 2. The transceiver comprises: (i) a transmitter, which has a data source, a pulse shaping filter, and a digital-to-analog-converter (DAC) for signal transmission; (ii) an ADC as the receiver front-end for signal reception and digitization; and (iii) a correlation-based SEC for the interference cancellation. The transceiver, first digitizes $S_{TR}(t)$ ¹ using an ADC and then re-samples it to obtain $S_{RI}[n]$. The digitized signal $S_{RI}[n]$ contains the FE information as well as the NE interference. From $S_{RI}[n]$, $S_{FR}[n]$ is then recovered post interference cancellation and equalization, using the correlation-based SEC and an equalizer, respectively.

A. Correlation-based SI and echo canceller

Major challenge in the TFD operation is to suppress the interferers present in $S_{TR}(t)$ without losing the simultaneously received FE information. The proposed TFD link suppresses S_{NI} by using a correlation-based cancellation approach to separate-out the FE data from $S_{TR}(t)$. The cancellation approach makes use of an adaptive Wiener filter [45] and the following two facts: (1) the inherent correlation that exists in between S_T and S_{NI} (the SI and echoes being delayed versions of S_T), for minimizing the latter; (2) the practically uncorrelated nature of the information generated by the two independent transceivers, for separating-out S_F . The filter determines the correlation coefficient c_k as given by Eq.(1), which represents the degree of correlation present in between $S_{RI}[n]$ and $T_k[n]$, and is used for minimizing this correlation. For uncorrelated NE and FE signals, this correlation-based interference cancellation approach is very helpful and unique. Using this approach, the interference cancellers can adjust

their coefficients even in the presence of S_F . The correlation-based approach, thus enables background cancellation for the interference in the TFD link.

$$\begin{aligned}
 c_k[n] &= c_k[n-1] + \frac{1}{N} \sum_{k=0}^{N-1} \{S_{RIC}[n] \times T_k[n]\}, \\
 &= c_k[n-1] + \frac{1}{N} \sum_{k=0}^{N-1} (S_F[n] \times T_k[n]) \\
 &\quad + \frac{1}{N} \sum_{k=0}^{N-1} (S_{NI}[n] \times T_k[n]) \\
 &\quad - \frac{1}{N} \sum_{k=0}^{N-1} (S_C[n] \times T_k[n]), \\
 &= c_k[n-1] + r_{\{S_F, T_k\}} + r_{\{S_{NI}, T_k\}} - r_{\{S_C, T_k\}}, \\
 &= c_k[n-1] + \Delta c_k[n], \tag{1}
 \end{aligned}$$

where, N - is length of the correlator, $r_{\{X, Y\}}$ - denotes the correlation between X and Y , and k - is the filter tap index. For interference cancellation, $r_{\{S_{NI}, T_k\}}$ should be equal to $r_{\{S_C, T_k\}}$, and $r_{\{S_F, T_k\}}$ should be low. The proposed correlation-based SEC generates a replica interference cancellation tap $S_C[n]$ as given by Eq. (2) to cancel S_{NI} present in $S_{RI}[n]$.

$$S_C[n] = \sum_{k=0}^{L-1} \{c_k[n] \times T_k[n]\}, \tag{2}$$

$$\begin{aligned}
 S_{RIC}[n] &= S_{RI}[n] - S_C[n], \\
 &= S_F[n] + r_{SI}[n], \tag{3}
 \end{aligned}$$

where, L - is length of the cancellation filter and k - is the filter tap index. For the cancellation, the SEC adaptively minimizes $r_{\{S_{RIC}, T_k[n]\}}$, and hence determines the correlation coefficient(s) $c_k(s)$ using Eq.(1), (2), and (3). As the SEC tries to minimize only $r_{\{RIC, T_k[n]\}}$ and not S_{RIC} , it preserves the uncorrelated components of S_{RIC} from cancellation. This cancellation, thus eliminates the interference and helps in re-

¹ Here, as an ADC is used directly as the receiver front-end without a hybrid, $S_{TR}(t)$ and $S_{RI}(t)$ are equivalent.

