An Improved Simulation Method for High-Speed Data Transmission through Electrical Backplane

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Abstract—Wireline transceivers for high-speed data transmission through backplane and ethernet cables are important applications for microelectronic and nanoelectronic CMOS technologies. Although many circuit simulators provide correct system-level and transistor-level simulations, they usually fail to give correct results for many highly lossy or highly dispersive channels. This paper discusses the advanced simulator that we developed. It can give correct simulation results for those channels. It can also process those channel model files to allow commercial circuit simulators to give correct results. This simulator when used together with commercial simulators gives correct system-level and transistor-level simulation results.

Keywords—Backplane, S-Parameter, Fourier Transform, causality, convergence, least mean square (LMS), intersymbol interference (ISI), finite impulse response (FIR), decision feedback equalizer (DFE), signal to noise ratio (SNR).

I. INTRODUCTION

Recent developments in routing and switching have increased high-speed serial data transmission to multi-gigabit/second or even tens of gigabit/second [1], [2]. Such a high data rate casts serious challenges to legacy backplanes, ethernet cables, and even optical fibers. A serialization/deserialization (serdes) transceiver is designed to reduce the various impairments that have been added to the received signals by the channel and recover clock and data correctly from the received signals. Serializer, deserializer, equalizer, and clock and data recovery (CDR) are 4 very important functional blocks of a high-speed serial data transceiver. The equalizer is essentially a mixed signal block that challenges conventional circuit simulators. However, the equalizer has become one of the most critical functional blocks of a high-speed serial data transceiver. Correct simulations of this block are crucial to transceiver design.

Many conventional circuit simulators have dedicated transmission line models. However, a legacy backplane is usually much more complex than what those models can describe. A better solution that many conventional circuit simulators provide is to treat a legacy backplane as a multi-port network. The network is a black box whose properties are described with multi-port S-Parameters. The S-Parameters are experimentally measured and saved in a file, for example, a 4-port S-Parameter file with an extension of .s4p. The file however does not necessarily contain data at frequencies that are essential for correct simulation. A circuit simulator usually gets those missing data by performing interpolation and/or extrapolation using numerical methods such as polynomial line fit, rational fit or Padé approximation. Despite the advancement of those numerical methods they sometimes lead to incorrect simulation results because they do not look into device physics of the backplane. Fig. 1 (a) shows the time-domain pulse response of a B20 through channel whose model file is available at IEEE 802.3ap website [3]. The pulse response is obviously wrong because the channel is at least 20-inch long, even free-space light will take 1.7 nanoseconds to travel from one end to the other. Fig. 1 (b) gives the transistor-level transient simulation of the received signal of a pre-emphasis equalized 10Gbps transmission through a 56-inch FR4 channel. The simulation fails to converge.

We have developed a simulator that can correctly perform behavioral simulation for those channels. It can also properly process the model files to allow commercial simulators to give correct simulation results. When our simulator is used together with a commercial circuit simulator and crosscheck is performed to guarantee they give the same simulation results, correct system-level simulation and transistor-level simulation can be achieved. This paper first gives an overview of our simulation method. The reason why a commercial circuit simulator usually gets those missing data by performing interpolation and/or extrapolation using numerical methods such as polynomial line fit, rational fit or Padé approximation is discussed. The design and simulation of a 2X oversampled FIR pre-emphasis backplane transceiver is given.

II. EQUALIZATION DESIGN PRINCIPLES

Fig.2 shows a popular backplane transceiver structure. A hybrid analytical model to describe this transceiver is given in Fig.3 [7].

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A backplane transceiver is basically a wideband system. However, the legacy backplane does not provide bandwidth up to multi-gigahertz. Therefore, when a rectangular waveform with 1-baudperiod travels through the legacy backplane, its high frequency contents experience much more energy loss than its low frequency contents. As a consequence, its amplitude is attenuated and its time-domain width is expanded into its neighboring symbols. In addition, the legacy backplane also adds other impairments such as noise, crosstalk, and reflection to the signal that passes through it.

