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

Sandeep Goyal and Shalabh Gupta

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.
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.

---

**TABLE I:** List of signal abbreviations used

| S_{TR} | Composite signal | S_{SI} | 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_{RRC} | Received far-end signal post interference cancellation |
| S_{R}  | Recovered and equalized far-end signal | S_{RI} | Received far-end signal with interference |

---

**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.
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 \( S_{RI} \) and \( S_{NI} \) (the SI and echoes being delayed versions of \( S_{RI} \)), for minimizing the latter; (2) the practically uncorrelated nature of the information generated by the two independent transceivers, for separating-out \( S_{FE} \). 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 \( S_{PE} \). The correlation-based approach, thus enables background cancellation for the interference in the TFD link.

\[
 c_k[n] = c_k[n-1] + \frac{1}{N} \sum_{k=0}^{N-1} \left\{ S_{RIC}[n] \times T_k[n] \right\},
\]
\[
 = c_k[n-1] + \frac{1}{N} \sum_{k=0}^{N-1} \left( S_{F}[n] \times T_k[n] \right)
\]  
\[
+ \frac{1}{N} \sum_{k=0}^{N-1} \left( S_{NI}[n] \times T_k[n] \right),
\]
\[
- \frac{1}{N} \sum_{k=0}^{N-1} \left( S_{C}[n] \times T_k[n] \right),
\]
\[
= c_k[n-1] + r_{S_{PE},T_k} + r_{S_{NI},T_k} - r_{S_{C},T_k},
\]
\[
= c_k[n-1] + \Delta c_k[n],
\]
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 \( r_{S_{PE},T_k} \) 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 \( S_{RI}[n] \).

\[
S_{C}[n] = \sum_{k=0}^{L-1} \left\{ c_k[n] \times T_k[n] \right\},
\]
\[
S_{RIC}[n] = S_{RI}[n] - S_{C}[n],
\]
\[
= S_{F}[n] + r_{S_{PE},n},
\]
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 \( r_{S_{RIC},T_k[n]} \) and not \( S_{RIC} \), it preserves the uncorrelated components of \( S_{RIC} \) from cancellation. This cancellation, thus eliminates the interference and helps in re-

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; \( T_{[n]} \) - \( k \)th delayed copy of \( T[n] \); \( \Sigma \) - accumulator; \( \mu \) - loop update rate; and \( L \) - length of the cancellation filter used by the correlation-based SEC.

\(^1\)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 (Proposed)    | 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 \( S_{RI}[n] \) for the supported cancellation mode (a) DPT and (b) DBG.

covering \( S_{FR}[n] \) from \( S_{RR}[n] \), 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 \( S_{NI}[n] \) by minimizing \( S_{RR}[n] \) 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 \( S_{RR}[n] \) 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.

\[
\Delta c_k[n] = \frac{1}{N} \sum_{k=0}^{N-1} S_{RI}[n] \times T_k[n],
\]

(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 \( S_{NI}[n] \)

\[
S_{FR}[t] = S_{C}[t] - S_{RI}[t],
\]

and \( S_{FR}[t] \). To cancel \( S_{NI}[n] \), the SEC generates a \( S_C[n] \) by minimizing \( \Delta_r[S_{RI}, T_k[n]] \). Any update in the value of \( c_k \) thus, depends only on the correlation between \( S_{RR}[n] \) and \( T_k[n] \) as given by Eq. (4). The minimization, hence reduces \( S_{RR}[n] \) down to \( S_r[n] \), 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. 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®.
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−α. The scalars are used to vary the fractions of the quantizer dynamic range occupied by SFR(t) and SNR(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 2XX−1 bits.

Fig.5 shows various signals of the correlation-based SEC in different test cases. The SEC faithfully recovers SFR[n] from SR[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 SE[n]s used for cancelling the NE interference, i.e. (1 − α) × T[n].

For the test cases, case1 and case2, BER is estimated statistically for SFR[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.

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

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 case1a and case1b, where a pre-trained filter (constructed in case1t) 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 case1a/case1b 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-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 SFI 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 Infinium oscilloscope with measurement bandwidth of 20 GHz – used for receiving SFR(t).