TABLE II: Cancellation modes supported by the correlation-based SEC

Cancellation mode	NE transmitter	FE transmitter	Correlation-based SEC	SEC training type
DPT	On	Off	On	Pre-trained
DBG (Proposed)	On	Don't care	On	Background

Here, DPT - digital pre-trained cancellation; and DBG - digital background cancellation.

Fig. 3: System model for the SEC along with $S_{RIC}[n]$ for the supported cancellation mode (a) DPT and (b) DBG.

covering $S_{FR}[n]$ from $S_{RIC}[n]$, wherein, some residual-SI (rSI) might also be present in case of an incomplete interference cancellation.

The proposed SEC supports interference cancellation in the following two modes as listed in Table II.

- (i) **Digital pre-trained cancellation (DPT):** In this mode, to determine the required values of all $c_k(s)$ the FE transmitter is deactivated; hence, $S_{RI}[n]$ contains only $S_{NI}[n]$. To cancel $S_{NI}[n]$, the SEC tries to generate a $S_C[n]$ that mimics $S_{RI}[n]$ by minimizing $S_{RIC}[n]$ down to system's noise floor, as shown in Fig. 3 (a). Updates in the value of any c_k depends only on the correlation between $S_{RIC}[n]$ and $T_k[n]$ as given by Eq. (4). For a stationary system, the correlation remains constant and independent of N . In such a scenario, the equivalent correlation can also be determined using the conventional LMS algorithm [46] as given by Eq. (5). In this mode, the SEC needs training at a rate similar to which the characteristics of the channel varies.

$$\begin{aligned} \Delta c_k[n] &= \frac{1}{N} \sum_{k=0}^{N-1} S_{RIC}[n] \times T_k[n], \\ &= r_{\{S_{RIC}, T_k\}}. \end{aligned} \quad (4)$$

$$\Delta c_k[n] = S_{RIC}[n] \times T_k[n]. \quad (5)$$

- (ii) **Digital background cancellation (DBG):** In this mode, the required values of $c_k(s)$ are determined in presence of the FE transmission, i.e., $S_{RI}[n]$ contains both $S_{NI}[n]$

TABLE III: Test cases for Matlab® simulations

Test case	Transmitter	Data sequence	Signalling scheme	Cancellation mode
Case1t	NE	PRBS-11	PAM4	DPT
	FE	Off	NA	
Case1a	NE	PRBS-11	PAM4	Off
	FE	PRBS-15	PAM4	
Case1b	NE	PRBS-11	PAM4	Off
	FE	PRBS-32	PAM4	
Case2a	NE	PRBS-11	PAM4	DBG
	FE	PRBS-15	PAM4	
Case2b	NE	PRBS-11	PAM4	DBG
	FE	PRBS-32	PAM4	
Case3	NE	PRBS-11	PAM4	DBG
	FE	PRBS-15	NRZ	

Here, NRZ - non-return-to-zero; PAM4 - pulse amplitude modulation with 4 signalling levels.

Fig. 4: Simulation setup used for validating the correlation-based interference cancellation approach in Matlab®. Here, $F[n]$ - FE transmitted data and Q - Quantizer.

Fig. 5: Various signals of the correlation-based SEC for different test cases as simulated in Matlab®.

and $S_F[n]$. To cancel $S_{NI}[n]$, the SEC generates a $S_C[n]$ by minimizing $r_{\{RIC, T_k[n]\}}$. Any update in the value of a c_k thus, depends only on the correlation between $S_{RIC}[n]$ and $T_k[n]$ as given by Eq. (4). The minimization, hence reduces $S_{RIC}[n]$ down to $S_F[n]$, as shown in Fig. 3 (b). This mode supports background training for the SEC, independent of $S_F[n]$ and variations in the channel characteristics.

