Progress In Electromagnetics Research C, Vol. 12, 237–251, 2010

IMPROVEMENT IN DATA TRANSMISSION EFFICIENCY IN COMMUNICATION SYSTEMS USING SCATTERING COMPENSATION TECHNIQUES

S. Raju

Department of Electrical and Electronics Engineering
American International University-Bangladesh
Dhaka-1213, Bangladesh

S. M. Salahuddin and M. I. Raza

Department of Electrical and Electronics Engineering
East West University
Dhaka-1212, Bangladesh

Abstract—The primary requirement for maximum power transfer and minimum power loss is matched impedance in a transmission system. However, static design variations in system such as parasitics and dynamic variations such as changes in transmission frequency will result in reflections. A redesign or reconfiguration of complex systems is neither easy nor cost effective. A noble technique for compensating reflections in a communication system is presented here. The proposed methodology adapts the system to change without any modification to the system physical configuration. In this methodology compensation signals are added by I/O drivers with programmable phase delay and drive strength adjustments to cancel reflections. The concept application is demonstrated for a narrow bandwidth antenna system. An operating frequency off the antenna frequency results in a degraded received message eye. Using the proposed technique, without modifying the antenna, reflections were compensated and a significantly improved data eye was produced, as measured by the enhancement of critical performance parameters. The architecture of an expanded driver to implement the concept is outlined here. An algorithm and flow chart to dynamically identify and compensate for reflections are also presented.

Corresponding author: M. I. Raza (iraza@ewubd.edu).

1. INTRODUCTION

In microwave communication systems transmission lines are used to connect different components of a system, such as antenna, filters, amplifiers, and mixers, both in transmitter and receiver side of a system. Transmission lines can also be viewed as connectors in chips, circuit boards, packages etc. in any digital system. When a signal travels through transmission lines, it is required that the signal holds its integrity while sustaining maximum power transfer. But when impedance mismatch occurs at different interfaces in the transmitting device, receiving device, and the transmission line, standing waves are generated, which will result in loss in the cable. This will eventually decrease the antenna radiation efficiency. Also, the antenna impedance is complex which varies in both magnitude and phase with variations in the operating frequency. There are also manufacturing tolerances which will inevitably introduce impedance mismatches in the system. Given the multitude of possibilities for impedance variations, it is important to build an inherent ability for such communication systems to identify and nullify these reflections. This will ultimately increase the system performance and antenna radiation efficiency and improve overall integrity of propagated message signals.

In digital systems impedance mismatch manifest in the form of degraded signal integrity and inter symbol interference, resulting in an increasing bit error rate. Equalization and pre-emphasis have been used both at the transmitting and receiving side to improve the bit error rates [1–3]. But reflection cancelation technique must be dynamic to accommodate perturbations in system design; otherwise reflections can not be fully compensated as system parameters change. Examples of such perturbations are manufacturing errors and changes in ambient factors, such as temperature and noise. In some microwave systems, impedance matching network are often used. Single stub, double stub and single or multi-section matching transformer are used as matching network [4, 5]. There are some disadvantages to these matching schemes. First, any shift in the operating frequency will result in a change of the impedance of the stub or transformer, i.e., any of this fixed matching network solution can ideally only have a perfect match at a single frequency. Also, there are parasitics associated with the termination scheme which will contribute to the impedance discontinuity, which being primarily reactive, is also a function of the frequency of operation. Its complexity increases when it is desirable to match a load at different operating frequencies. Moreover the terminations also consume power, adding to the power budget. With increasing frequencies, the effects of these attributes grow in significance. Recently, non-uniform microstrip transmission line has been proposed for impedance matching [6]. Several other methods such as coplanar waveguide feed structure have been proposed earlier for impedance matching [7]. Wideband differential phase shifter using microstrip non-uniform transmission line has also successfully minimized reflections [8]. Several other echo cancelation techniques have also been proposed where echo is canceled in high frequency circuits [9, 10]. But the potential drawback of these proposals is that any system parameter changes after fabrication may produce static discontinuities which cannot be addressed. To overcome such shortcomings, a matching scheme should have dynamic adjusting capability. Such dynamic concept has been proposed earlier to compensate digital systems [11]. Here a concept is proposed for high frequency communication systems operating in wireless environments.

