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Data communication in read-out systems: how fast can we go over copper wires?

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Abstract

In a digital X-ray imaging system, data has to be transmitted from the detector to the storage system. In future digital X-ray imaging systems, higher data rates will be needed. For some applications, e.g. protein crystallography at synchrotron beams, data rates in the order of gigabits per second are expected. Present trend for such systems is to move from a parallel data bus towards a high-speed serial readout. For high speed signaling over short distances (up to 10 m) the attenuation of copper cables is low enough to permit multi-gigabit per second speeds. In this article, an overview will be given of problems encountered in high speed data transmission over copper cable and techniques will be shown to overcome these problems. The bandwidth bottleneck in short distance communication is in the IC-technology and not in the channel. The cable transfer function results in inter-symbol interference (ISI). The skin-effect is the most significant cause of ISI for short length (10 m) coaxial copper cables. Fortunately, equalization can compensate for these effects. An equalizer has a transfer function that is the inverse of the channel transfer function. With the correct equalizer, a very low Bit Error Ratio (BER) can be achieved. The measured RG-58U cable ($\tau_1 = 0.12$ ns) could transmit at a bit rate of 8.3 Gbps, with a BER of 10^{-12}. Multi-gigabit speeds are possible over short length coaxial copper cables.

Keywords: Data-communication; Copper; Medipix; Jitter; Equalization

1. Introduction

In a digital X-ray imaging system, data has to be transmitted from the detector to the storage system (see Fig. 1).

For some applications, e.g. protein crystallography at synchrotron beams, data rates in the order of gigabits per second are expected. As an example, assuming a frame rate of 1000 frames per second, a pixel density of 40.000 pixels/cm^2 (Medipix2 [1]), and a counter depth of 16 bits per pixel, for a module area of 20 cm^2 we obtain a raw

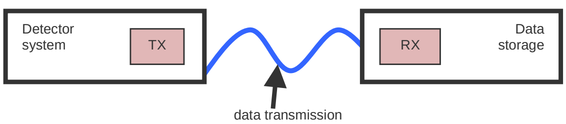

Fig. 1. An imaging system.

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data rate of 12.8 Gbps. The present trend for such systems is to move from a parallel data bus towards a high-speed serial readout (see Ref. [2]).

Glass fiber has almost ‘unlimited’ bandwidth (THz range), but often the bandwidth limiting factor in short length interconnects is in the IC-technology and not in the channel. Moreover, for driving optical cables, lasers are necessary and in the receivers we need photodiodes, so the system will be more elaborate to connect than a system using copper cables. Short length copper (length up to 10 m) can have very high bandwidth (multi-GHz range) and signal-to-noise ratio. For a given cable length, a coaxial cable has a higher bandwidth than a twisted pair cable, although the latter is cheaper. Also, characteristic impedance continuity and shielding is better for the coaxial cable. Therefore, we assume the use of coaxial cable in order to reach the highest possible transfer speed. Another advantage of copper channels is the very low thermal noise (1 nV/Hz, assuming 50 Ω termination).

In this article, an overview will be given of problems encountered in high-speed data transmission over copper cable and techniques will be shown to overcome these problems.

2. Bandwidth and spectral efficiency

The number of bits per second that a system can transmit over a given channel is a function of two parameters:

- bandwidth BW [Hz]
- spectral efficiency se(f) [bps/Hz]

$$\text{bit rate} = \int_0^{\text{BW}} \text{se}(f)\,\text{d}f \quad [\text{bps}].$$

In Fig. 2, the bit rate is shown as a function of these two parameters. It is important to see that a given bit rate can be achieved with different combinations of bandwidth and spectral efficiency.