B. Validation using simulations in Matlab®

Simulations were performed in Matlab® for validating the used correlation-based interference cancellation approach. The simulation setup is shown in Fig. 4, where Transmitter-NE and

Fig. 6: Coefficients of the correlation filters used for generating the replica interference cancellation tap $S_C[n]$ in different test cases simulated in Matlab[®]. The filters constructed for the cancellation using the two SEC modes show a good conformal matching.

Transmitter-FE, respectively generates the NE data $T[n]$ and the FE data $F[n]$. The setup also comprises: (i) a quantizer (Q) used as the receiver front-end; (ii) a matched filter as the receiver; (iii) the correlation-based SEC for suppressing the interference; and (iv) two scalars with gains α and $1-\alpha$. The scalars are used to vary the fractions of the quantizer dynamic range occupied by $S_F(t)$ and $S_{NI}(t)$. The functionality of the SEC has been validated, using the test cases as listed in Table III, where, PRBS-XX² is a pseudo-random-bit-sequence of length $2^{XX}-1$ bits.

Fig. 5 shows various signals of the correlation-based SEC in different test cases. The SEC faithfully recovers $S_{FR}[n]$ from $S_{RI}[n]$ in both cancellation modes, independent of the modulation scheme used by Transmitter-FE. Correlation filters used for the recovery are shown in Fig. 6. These correlation filters, constructed in the two cancellation modes, show a good conformal matching. The filters were constructed to generate the $S_C[n]$ s used for cancelling the NE interference, i.e. $(1-\alpha) \times T[n]$.

For the test cases, case1 and case2, BER is estimated statistically for $S_{FR}[n]$, using Eq.(6), as given in [46]. The estimated BERs for the test cases are plotted in Fig. 7. The error performance has been estimated for a system ENOB (effective-number-of-bits) of 5.5 bits.

$$\text{BER}_{\text{avg}} = \frac{3}{4} \times Q(\sqrt{0.8 \times \text{average SNR per bit}}), \quad (6)$$

where, SNR is the ratio of the signal power and noise power. The plots show an improvement in the BER as the strength of the interference decreases for all the cases. This improvement in the BER is due to the availability of more resolution bits for the FE signal.

The BER curves for casel1a and casel1b, where a pre-trained filter (constructed in casel1t) has been used for the cancellation, show similar error performance irrespective of the used FE data sequence. On the other hand, the BER plot for case2b shows an error performance similar to casel1a/casel1b but better than case2a. The performance improves because case2b has a lower cross-correlation between the FE and NE data sequences as compared to case2a. Thus, as the cross-

Fig. 7: Estimated BER for the recovered FE signal $S_{FR}[n]$. The BER improves with a decrease in the strength of the interference. The plots show an improvement in the BER for the background cancellation mode as the cross-correlation between the FE and the NE data decreases.

TABLE IV: List of the TFD links demonstrated.

TFD link	Transmitter-NE		Transmitter-FE		Aggregate data rate
	SQ ₁ : PRBS-11		SQ ₂ : PRBS-15		
	Baud-rate (data-rate)	Signalling scheme	Baud-rate (data-rate)	Signalling scheme	
LA _{PAM4}	25 Gbaud (50 Gbps)	PAM4	25.6 Gbaud (51.2 Gbps)	PAM4	101.2 Gbps
LB _{PAM4}	25 Gbaud (50 Gbps)	PAM4	25 Gbaud (50 Gbps)	PAM4	100 Gbps
LC _{PAM4}	20 Gbaud (40 Gbps)	PAM4	20 Gbaud (40 Gbps)	PAM4	80 Gbps
LD _{NRZ}	32 Gbaud (32 Gbps)	NRZ	32 Gbaud (32 Gbps)	NRZ	64 Gbps

TABLE V: System configurations for measurements.

System configuration	Ch ₁ (Transmitter-NE)	Ch ₁ (Transmitter-FE)
SC1	On	On
SC2	Off	On

correlation between the FE and NE data decreases, the two operating modes of the SEC, i.e. DBG and DPT, have similar error performances. Therefore, with uncorrelated NE and FE signals, the correlation-based SEC can reliably operate in the background - to eliminate S_{NI} independent of modulation-schemes used by the two transceivers. This approach, thus obviates the need to have a dedicated startup protocol for the two transceivers used in the TFD links, without compromising on the link error performance.