**TABLE IV: List of the TFD links demonstrated.**

| TFD link | Transmitter-NE | Transmitter-FE | Aggregate data rate |
|----------|----------------|----------------|-------------------|
| 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_NIRZ | 32 Gbaud (32 Gbps) | NRZ | 32 Gbaud (32 Gbps) | NRZ | 64 Gbps |

**TABLE V: System configurations for measurements.**

| System configuration | Ch1 (Transmitter-NE) | Ch1 (Transmitter-FE) |
|----------------------|-----------------------|----------------------|
| SC1                  | On                    | On                   |
| SC2                  | Off                   | On                   |

2The higher the difference between XX values of the two PRBS sequences lesser will be the cross-correlation present in between the two.
On the unwanted interference due to the NE signal $S_{TR}$ of 80 GSps. The $S_{TR}$ from the two ends of the channel; and the received signal were used as buffers and are shown in Fig. 8.

The decrease in the number of resolution bits is due to the additional noise received from the DACs of the AWG 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 $S_T(t)$, 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\,\Omega 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, 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 the cancellation modes ADPT and ADBG, prior to cancellation.

- 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 $S_T(t)$, and has been used for performing the SI-suppression in the analog domain.

The setup was operated in the measurement configurations as shown in Fig. 8. Figures 9 and 10 show the measurements on the additional noise received from the DACs of the AWG channels (Ch1 and Ch2) 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 $S_{RF}(t)$ was sampled using the oscilloscope at a sample rate of 80 GSps. The $S_{RF}(t)$ contains the desired FE information and the unwanted interference due to the NE signal $S_T(t)$. To suppress the interference present in $S_{RF}(t)$, the correlation-based interference cancellation was performed using Matlab\textsuperscript{®}, 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 $S_{RF}(t)$. Whereas, in the cancellation modes ADPT and ADBG, prior to cancellation.
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 else 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 $L_{PAM4}$, 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 $L_{PAM4}$, with and without interference cancellation and equalization, are plotted in Fig. 12. Fig. 12(a) and Fig. 12(b) show eye-diagrams 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 $S_{FR}[n]$ with and without equalization in the mode UD, respectively.

Fig. 13 shows bath-tub curves for the equalized $S_{FR}[n]$. The
The FE and NE signals. The BER curves, for the cancellation other links because of the used NRZ signalling scheme for links. The LD
 performance with the SI-suppression for all the demonstrated TFD 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\textsubscript{NRZ} → UD < ADPT < ADBG < DBG < DPT
- Fig. 13(b): LA\textsubscript{PAM4} → UD < ADPT < ADBG < DBG ∼ DPT
- Fig. 13(c): LC\textsubscript{PAM4} → UD < ADPT < DPT < ADBG < DBG
- Fig. 13(d): LB\textsubscript{PAM4} → UD < ADPT < DPT < ADBG ∼ DBG

The bath-tub curves show a better BER performance for the backgroun d 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\textsubscript{PAM4} (which supports the highest throughput) show improved error performances than both LB\textsubscript{PAM4} and LC\textsubscript{PAM4}. This improvement in the error performance for LA\textsubscript{PAM4} is because of the lower cross-correlation between the transmitted sequences SQ\textsubscript{1} and SQ\textsubscript{2}. The link LA\textsubscript{PAM4} has a low cross-correlation as transmitting SQ\textsubscript{1} and SQ\textsubscript{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\textsubscript{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\textsubscript{PAM4} show a raw BER better than 10\textsuperscript{−7} and 10\textsuperscript{−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\textsubscript{1}/A\textsubscript{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.

REFERENCES

[1] J. Macri, “AMD’s next generation GPU and high bandwidth memory architecture: FURY,” in IEEE Hot Chips, 2015, pp. 1–26.
[2] D. S. Green, “Common Heterogeneous Integration and Intellectual Property (IP) Reuse Strategies (CHIPS),” [Online]. Available: https://www.darpa.mil/attachments/CHIPSoverview% 20Sept212016ProposerDay.pdf [Accessed: 20-08-2020].