In this paper, a reflection cancelation methodology is proposed which has significant improvements over previous techniques. The proposed technique can be used in systems where the operating frequency may vary or system operating parameters may change. This ability to tune the compensation feature in real time, through a simple calibration routine, not only improves the power budget, but also the signal to noise ratio at load end. Similarly, if any system parameter changes beyond design parameters during fabrication, reflections will happen. This can also be addressed with the proposed architecture.

The technique proposes an I/O driver architecture which will have programmable reflection compensation capabilities. The driver will first obtain the signature of the system, looking at the signal waveform at the point where signal quality is of interest. The residual reflections observed at that point of interest can then be compensated with time delayed and scaled transmitted signals. The compensation signals are transmitted from the driver during regular operation of the system using appropriate phase and magnitude.

The nature of the reflection depends on the system, load or topology discontinuity. A part of the source driven signal is sent back to the source driver. Then impedance discontinuity at the driver will result in a part of the load signal to be retransmitted back to load. To reduce these reflections, the driver will be programmed to either source or sink additional energy to cancel either the negative or positive reflection from the source. The compensation signal will be of opposite sign of the signal reflected by the source. The ability to provide measured compensation for reflections at the source also empowers usage of incident wave reflection for greater received signal amplitude.

2. MATHEMATICAL ASPECT

To introduce this methodology, let us first consider a transmission line with a source and a load, as shown in Figure 1. It is driven by a continuous wave (CW) source $V_s(t)$ with an impedance Z_s. The characteristic impedance of the point to point interconnect is Z_C which is terminated with load Z_L.

Impedance discontinuity at source and load ends will introduce reflections from both ends. From the voltage reflection diagram or a lattice diagram the voltage distribution along the transmission line in time domain is calculated. Considering all reflections, the total voltage summed at the source end $z = -l$ and at the load end $z = 0$, is

$$V(-l,t) = \frac{Z_C}{Z_S + Z_C} \left[\sum_{n=0}^{\infty} (1+\Gamma_L)e^{-\gamma l} \left(\Gamma_S \Gamma_L e^{-2\gamma l}\right)^n V_S(t)u(t-(2n+1)T_D)\right] \quad (1)$$

where n is the number of reflections at source side and

$$V(0,t) = \frac{Z_C}{Z_S + Z_C} \left[V_S(t)^+ \sum_{n=1}^{\infty} (1+\Gamma_S)\Gamma_L e^{-2\gamma l} \left(\Gamma_S \Gamma_L e^{-2\gamma l}\right)^{n-1} V_S(t)u(t-2nT_D)\right] \quad (2)$$

where n is the number of reflections at load side.

Here, Γ_L is reflection coefficient at the load side, Γ_S is reflection coefficient at the source side, l is transmission line length, γ is propagation constant and T_D is one way delay of the transmission line. To eliminate reflections from the source end, a signal canceling the first reflection is added at a delayed time. The added signal x, is given by

$$x = -\Gamma_S \Gamma_L e^{-2\gamma l} \frac{Z_C}{Z_S + Z_C} V_S(t)u(t-2T_D) \quad (3)$$

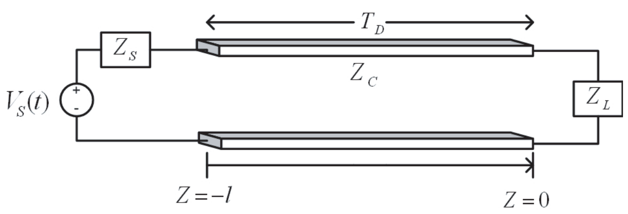

Figure 1. Transmission line terminated with load impedance Z_L.