In the graph, stars are used to indicate a number of case studies. As an example, Asymmetrical Digital Subscriber Line (ADSL) reaches a bit rate of 10 Mbps using a bandwidth in the order of 1 MHz and a spectral efficiency in the order of 10 bps/Hz. ADSL transmits data over the Plain Old Telephone System (POTS) twisted pair wires of up to 6 km long. To achieve such high spectral efficiencies, Digital Signal Processing (DSP) is exploited. This can be done because the wire bandwidth is much lower than the processing speed of the electronics. For very high bit rates this is not the case. An example from Ref. [3] reaches a bit rate of 8 Gbps (at a Bit Error Ratio of 10^{-7}) using a bandwidth of 2 GHz and a spectral efficiency of only 4 bps/Hz (4-level Pulse Amplitude Modulation). The latter system transmits data over coaxial cable of 10 m. An overview of high-speed serial communication circuits is given in Ref. [4].

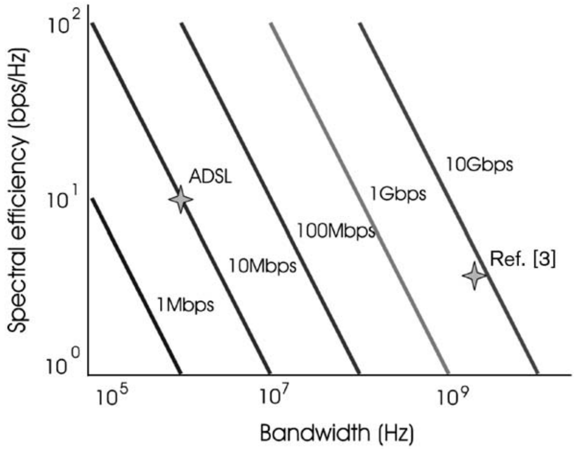

Fig. 2. Bit rate as a function of bandwidth and spectral efficiency.

Some examples of standards on the market are Gigabit Ethernet (1 Gbps over 100 m), Universal Serial Bus (USB2.0: 480 Mbps over 2 m) and FireWire (FireWire2: 800 Mbps over 2 m).

3. Cable modeling

In this section, time and frequency domain models for the coaxial copper cable will be given. This model will be fitted to Time Domain Transmissometry (TDT) measurements done on a real cable.

A coaxial copper cable can be modeled in the time domain as described in Ref. [5,6]. This model is valid for a characteristically terminated cable, at frequencies where the skin depth is low compared to the conductor diameter. First, the time constants are calculated as follows:

$$\tau_0 = l\sqrt{L_e C} \tag{1}$$

$$\tau_1 = \frac{l^2\lambda^2}{2Z^2} \tag{2}$$

$$\tau_2 = \frac{\tan\delta}{2}\tau_0 \tag{3}$$

where

$$L_e = \frac{\mu}{2\pi}\ln\left(\frac{b}{a}\right) \tag{4}$$

$$C = \frac{2\pi\varepsilon}{\ln\left(\frac{b}{a}\right)} \tag{5}$$

$$\lambda = \frac{1}{2\pi}\left(\frac{1}{a}+\frac{1}{b}\right)\sqrt{\frac{\mu}{2\sigma}} \tag{6}$$

where L_e is the external series inductance per meter (related to the field between conductors), C the shunt capacitance per meter, l the length of the cable, a the radius of the center conductor, b the distance from the center to the shield, Z the characteristic impedance, μ the magnetic permeability, ε the electric permittivity, σ the conductivity, and $\tan\delta$ the loss tangent of the dielectric. The three parts of the impulse response are:

$$h_0(t) = \delta(t-\tau_0) \tag{7}$$

$$h_1(t) = \frac{\sqrt{\tau_1}}{2t\sqrt{\pi t}}\cdot e^{-\tau_1/4t} \tag{8}$$

$$h_2(t) = \frac{1}{\pi\tau_2}\frac{1}{1+\left(\frac{t}{\tau_2}\right)^2} \tag{9}$$

where $h_0(t)$ represents the propagation delay, $h_1(t)$ represents the conductor losses and $h_2(t)$ represents the dielectric losses. The total impulse response $h(t)$ can be calculated from the convolution $h_0 * h_1 * h_2$. In Fig. 3, $h_1 * h_2$ is shown.