The equalization design is a critical part of a backplane transceiver. The design of equalizer of a backplane transceiver is to reduce the side effects of those impairments. From a system’s point overview, a successful transmission means that the received sequence $a(k)$ should be a delayed version of the original sequence $a(k)$. This is analytically expressed as

$$h(nT + \phi) = \begin{cases} 
0, & \text{when } n \neq n_0 \\
\alpha, & \text{when } n = n_0
\end{cases}$$

where $h$ is the time domain channel response, $n_0T + \phi$ is the delay, and $\alpha$ is the amplitude of the received sequence. When expressed in the frequency-domain, it becomes

$$\frac{1}{T} \sum_{k=-\infty}^{\infty} H(\omega - k\omega_0) = \alpha$$

In equation (2) we use a new coordinate where the time origin is at $n_0T + \phi$. The reason is that if no sample is made at the time origin, the conversion from a continuous Fourier Transform (CFT) to a discrete Fourier Transform (DFT) will be unnecessarily complicated. In practice, some systems may apply nonlinear operations such as modulus in duobinary (edge equalizer). A good method applied in those systems is to define the voltage transfer function as $V(n)$ or $I(n)$. If we substitute $a(n)$ and $b(n)$ with $V(n)$ in the definition of S-Parameters, we get the voltage at any port (port n). Therefore, the rest of this paper will concentrate on our simulator. We will focus on a 2X oversampled FIR pre-emphasis NRZ equalizer for a B20 backplane channel. Our work on conventional 1X FIR pre-emphasis, edge equalizer and DFE can be found in our recent publications.

III. DESIGN OF A 2X OVERSAMPLED FIR PRE-EMPHASIS NRZ EQUALIZER

The first step to design a 2X oversampled FIR pre-emphasis NRZ equalizer is to convert the .s4p file into voltage or current transfer function. To do this, we need to define the transceiver circuit topology. We use two 1X FIR pre-emphasis branches to form a half-rate structure. Therefore, for a 10-gigabit/second transmission our rectangular waveform is still 100-picosecond wide. The conversion from S-Parameter matrix to a transfer function is done as follows. First, the voltage and current at a given port (port n) has a relationship as:

$$V(n) = E(n) - I(n) \cdot Z_0$$

where $Z_0(n)$ is the internal impedance of the stimulus $E(n)$ at port n. The voltage $V(n)$ and current $I(n)$ can be expressed with the incident wave $a(n)$ and reflected wave $b(n)$ as follows.

$$V(n) = \sqrt{Z_0} \cdot [a(n) + b(n)]$$

$$I(n) = [a(n) - b(n)] / \sqrt{Z_0}$$

where $Z_0$ is the characteristic impedance of the backplane. Solving the equations, we can express $a(n)$ and $b(n)$ in terms of $V(n)$ or $I(n)$. If we substitute $a(n)$ and $b(n)$ with $V(n)$ in the definition of S-Parameters, we get the voltage at any port (port n); if we substitute $a(n)$ and $b(n)$ with $I(n)$, we get the current at any port. We can define the voltage transfer function as
\[ VTF_{m} = \frac{V(n)}{V(m)} \]  

(6)

For differential backplane we usually define

\[ VTF_{mjk} = \frac{[V(j) - V(k)]}{[E(m) - E(n)]} \]  

(7)

Once the voltage or current transfer function is obtained from the .s4p file, we need to interpolate, extrapolate the transfer function to low frequencies and high frequencies and filter out measurement errors. To do this, we need to look into the device physics of backplane. A piece of backplane is usually some strip lines connected with some connectors. The strip line can be described with equation (8) and (9) in transmission line theory.

\[ z = j2\pi f_0 g_1 + 2 g_2 \sqrt{j \frac{2 \pi f_0 \mu_0}{\sigma}} \]  

(8)

\[ y = 2\pi f_0 (j e^{j\beta} + e, \tan \delta) / g_1 \]  

(9)

where \( z \) and \( y \) are impedance per unit length and admittance per unit length, respectively, \( g_1 \) and \( g_2 \) are parameters decided by geometries of the transmission line and their expressions are given in [5] and [6], respectively, \( \varepsilon_0 \) and \( \mu_0 \) are permittivity and permeability in free space, respectively, \( \varepsilon_r \) is relative permittivity of the dielectric, \( \epsilon^{\text{eff}} \) is the effective dielectric constant of the substrate, \( \alpha \) is the conductivity of the strip, and \( \tan \delta \) is the loss tangent of the substrate. The propagation constant \( \gamma = \alpha + j\beta \) and the characteristic impedance \( Z_0 \) can be derived from equation (8) and (9). It can be proved from the equations and it has been wide accepted that at low frequencies skin effect dominates the attenuation factor \( \alpha \) and it is approximately proportional to the square root of frequency. At high frequencies, the loss tangent of substrate dominates and it becomes approximately proportional to frequency. By applying this observation, we can correctly interpolate and extrapolate the transfer function to low frequencies and high frequencies. We also define a corner frequency where the system noise becomes to dominate the measurement and filter out the noise accordingly. It is worth noticing that in order to get very fine time resolution; we usually need to extrapolate the transfer function to very high frequencies.