The compensation signal x is a scaled source signal delayed by time $t = 2T_D$. Until time $t = 2T_D$ only the original signal propagating from the source to load and the first reflection from load to source will exist on the line. However, all other higher order reflections after $2T_D$ are cancelled out with the introduction of the compensation signal x. As indicated above, the definition of the amplitude and phase of the signal x depends on reflection coefficients at source and load, propagation constant and transmission line length.

In this paper we demonstrate this scheme for a communication model based on amplitude modulation. For our scheme we assume a message $m(t)$ modulating the carrier signal $c(t)$. In our example, the message is a low frequency digital signal. The modulated signal,

$$V_s(t) = m(t)A_c \cos(\omega_c t) \tag{4}$$

where the carrier signal is $c(t) = A_c \cos(\omega_c t)$.

To verify our proposed methodology we have implemented this scheme to a worst case situation where impedance mismatch occurs both at source end and load end. As the signal propagates back and forth from the load end and source end, a reflection will persist in the transmission line. The compensation signal for the amplitude modulated signal, modifying Equation (3), will be given by

$$x = -\Gamma_s \Gamma_L e^{-2\gamma l} \frac{Z_C}{Z_S + Z_C} m(t)u(t - 2T_D) \cdot A_c \cos(\omega_c t - 2\omega_c T_D) \tag{5}$$

The driver architecture is proposed here which will add the above signal, as shown in Equation (5), to the source signal.

3. I/O DRIVER MODEL

An architecture is introduced here for the driver designed with the intent to compensate reflections. The schematic description is shown in Figure 2. The circuit has programmable features designed to set the amplitude and delay of the driver compensation signal. The ultimate goal of this driver architecture is to generate the reflection compensation signal as described in Equation (5) and add this signal with the main driver at the proper time.

First, the carrier signal is fed into a dynamic phase shifter which is set by the DREG to produce $A_c \cos(\omega_c t - 2\omega_c T_D)$. Then the analog message signal $m(t)$ is converted into digital data using high speed analog to digital converter (ADC). The sampling frequency of the ADC must be several times the Nyquist rate to ensure an accurate signal trace. The digital data is then fed into a multiplier, in which the multiplying factor $\Gamma_S \Gamma_L e^{-2\gamma l} \frac{Z_C}{Z_S + Z_C}$ is defined based on the magnitude

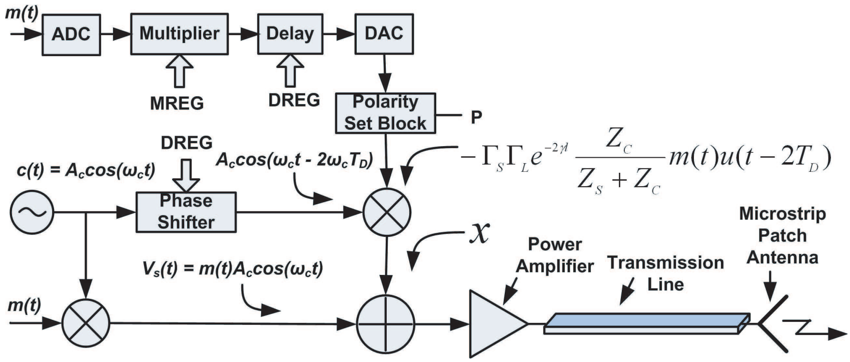

Figure 2. I/O driver model.

register (MREG). As compensation is done after $2T_D$ seconds, those digital data are time shifted by a delay module which is controlled by the delay register (DREG). The digital data is then converted into an analog signal by a high speed digital to analog converter (DAC) before it is added to the source signal. The polarity of compensation signal can be positive or negative depending upon the nature of the discontinuity where the reflection originates. If the polarity of compensation signal x is negative, the polarity pin P is set to 1, then the polarity set block will be used as an inverting amplifier. Otherwise $P = 0$ and polarity is kept as it was. At last the polarity set block adjusts the polarity to acquire the signal

$$-\Gamma_S\Gamma_L e^{-2\gamma l}\frac{Z_C}{Z_S+Z_C}m(t)u(t-2T_D)$$

as shown in Figure 2. This signal is then multiplied by generated phase shifted carrier, $A_c\cos(\omega_c t - 2\omega_c T_D)$ to produce reflection compensation signal x as describe in Equation (5). This compensating signal is then added with the main driver output.