It is also insightful to look at the frequency domain transfer function. It can be described as

$$H(j\omega) = e^{-\gamma l} \tag{10}$$

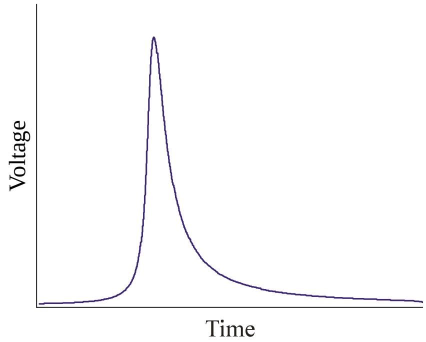

Fig. 3. Impulse response $h_1(t) * h_2(t)$.

where

$$\gamma = \sqrt{(R+j\omega L)(G+j\omega C)} \tag{11}$$

$$R = \lambda\sqrt{\omega} \tag{12}$$

$$L = L_e + L_i = L_e + \lambda/\sqrt{\omega} \tag{13}$$

$$G = (\tan\delta)\omega C \tag{14}$$

where R is the series resistance per meter, L_i is the internal series inductance per meter (related to the field inside conductors, so skin-effect dependant) and G is the shunt conductance per meter. TDT Measurements on a 10 m long cable resulted in a propagation delay value of $\tau_0 = 51\times10^{-9}$ s. The measured impulse response was fit onto the modeled impulse response using $\tau_1 = 1.2\times10^{-10}$ s and $\tau_2 = 1.0\times10^{-11}$ s. We fix the parameters $l = 10$ m, $Z = 50\,\Omega$, $\mu = 4\pi\times10^{-7}$ H/m, $\varepsilon = 2.25\cdot 8.854\times10^{-12}$ F/m, $\sigma = 5.8\times10^{-7}$ S/m and $\tan\delta = 4\times10^{-4}$. Remaining parameters a and b are tuned until the correct τ_1 and τ_2 are found resulting in $a = 0.25\times10^{-3}$ m, $b = 1.7\times10^{-3}$ m.

The skin effect causes high-frequency waves to travel closer to the surface of the conductor so that they see a higher resistance than waves of lower frequency and which makes the field inside the conductor frequency dependent. This is expressed in the formula with a frequency-dependent R and L. The skin depth is related to conductivity and frequency as follows [6]:

$$\partial_s = \sqrt{\frac{2}{\omega\mu\sigma}} \tag{15}$$

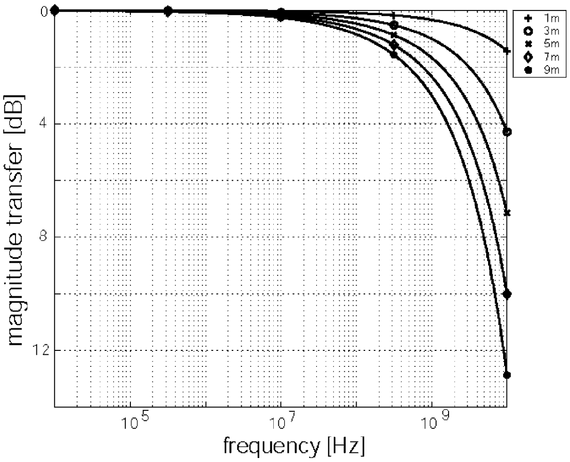

Fig. 4. Cable transfer: cable attenuation vs. frequency for 5 cable lengths.

The conductor losses and dielectric losses cause attenuation of the HF components of the signal, as shown in Fig. 4 (frequency domain magnitude transfer). In the copper cable modeled in this paper, the conductor losses are dominant over the dielectric losses ($\tau_1 \gg \tau_2$). Therefore, we will only take τ_1 into account in the rest of this paper.