IV. MEASUREMENT SETUP AND RESULTS

Measurements have been performed for demonstrating the TFD links as given in Table IV, over a 1 m long PE-P160 coaxial cable. The measurement setup, as shown in Fig. 8, uses a correlation-based SEC for the interference cancellation in the digital domain. The digital domain cancellation was performed on the signals received, with and without the SI-suppression performed at the hardware level. The TFD links have been demonstrated for a channel response as shown in Fig. 9, using bench-top equipment, which include

- A Keysight DSOZ204A Infiniium oscilloscope with measurement bandwidth of 20 GHz – used for receiving $S_{RI}(t)$.

²The higher the difference between XX values of the two PRBS sequences lesser will be the cross-correlation present in between the two.

Fig. 8: Measurement setup used for demonstration of the TFD links using off-the-shelf components. The SI has been suppressed in the analog domain using a SI-suppression tap $S_{TA}(t)$, and the SSI and echoes are further cancelled in the digital domain using the correlation-based SEC. Here, $S_{FR}[n]$ - equalized $S_{RIC}[n]$; $S_T(t)$ - NE transmitted signal; $S_F(t)$ - FE transmitted signal; $\bar{S}_T(t)$ - inverted copy of $S_T(t)$; Ch_i - i^{th} channel of the AWG; and A_i - i^{th} used amplifier.

Fig. 9: Response of the channel used for demonstration of the TFD links. Plot shows a relative channel loss, i.e. $S_{31} - S_{32}$, of nearly 10 dB at 12.5 GHz.

- A Keysight M8195A 2-slot AXIe arbitrary waveform generator (AWG) for generation of the NE signal $S_T(t)$ and the FE signal $S_F(t)$, and a SI-suppression tap $S_{TA}(t)$. The $S_{TA}(t)$ is generated using $\bar{S}_T(t)$, and has been used for performing the SI-suppression in the analog domain.

The measurement setup, as shown in Fig. 8, consists of three 50 Ω 3-port resistive broadband power dividers/combiners (BBPD)s. Each of these BBPDs has a bandwidth of 18 GHz and an insertion loss of at least 6 dB. In this setup, the BBPD₁/BBPD₂ combines the desired FE signal $S_F(t)$ and the NE interference i.e. the undesired SI and echoes from the NE transmitter. Whereas, the BBPD₃ was used to perform the SI-suppression, by summing the signals $S_{TR}(t)$ and $S_{TA}(t)$. The setup was operated in the measurement configurations as given in Table V. The number of available resolution bits for the measurements was determined by curve fitting sinusoidal waves of frequency 1 GHz [47]. The system configurations SC1 and SC2, have a lower number of available resolution bits as compared to the standalone oscilloscope, as shown in Fig. 10. The decrease in the number of resolution bits is due to the additional noise received from the DACs of the AWG channels (Ch_1 and Ch_2) and from the amplifiers A_1/A_2 , which were used as buffers and are shown in Fig. 8.

For the measurements, the $S_T(t)$ and $S_F(t)$ were transmitted from the two ends of the channel; and the received signal $S_{RI}(t)$ was sampled using the oscilloscope at a sample rate of 80 GSps. The $S_{RI}(t)$ contains the desired FE information and the unwanted interference due to the NE signal $S_T(t)$. To

TABLE VI: Cancellation modes used in the TFD links.

Cancellation mode	System configuration	$S_{TA}(t)$ - SI-suppression tap	SEC training type
DPT	SC1	Off	Pre-trained
DBG	SC1	Off	Background
ADPT	SC1	On	Pre-trained
ADBG	SC1	On	Background
UD	SC2	Off	Off

Here, DPT - digital pre-trained cancellation; DBG - digital background cancellation; ADPT - analog domain suppression plus digital pre-trained cancellation; ADBG - analog domain suppression plus digital background cancellation; UD - unidirectional.