4. SYSTEM CALIBRATION

The transfer function of a system and thus its signature defines the signal propagation and reflections within a system. The accurate signature of the system will however depend on the specific operating conditions, system configuration, and the working environment of the system. It is therefore necessary to calibrate a system prior to activation. During calibration, information regarding reflection compensation is measured, extracted, and stored. Then, the system is initialized prior to operation using the information obtained during calibration. In this instance the multiplier, phase shifter and delay modules are programmed for reflection cancellation.

In the calibration routine, a Gaussian monocycle pulse is proposed for use to measure reflections transmitted back to the driver (Figure 3). The center frequency of the monocycle pulse should be equal to the frequency of the carrier signal used in source driver. As the system component responses are functions of frequency (such as the antenna input impedance being reactive) — the measured reflections will also be dependent on the spectral content (frequency) of the signal source. The spectral content of the signal will not only vary the delay and phase of any scattering but also its strength (reflection coefficient $\Gamma(\omega)$ is a function of frequency) [11]. The measured delay and magnitude obtained from the calibration routine is stored into appropriate registers which are then used to program the I/O driver.

In the calibration routine, the system is characterized at the carrier frequency (unmodulated, i.e., without considering the message signal). The characterization at the carrier frequency is considered sufficient for the following reasons. In a typical communication system, the message signal bandwidth is very small compared to the carrier frequency. For example, in the case of a GSM network, the carrier frequency is 1.9 GHz (ω_c), while the channel bandwidth is 200 kHz (ω_m). For the given example, the system will operate at the following pass band $\omega_c \pm \omega_m$, which will be well within the bandwidth of the relevant antenna. However, the reflection measured at 1.9 GHz will not be significantly different from either 1.8998 GHz or 1.9002 GHz. Also, given that the message signal content is dynamic, it is not possible to compensate the transmitted signal without violating causality or adding delay. Therefore, it is our assertion that using the carrier signal to calibrate the system response is sufficient to accurately compensate for scattering effects, which is also supported by our calculated results, as shown in the later section.

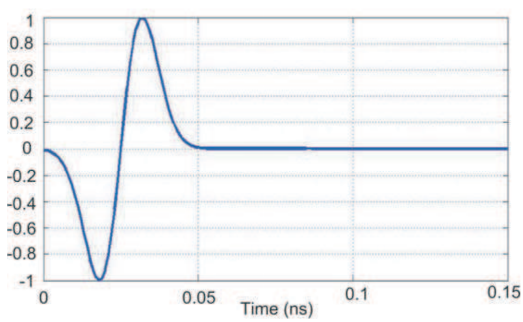

Figure 3. Gaussian monocycle pulse.

A flow chart on how to design the calibration and initialization sequence is presented in Figure 4. In the beginning, a magnitude counter (MCount) and a delay counter (DCount) is set to zero. Then the Gaussian monocycle pulse is transmitted by the driver as soon as the system is powered on or a reset is requested (for the purpose of calibration). The driver is continuously observed to detect the reflected signal ($V_{reflected}$).