4. Signal distortion in copper wires

In this section, the signal distortion caused by a coaxial copper cable will be described. When transmitting digital data over a short range of copper cable at gigabit rates, inter-symbol interference (ISI) is one of the major hurdles to be taken. It is caused by skin effect and dispersion. In the examples, it is assumed that Pulse Amplitude Modulation (PAM) is used for the signaling. For 2-level PAM, the scheme is: for a binary '1' a high voltage $+A_m$ is transmitted, for a binary '0' the inverted voltage ($-A_m$) is transmitted. Thus the signal constellation will look like $\{-A_m, A_m\}$. The distance between separate levels is $A = 2A_m$. Using 4-PAM (4 possible signal levels) will double the spectral efficiency of the system. In that case the signal constellation will look as follows: $\{-A_m, -1/3A_m, 1/3A_m, A_m\}$. For 4-level PAM, the distance between separate levels is $A = 2/3A_m$. In general, for n-level PAM,

$$A = 2A_m/(n-1). \tag{16}$$

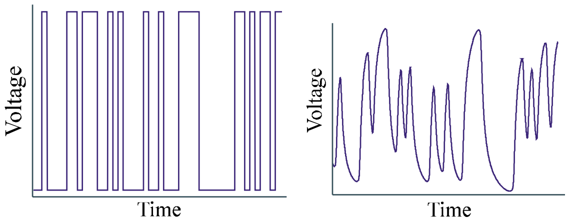

Fig. 5. Baseline wandering caused by inter-symbol interference on 2-level PAM signal (left: input signal, right: output signal).

If the symbol length T_s of a PAM signal becomes lower than τ_1, the received signal will start to show 'baseline wandering' as shown in Fig. 5. This is a result of the summation of tails of the impulse response. It can also be understood from a frequency domain point of view by noticing that the low frequencies (DC/baseline) are stronger in the received signal than the high frequencies.

ISI can be decreased by lowering the PAM signaling speed (less bps). It comes down to lowering $1/T_s$ with respect to $1/\tau_1$. This can be understood from the long-tailed impulse response of Fig. 3: if we wait longer before sending the next symbol, the tail from the previous symbol will be lower in voltage. So, ISI is dependent on the baud rate (= symbol rate).

5. Bit error ratio analysis

How can we quantify the effect that ISI has on the link quality? This can be done using the Bit Error Ratio (BER), which is the ratio between error bits and total bits sent. The BER is a function of the Signal-to-Noise ratio (SNR). Here it is assumed that ISI causes a Gaussian amplitude noise distribution. In that case, the following formula can be used for calculation of the BER [7]:

$$P_e = \frac{1}{\sqrt{2\pi}} \int_{A/\sigma_A}^{\infty} e^{(-y^2/2)}\, dy \tag{17}$$

where A is the distance between levels, σ_A is the standard deviation of the amplitude, and P_e is the error probability. For example, when $A/\sigma_A = 10$, $P_e \approx 1 \times 10^{-23}$.

The ISI contribution in σ_A is linearly dependent on A_m. In the previous section it was shown that $A = 2A_m/(n-1)$. From this it follows that $A/\sigma_A \sim 1/(n-1)$, which implies that increasing the number of PAM levels (while all other conditions remain the same) will degrade the BER. For example, when $A/\sigma_A = 3.3$, $P_e \approx 1 \times 10^{-4}$.

There will also be a timing uncertainty (jitter, or 'phase noise') in any practical circuit. This could come from the Phase Locked Loop (PLL), for example. For calculating the influence of jitter on the BER, it is assumed that the jitter also has Gaussian time distribution. Assuming a triangular eye shape, timing jitter can be translated linearly into an equivalent amplitude noise [8]:

$$\sigma_{A_{eq}} = A\left|\frac{\sigma_\phi}{\pi}\right| \tag{18}$$

$$\sigma_{tot} = \sqrt{\sigma_A^2 + \sigma_{A_{eq}}^2} \tag{19}$$

where σ_Φ is the standard deviation of the jitter in radians, σ_{Aeq} the equivalent standard deviation of amplitude, and σ_{tot} is the total (amplitude-translated) standard deviation. In Fig. 6, these functions have been plotted.