Fig. 10: Measured value of resolution bits available for the standalone oscilloscope, and of the measurement system in configurations: SC1 and SC2. The decrease in resolution bits for SC1/SC2 as compared to the oscilloscope is due to the additional noise received from the active AWG channels and the amplifiers.

suppress the interference present in $S_{RI}[n]$, the correlation-based interference cancellation was performed using Matlab®, in the modes as mentioned in Table VI. The digital domain cancellation have been performed using pre-trained SECs for the modes DPT and ADPT. While for the modes DBG and ADBG, the SECs operate in the background for the cancellation.

In the cancellation modes DPT/DBG, only the digital domain cancellation has been performed on $S_{RI}[n]$. Whereas, in the cancellation modes ADPT and ADBG, prior to cancellation

Fig. 11: Coefficients of the digital correlation filters used for cancelling the interference present in $S_{RI}[n]$: (a) Without the SI-suppression in the analog domain and (b) with the SI-suppression performed using $S_{TA}(t)$ at the hardware level.

in the digital domain, the SI-suppression was also performed at the hardware level, by summing $S_{TR}(t)$ and $S_{TA}(t)$. The use of $S_{TA}(t)$ helps in re-claiming the dynamic range of the oscilloscope's ADC, which was also occupied by the SI. Post interference cancellation in these modes, the FE signal was then recovered as $S_{RIC}[n]$. The $S_{RIC}[n]$ has been further equalized to obtain $S_{FR}[n]$ in Matlab®. The equalization has been performed to compensate for the channel losses using a feed-forward equalizer. For the performance evaluation of the SECs used in the TFD links, uni-directional (UD) measurements were also performed. In the UD mode, the SEC is turned-off as $S_{RI}[n]$ contains only the FE signal, and needs only equalization to compensate for the channel losses.

For the TFD link LA_{PAM4} , coefficients of the constructed cancellation filters are shown in Fig. 11. To cancel the interference, these filters were used to minimize the correlation present between $S_{RI}[n]$ and the delayed replicas of S_T . The constructed correlation filters for the mode DPT and DBG plotted in Fig. 11(a) show a strong main lobe along with side lobes. The main lobe is due to a strong SI, whereas the side lobes are because of echoes and the pulse shaping filter used for the data transmission. For the modes, ADPT and ADBG, suppression performed in the analog domain using $S_{TA}(t)$ suppresses the main lobe significantly. The filter for the mode ADBG, as shown in Fig. 11(b), because of its larger length, shows a higher variation in the coefficient values as compared to that for the mode ADPT.

Eye-diagrams of the FE signal for the link LA_{PAM4} , with and without interference cancellation and equalization, are plotted in Fig. 12. Fig. 12(a) and Fig. 12(b) show eye-diagrams

Fig. 12: Eye-diagrams for (a) $S_{RI}[n]$ without any interference cancellation, (b) $S_{RI}[n]$ with SI-suppression performed using the $S_{TA}(t)$ in the analog domain, (c)–(g) recovered $S_{RIC}[n]$ post interference cancellation in the digital domain using the cancellation mode as mentioned in the subscript, and (h)–(l) recovered and equalized $S_{FR}[n]$ in the cancellation modes as mentioned in the subscript. These eye-diagrams are plotted for the link LA_{PAM4} , and are for 10^4 symbols of the signals.

for the received $S_{RI}[n]$ with and without analog domain suppression, respectively. Eye-diagrams of $S_{RIC}[n]$ for the four cancellation modes DPT, DBG, ADPT, and ADBG are shown in Fig. 12(c-f), respectively. Eye-diagram of $S_{FR}[n]$ for mode ADPT/ADBG shows a better eye-opening than the modes DPT/DBG, even with a reduction in the number of available resolution bits. This improvement in the eye-quality is because of the SI-suppression performed using $S_{TA}(t)$, which allows the FE signal to re-claim the dynamic range of the oscilloscope's ADC. Fig. 12(g) and Fig. 12(l) show eye-diagrams for the received $S_{RIC}[n]$ with and without equalization in the mode UD, respectively.