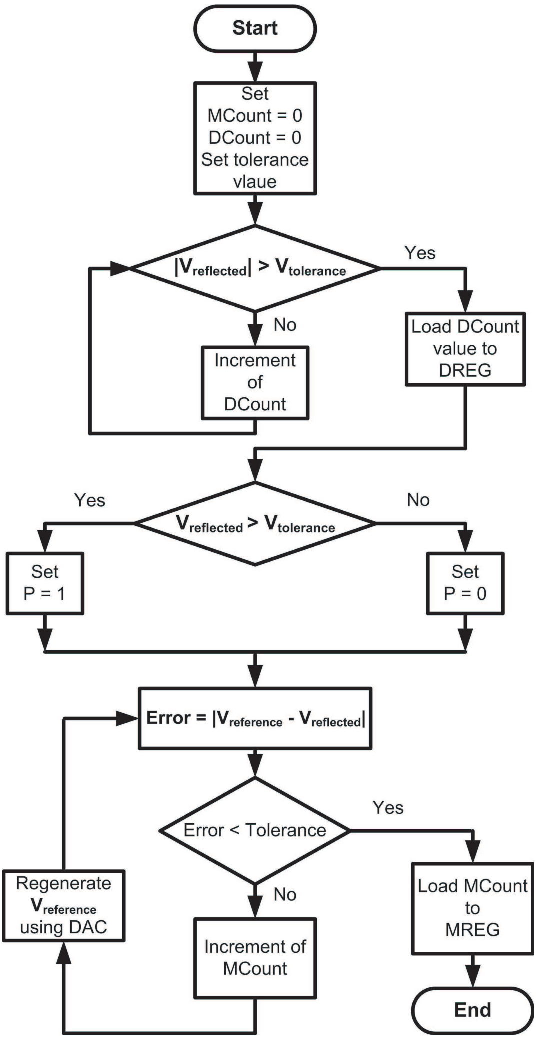

Figure 4. Calibration flow chart.

The detected signals are stored into analog storage devices, such as low leakage capacitors. The lifetime and accuracy of the stored charge is critical. It should not discharge to a point where the content is not reliable nor is it accurate. In this section details are discussed on how the delay, strength, and polarity of the reflected signal can be measured. However, before a signal is measured, a distinction must be made between noise and soft signal. A discretization level is set for all measurements according to the desired accuracy of the detection. We will refer to that as the tolerance level from this point forward. It should be noted that a detected signal is not registered as reflection if the strength of the signal is less than the tolerance, which translates the reflection into noise and is thus ignored by the system and is not compensated. With continuous sampling at the transmitter end, if a reflected signal greater than the tolerance level is detected, a timer circuit determines the time between the transmission of the pulse and the time of this detection.

The time between the transmission and reflected signal can be reliably determined by detecting the time lapse between the zero crossings of the transmitted pulse and the reflected pulse near the driving end of the circuit. This time is converted to a digital equivalent and then programmed into DREG. This register content is then used during normal operation to define the compensation signal delay period.

For accurate cancellation, the correct polarity of the reflected signal from the transmitter end must be known. The polarity of the reflected signal depends on the reflection coefficient at both the load end and the transmitter end. To determine the polarity the reflected Gaussian monocycle pulse is differentiated. A peak detector is then used to measure the peak value in the derivative of the monocycle pulse. If the largest peak is a positive value, then the reflection is positive, P is set to '1'. If the largest peak is of negative value, then the reflection is negative, P is set to '0'. To implement this routine, at first the reflection is passed through a Gaussian band pass filter, with center frequency equal to the intended carrier. This will ensure maximization of the received pulse amplitude at the intended carrier frequency, while attenuating other frequencies.

To determine MREG contents, the highest peak value that is detected in the polarity test is stored in a temporary device and input to a comparator. The other input is the analog equivalent of the MCount register value $V_{reference}$, which will be incremented starting from zero.The increment is continued until the magnitude of the MCount reference value ($V_{reference}$) is similar to $V_{reflected}$. The binary equivalent of the final MCount value is stored in MREG.

The overall calibration process will not necessarily stop with a single detection as the system may have multiple reflections which will result in the detection of several reflected signals. The amplitudes of each of this signal must be recorded along with the delay time of each reflected pulse.

As all system parameters are functions of ambient environmental condition and process, system parameter can change dynamically. And this will also change the nature of scatterings. Therefore, the calibration routine delay and magnitude associated with reflected will change as well. This requires recalibration and thus resetting the value of DREG and MREG at appropriate intervals.