When we plot the BER as a function of the normalized symbol length, a logarithmic relation is found. For increasing T_s/τ_1, the BER decreases (see Fig. 7). For example, a BER of 10^{-12} requires $T_s/\tau_1 \approx 8$.

6. Equalization

It is possible to compensate for the frequency dependent transfer function of the channel by using an equalizer circuit. This circuit has an

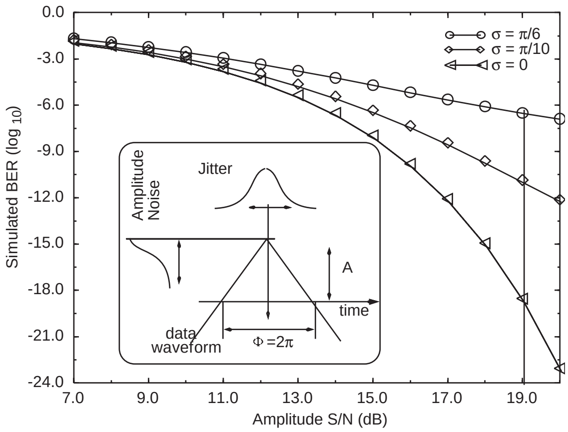

Fig. 6. Bit error ratio as a function of S/N and jitter [8].

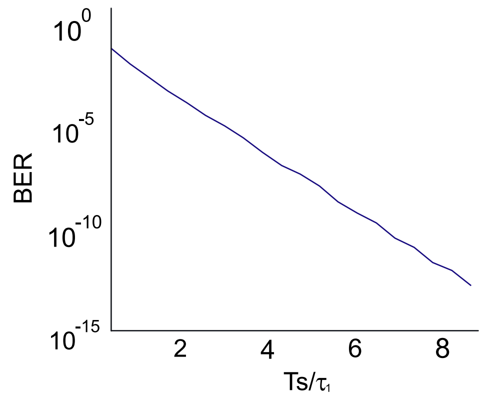

Fig. 7. BER of 2-PAM vs. normalized symbol length T_s/τ_1.

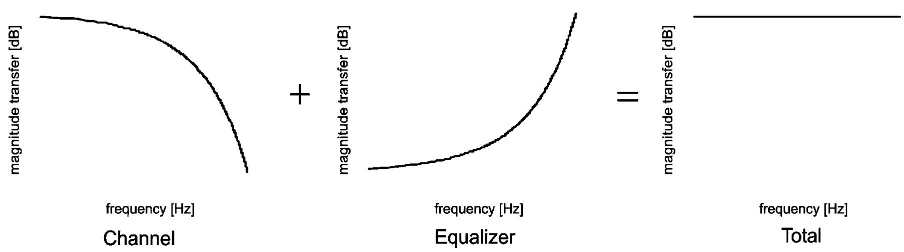

Fig. 8. Equalization.

inverse transfer function and the combination of channel + equalizer will give a flat response again. This is illustrated in Fig. 8.

In today's market, 3.2 Gbps cable equalizers are already available. A relatively simple equalizer implementation is the analog Finite Impulse Response (FIR) filter as described in Ref. [3]. The distance between samples that are weighed and summed by the FIR is chosen there as T_b (the symbol time), and 2 or 3 filter taps are used to reach a BER of 10^{-7} for 4-level PAM at 8 Gbps over 10 m coaxial cable. In Fig. 9, the equalized and un-equalized pulse shapes are shown for the above-mentioned channel and equalizer for a symbol length T_s of 250 ps.