The second step is to get the time-domain pulse response of the backplane channel. In this step we should check if the transfer function is causal. We should force it to be causal if it is not. In addition, we should take into account the rectangular waveform as part of the channel. We define the frequency-domain product of the transfer function of the channel and the CFT of the rectangular waveform as “channel response”.

The third step is to define the target response of the 2X oversampled FIR pre-emphasis NRZ equalizer. As an example, the target response is defined as a raised cosine function.

\[ Fg(\Omega) = \begin{cases} 0, & \text{when otherwise} \\ 1 + \cos(\pi \Omega), & \text{when } |\Omega| \leq 1 \end{cases} \]  

(10)

where \( \Omega \) is the normalized frequency relative to data rate. Once the circuit topology is fixed, \( Fg(\Omega) \) should be multiplied with a constant to reflect maximum output power constraint. In the fourth step we define a time offset of the peak of the channel pulse response to the time origin of the target response. We need it to avoid unnecessary complicated operations to convert CFT to DFT. We also need this offset to guarantee a causal FIR equalizer. This time offset is added to the frequency-domain channel response as a positive phase shift.

In the fifth step we should sample the continuous-time channel response. This is necessary if we use time-domain methods to derive the filter coefficients. One of such time-domain method is to sample both the time-domain channel response and the target response. The filter coefficients are derived by de-convoluting the channel response and the target response. Another time-domain method is to least mean square (LMS). The first time-domain method only takes very limited number of samples of the time-domain channel response. It fails to give optimum results for very challenging channels. The LMS method may ends up with large residual errors if the first tap is the main tap and the tap with the maximum coefficient. The reason is that it is unable to remove pre-cursor ISI for very challenging channels even for pre-emphasis.

Frequency-domain methods to derive the filter coefficients have many advantages over time-domain methods. The DFT of the channel response \( (H_{\text{ch}}) \) is a folded version of its CFT \( (H_{\text{c}}) \).

\[ H_{\mu}(\omega) = \frac{1}{T} \sum_{k=-\infty}^{\infty} H_c(\omega - k\omega_0) \]  

(11)

The frequency-domain filter response is the division of the target response by the folded channel channel response.

\[ \text{Filter}(\omega) = \frac{Fg(\omega)}{H_{\mu}(\omega)} \]  

(12)

Fig. 4. Transfer function of a B20 through channel

When we perform inverse discrete Fourier Transform of the frequency-domain filter response, we get the coefficients of the filter. The coefficients are usually very long and should be truncated near the peak coefficient. For a 2X oversampled equalization system, it should meet the Famous Nyquist’s first criterion if it is decimated to 1X.
IV. RESULTS ON A 2X OVERSAMPLED FIR PRE-EMPHASIS

Fig. 4 shows the differential transfer function of a B20 through channel converted from its S-Parameter matrix. The magnitude and group delay are both shown. Fig. 5 shows the frequency-domain filter response and the desired filter response of a 2X oversampled pre-emphasis equalizer for 10Gbps data transmission over a B20 differential channel. In all curves, the blue curves are desired ones; the red curves are designed ones.

Fig. 6 shows the folded frequency-domain joint channel response of the same system. The joint channel is defined as the system from the digital input to the digital output. Fig. 7 shows the decimated (down sampled to data rate) joint channel response. It becomes to satisfy the Nyquist’s first criterion. Fig. 8 shows the channel pulse response before and after equalization. The differential input is a 1V rectangular pulse of 100ps long from a 50ohm voltage source.

V. CONCLUSION

We have introduced some advanced techniques to improve mixed signal high-speed transceiver simulation. Our simulator can give correct simulation results for many channels that some advanced commercial circuit simulators fail. It can also process those channel model files to allow commercial circuit simulators to give correct results. This simulator when used together with commercial simulators gives correct system-level and transistor-level simulation results.

REFERENCES