5. SIMULATION SETUP AND RESULTS

To demonstrate the proposed architecture a simple AM transceiver system is used as shown in Figure 5. This system has been modeled and simulated for demonstration of the proposed concept. On the transmitter side a microstrip patch is considered for the antenna. A transmission line propagates an amplitude modulated high frequency signal from a power amplifier to the transmitting patch antenna. The S_{11} return loss of the antenna, characterizing the impedance matching between the transmitting network and the feed network of the antenna is shown in Figure 6. The magnitude of S_{11} of the antenna in decibels represents how much power is reflected back from the antenna. At 23 GHz, which is the resonant frequency of the antenna, minimum

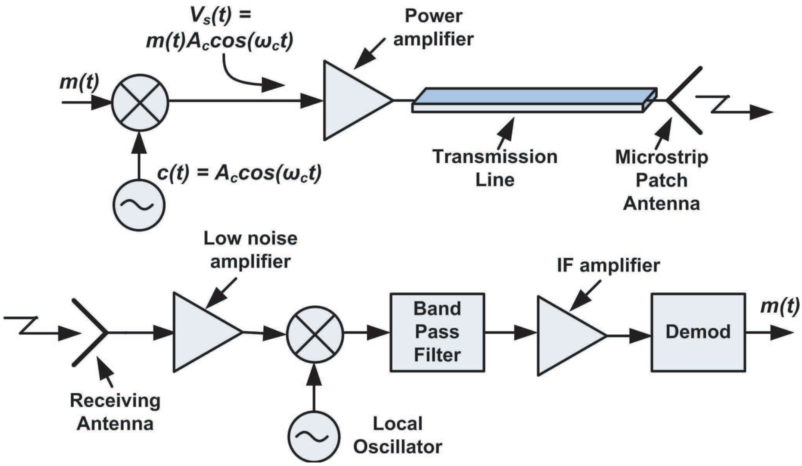

Figure 5. Simulation setup of an AM transceiver system.

amount of power will be reflected. For a carrier frequency of 20 GHz a considerable amount of transmitter signal is reflected back from the antenna. At 20 GHz, the scattering (reflections) will be sustained in the transmission line through multiple reflections and will corrupt future signals. The sustained reflections will deteriorate the VSWR and antenna efficiency. Eventually this effect will corrupt received signal at receiver end.

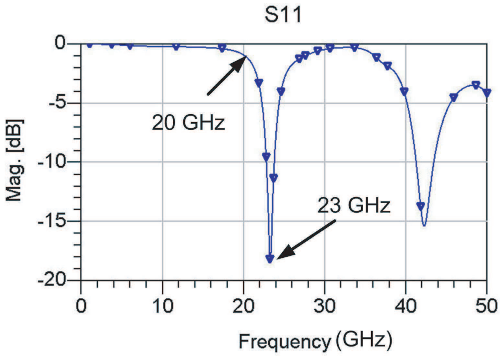

Figure 6. S_{11} parameter of transmitting microstrip patch antenna.

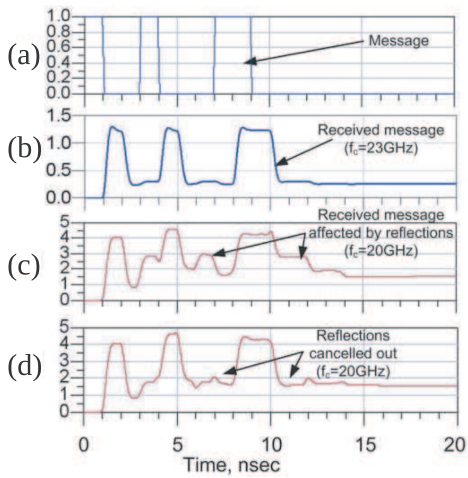

Figure 7. (a) Data is transmitted at carrier frequency of 23 GHz which is same as antenna resonant frequency. (b) Demodulated message is good. (c) Message is corrupted with reflections when transmitted at non-resonant frequency of 20 GHz. (d) Using cancellation technique the received message is 'cleaned'.