An eye diagram can be plotted to visualize the decrease in BER thanks to equalization. An eye diagram has one (or two) bit times on the x-axis and the signal voltage on the y-axis. It can be interpreted as an oscilloscope image where the x-axis is triggered by the symbol clock; $t = 0$ corresponds to the start of a symbol. All consecutive bits are drawn on top of each other. The effect of equalization now becomes clearly visible (Fig. 10, $T_s = 250$ ps, 3-taps symbol-spaced equalizer):

In Fig. 11, the effect of equalization on the BER is shown. For $T_s/\tau_1 = 1$, a very low BER is possible, compared to Fig. 7. For that T_s/τ_1, the measured 10 m long RG-58U cable ($\tau_1 = 0.12$ ns) could transmit data at a bit rate of 8.3 Gbps.

7. Conclusions

In future digital X-ray imaging systems, higher data rates will be needed. For high speed signaling over short distances (up to 10 m) the attenuation of copper cables is low enough to permit multi-gigabit per second speeds. The bandwidth

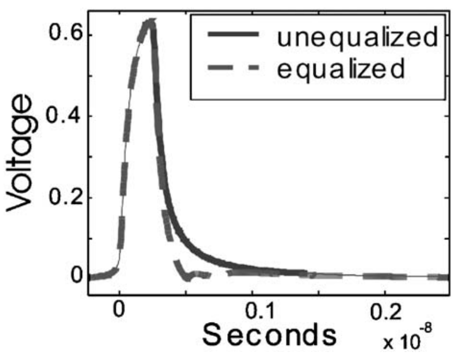

Fig. 9. Time domain response with (dashed line) and without equalization (solid line).

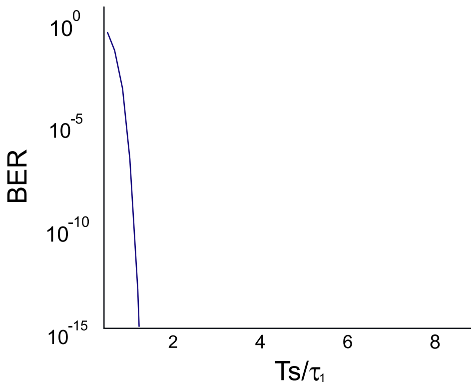

Fig. 11. BER vs normalized symbol length T_s/τ_1 for equalized channel.

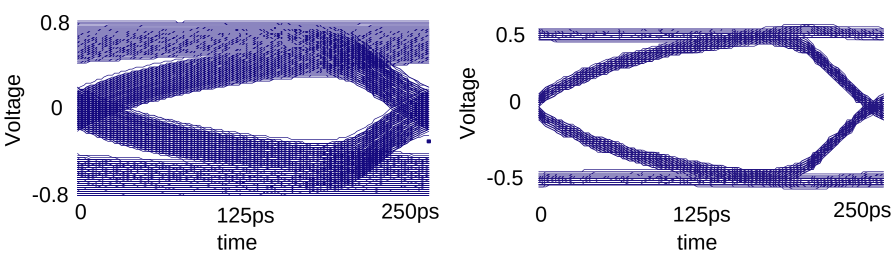

Fig. 10. Eye diagrams: unequalized, BER = 1×10^{-2} (left), equalized, BER = 1×10^{-12} (right).

bottleneck in short distance communication is in the IC-technology and not in the channel. However, the cable transfer function results in intersymbol interference (ISI). The skin-effect is the most significant cause of ISI for short length (10 m) coaxial copper cables. The amount of ISI is logarithmically dependent on the quotient of T_s/τ_1 (assuming that $\tau_1 \gg \tau_2$). Fortunately, equalization can compensate for these effects. An equalizer has a transfer function that is the inverse of the channel transfer function. With a 3-tap symbol spaced equalizer, for $T_s/\tau_1 = 1$ a BER of 10^{-12} can be achieved. The measured RG-58U cable ($\tau_1 = 0.12$ ns) could transmit at a bit rate of 8.3 Gbps. Multi-gigabit speeds are possible over short length coaxial copper cables.

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