Fig. 13 shows bath-tub curves for the equalized $S_{FR}[n]$. The

Fig. 13: Estimated bath-tub curves for the equalized FE signal $S_{FR}[n]$ of the demonstrated TFD links: (a) LD_{NRZ} , (b) LA_{PAM4} , (c) LC_{PAM4} , and (d) LB_{PAM4} . For the links, the cross-correlation between the transmitted sequences SQ_1 and SQ_2 increases from (a) to (d). The links with the lower correlation show a better BER performance for the background cancellation modes.

curves have been plotted using the estimated BERs [46]. The relative BER performances of the demonstrated links for different cancellation modes can be summarized as follows:

Fig. 13(a): $LD_{NRZ} \rightarrow UD < ADPT < ADBG < DBG < DPT$

Fig. 13(b): $LA_{PAM4} \rightarrow UD < ADPT < ADBG < DBG \sim DPT$

Fig. 13(c): $LC_{PAM4} \rightarrow UD < ADPT < DPT < ADBG < DBG$

Fig. 13(d): $LB_{PAM4} \rightarrow UD < ADPT < DPT < ADBG \sim DBG$

The bath-tub curves show an improvement in the error performance with the SI-suppression for all the demonstrated TFD links. The LD_{NRZ} shows a better error performance than the other links because of the used NRZ signalling scheme for the FE and NE signals. The BER curves, for the cancellation

modes, ADPT and DPT, show a drop in the error performances with an increase in the operating baud rates for all links. On the other hand, for the modes ADBG and DBG, the BER curves for LA_{PAM4} (which supports the highest throughput) show improved error performances than both LB_{PAM4} and LC_{PAM4} . This improvement in the error performance for LA_{PAM4} is because of the lower cross-correlation between the transmitted sequences SQ_1 and SQ_2 . The link LA_{PAM4} has a low cross-correlation as transmitting SQ_1 and SQ_2 at different baud rates, lowers their concurrent repetition frequency. This decreased cross-correlation facilitates a better interference cancellation in the background cancellation modes, which is consistent to the simulation results.

The demonstrated TFD link LA_{PAM4} , supports an aggregated bidirectional throughput of 101.2 Gbps over a channel loss of nearly 10 dB at the Nyquist frequency. The bath-tub curves for link LA_{PAM4} show a raw BER better than 10^{-7} and 10^{-5} in the cancellation mode ADPT and ADBG, respectively, meeting the present-day error performance [48]. The demonstrated TFD link, thus supports a data transfer throughput of 100 Gbps over a single-lane, even when the transceivers operate at different baud rates. This independence eliminates the need to have a dedicated clock synchronization procedure or a training sequence for cancellation of the NE interference in the transceivers.

All these measurements were performed using single-ended signals, bandwidth-limited BBPDs, and 12-inches co-axial patch cables, which have been used for connecting the BBPDs to the amplifiers A_1/A_2 . The performance of the SECs used in DBG/ADBG can be further improved by using data sequences with low cross-correlation for the FE and NE data. The error performance of the demonstrated links can be enhanced significantly by using better components or in an integrated transceiver, where these limitations will not be present.

V. CONCLUSION

This paper introduces and validates the concept of true full-duplexing (TFD) for transceivers used in bidirectional links. The use of TFD in transceivers doubles the data throughput per lane and offers a reduction in routing density. The proposed TFD approach allows for the usage of independent modulation schemes and independent baud-rates by the transceivers on either side of the same electrical interconnect. This independence obviates the need of having a dedicated start-up protocol or a training sequence for the used transceivers. The proposed TFD transceivers use a correlation-based technique to support background cancellation of the near-end interference. We have successfully demonstrated a 100 Gbps TFD link over a 1 m long coaxial cable with background interference cancellation, which is the highest bidirectional throughput reported to date for a single-lane interconnects.

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