Figure 7 demonstrates the performance of the transceiver system at two different carrier frequencies 23 GHz and 20 GHz. In both the cases, the system has been calibrated and the simulated active system response is shown for both at the relevant carrier frequencies, i.e., at 23 GHz in Figures 7(a)–(b) and 20 GHz in Figures 7(c)–(d), respectively. Figure 7(a) shows the actual message that is being transmitted through the transmission line system. The first set of data shows the system operating at the resonant frequency of the antenna, which is 23 GHz. It is shown in Figure 7(b) that the integrity of the demodulated message signal is good, meaning that the transmitted message is detected well by the receiver, as compared with the actual data in Figure 7(a). Given the return loss at 23 GHz is around -18 dB, the amount of reflected signal is very negligible, therefore the transmission is good. For the second case, the return loss of the antenna at 20 GHz is rather high, at -1.5 dB, indicating a significant mismatch and thus reflections, rendering the received message unreliable, as shown in Figure 7(c). Implementing our proposed I/O driver model (as shown in Figure 2) we have compensated the reflections. In Figure 7(d), a successful demodulation shows that the received message is relatively clean, i.e., reflections have been cancelled out.

If however, reflections were present in the transmission line, as in Figure 7(c), all higher order reflections will have significant effect on signal to noise ratio (SNR) and jitter. As our proposed methodology significantly cancels all higher order reflections, the system signal to noise ratio (SNR) and deterministic jitter will also improve.

Table 1. Analysis of message eye for both an uncompensated system and a compensated system.

Parameters	Uncompensated system	Compensated system
Eye Height (V):	0.71897	2.16814
Eye Width (s):	7.40576×10^{-10}	8.691796×10^{-10}
Eye Opening Factor:	0.47898	0.84917
Eye Signal_to_Noise:	1.91933	6.62998
Eye Jitter (PP) (s):	1.02439×10^{-9}	1.374723×10^{-10}
Eye Level One (V):	3.55770	4.362426
Eye Level Mean (V):	2.57327	2.981087
Eye Amplitude (V):	1.96884	2.762679

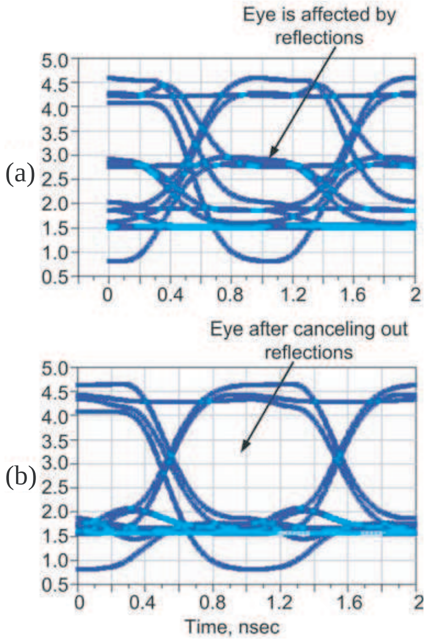

Figure 8. Eye diagram of received message pulse train (a) before and (b) after reflection cancellation.

Eye diagrams have been used previously to determine the quality of data transmission [1, 12]. In this paper we make similar use of the data eye. The effectiveness of the reflection cancellation technique is clearly evident from the analysis of the eye diagram generated using the message content of the transmitted signal. Figure 8(a) shows the degradation of the eye which is significantly compromised due to reflection. It will result in the transmission of erroneous data. However, as shown in Figure 8(b), after compensation the eye has opened up significantly, being relatively free from reflections. The improvement in the data transmission is evident in several eye characterization parameter enhancements, as shown in Table 1. The table compares the system simulation results for an uncompensated system where reflections are not cancelled out and a compensated system where reflections are cancelled out. In the table it is shown that eye height, eye width and eye opening factor are significantly increased. As the data in the Table 1 shows, the most significant improvement in the received signal can be best measured by the opening in the received eye. A larger eye opening has a significant impact on the quality of the transmission system — A lower bit error rate (BER), meaning a higher accuracy in data transmitted and thus lesser transmit repeat requests, which will increase the data transmission speed and also reduce load on traffic systems. A more coherent eye also means smaller signal strength is acceptable which can translate to more power saving, larger area of coverage, and more advanced lower voltage advanced technology usages in electronics circuits. The smaller jitter in the transmitted data not only provides for lower BER but also allows for support for higher frequencies (bands).

6. CONCLUSION

This paper has presented a technique for use in communication systems to reduce reflections in a dynamic environment without modifying the physical construct and configuration. A programmable I/O driver architecture is introduced which will be configured to cancel reflections that can be generated by any static or dynamic changes in the transmission line. To configure the system an efficient calibration routine is presented that will identify the magnitude and delay of the reflected signals. The compensation routine then adds this signal at an appropriate time which will minimize the reflections by cancellation. A setup is presented for an AM communication system model where a shift in the carrier frequency away from the antenna resonant frequency is simulated. Results show that the programmable I/O driver successfully cancels reflections. The success of the technique is measured using the eye of the received message signal. This methodology can be implemented in other communication systems as well. The advantage of the methodology is in its ability to adapt the system transmitter to the potential changes that a system can be encumbered with without damaging the data transmission reliability. The proposed method does not require any change in the overall system configuration but a modification in the system transmitter, therefore any existing system can be upgraded with a simple change.

REFERENCES

1. Zerbe, J., et al., "Equalization and clock recover for a 2.5–10 GB/s 2-PAM/4-PAM backplane transceiver cell," *IEEE Journal of Solid-state Circuits*, Vol. 38, No. 12, 2121–2130, Dec. 2003.
2. Papagiannakis, I., C. Xia, D. Klonidis, W. Rosenkranz, A. N. Birbas, and I. Tomkos, "Electronic distortion equalisation by using decision-feedback/feed-forward equaliser for transient and adiabatic chirped directly modulated lasers at 2.5 and 10 GB/s," *IET Optoelectronics*, Vol. 3, 18–29, Feb. 2009.

3. Sewter, J. and A. C. Carusone, "A 3-Tap FIR filter with cascaded distributed tap amplifiers for equalization up to 40 GB/s in 0.18-μm CMOS," *IEEE Journal of Solid-state Circuits*, Vol. 41, 1919–1929, Aug. 2006.
4. Pozar, D. M., *Microwave Engineering*, 3rd edition, John Wiley & Sons, Inc. Publication, 2005.
5. Collin, R. E., *Foundations for Microwave Engineering*, 2nd edition, *IEEE Press Series on Electromagnetic Wave Theory*, 2001.
6. Khalaj-Amirhosseini, M., "Wideband or multiband complex impedance matching using microstrip nonuniform transmission lines," *Progress In Electromagnetics Research*, PIER 66, 15–25, 2006.
7. Rao, Q., T. A. Denidni, A. R. Sebak, and R. H. Johnston, "On improving impedance matching of a CPW FED low permittivity dielectric resonator antenna," *Progress In Electromagnetics Research*, PIER 53, 21–29, 2005.
8. Khalaj-Amirhosseini, M., "Wideband differential phase shifter using microstrip nonuniform transmission lines," *Progress In Electromagnetics Research Letters*, Vol. 3, 151–160, 2008.
9. Neshat, M., A. Heidari, D. Saeedkia, and S. Safavi-Naeini, "Echo cancellation in pulsed terahertz integrated circuits," *33rd International Conference on Infrared, Millimeter and Terahertz Waves, 2008*, Sep. 1–2, 2008.
10. Staubli, P. and P. Heinzmann, "Bidirectional 2 Mbit/s transmission over a single fibre using electrical echo cancellation," *Electronics Letters*, Vol. 22, 534–535, May 1986.
11. Raju, S., S. M. Salahuddin, and I. Raza, "Reflection cancellation from high speed transmission line," *PIERS Proceedings*, 731–735, Beijing China, Mar. 23–27, 2009.
12. Sethares, W. A., R. A. Kennedy, and Z. Gu, "An approach to blind equalization of non-minimum phase systems," *International Conference on Acoustics, Speech, and Signal Processing, 1991*, Vol. 3, 1529–1532, Apr. 1991.