A Comparison of Dissipated Power and Signal-to-Noise Ratios in Electrical and Optical Interconnects

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Abstract—A comparison between dissipated power and signal-to-noise ratios (SNR’s) in electrical and optical interconnects is performed. It is shown that in the absence of amplification to logic voltage levels the electrical interconnection requires much lower signal powers. However, if the amplification in the receiver is included, comparable total power dissipation and SNR’s result under the constraint of equal output voltage.

Index Terms—Interconnections, optical communications, optical interconnections, noise.

I. INTRODUCTION

There has been a significant interest in employing optics for interconnects in recent years and the subject has been treated in numerous articles and workshops [1]–[3]. At the same time, there has been an increased interest in comparing optics and electronics in the interconnect applications [4]. The alleged advantages of optics are primarily the lower crosstalk and increased packaging density achieved. In [4], an elegant discussion on the difference between the signal integrating capability of the optical receiver versus the impedance matching requirements of the electrical receiver is made. It shows a signal power advantage over the optical interconnect, due to the possibility of performing impedance conversion in the optical case. However, the paper is based on the assumption that one requests a peak voltage of 1 V. This corresponds to current logic voltage swings, but poses no fundamental limit and does in fact correspond to a huge signal-to-noise ratio SNR dB for a thermally limited transmission system, where we assume a noise bandwidth of 10 GHz. This SNR corresponds to a bit error rate BER on the order of $10^{-9}$ if Gaussian statistics is assumed.

The choice of these large voltages in digital communications is due to the fact that one often has selected the same signal voltage for communications as for logic operations. However, if one wants to limit power dissipation in order to minimize heating or power consumption, one should certainly make another choice. We also emphasize the importance of SNR as a principal parameter. An SNR = 30 dB corresponding to BER = $10^{-30}$ is considered to be sufficient in this paper.

II. SIGNAL ENERGY CALCULATION

The electrical SNR can be calculated as follows (see Fig. 1). The electrical power $P_{el}$ delivered to and dissipated at the receiver is for a ONE pulse

$$P_{el} = \frac{V_{el}^2}{Z}$$

where $V_{el}$ is the voltage which has a rectangular pulse shape. A matched load $R_1$, normally 50 $\Omega$, represents the load impedance. The electrical energy $E_{el}$ is

$$E_{el} = P_{el}T_{bit}$$

where $T_{bit}$ is the bit period.

The noise bandwidth of a receiver is usually around $B = 0.7/T_{bit}$ and is a compromise between minimizing noise and minimizing intersymbol interference. In this paper, we choose $B = 1/T_{bit}$ and hence the noise bandwidth equals the bit rate. The SNR is

$$SNR_{el} = \frac{V_{el}^2}{4kT(R_1/2)B} = \frac{E_{el}}{2kT}.$$  

Fig. 1. Electrical interconnect equivalent circuit. The characteristic impedance, the load, as well as the generator impedance is $R_1$. The circle represents the generator.
There are two noise sources which are both thermal. The first stems from the transmitter and losses in the transmission line. It is assumed that the transmitted signal has a noise temperature of $T$ and that the transmission line has the physical temperature $T$. The second stems from the matched load $R_2$ which also has the temperature $T$. In the equivalent circuit shown in Fig. 1 no bandwidth limitation is apparent. The bandwidth is set by a following filter (in practice a filtering amplifier) and a typical filter constitutes a RC-filter with time constant $\tau$, and hence $B = 1/(4\tau)$.

Fig. 2 shows the equivalent circuit of the optical detector, which is a PIN-diode which is reversed biased to make it fast and efficient. The transit time is neglected and the capacitance $C$ can be below 1 fF. We get

$$E_{\text{opt}} = P_{\text{opt}} T_{\text{shot}}$$

where $P_{\text{opt}}$ is the optical power in a ONE and $E_{\text{opt}}$ is the optical pulse energy. Further, if the quantum efficiency of the detector is unity, we write the current as

$$I = P_{\text{opt}} \frac{q}{h\nu}$$

The voltage $V_{\text{opt}}$ is

$$V_{\text{opt}} \approx IR_2 = \frac{TT_{\text{shot}}}{4C}.$$ 

The SNR is

$$\text{SNR}_{\text{opt}} \approx \frac{V_{\text{opt}}^2}{2TT_{\text{shot}}IB + 4kTR_2B}.$$ 

Here the peak voltage is approximated by $V_{\text{opt}}$. There are two noise sources. The first stems from the transmitter and the losses of the optical waveguide. It is assumed that the laser has no excess noise and emits a coherent Poissonian state and hence the noise will appear as shot noise in the receiver whether there are transmission losses or not. The second source is thermal and stems from the load resistance $R_2$.

Using (4)–(6) the SNR can be written as

$$\text{SNR}_{\text{opt}} = \frac{(E_{\text{opt}}^2)^2}{16kTC + 2qE_{\text{opt}}^2 \frac{1}{h\nu}}.$$ 

Thus, $E_{\text{opt}}$ involves the common SNR, the diode capacitance $C$ in addition to fundamental constants, temperature and photon energy. Equation (8) shows the central importance of the capacitance in an optical receiver in the sense that the optical receiver will be shot noise limited for $C \rightarrow 0$.

Fig. 3 shows the electrical and optical signal power $P_{\text{el}}$ (solid line) and optical received signal power $P_{\text{opt}}$ (dashed line is for capacitance of 10 fF and 2500 $\Omega$ and dotted line is for 1 fF and 25 k$\Omega$) versus SNR at a noise bandwidth of $B = 10$ GHz. At points A and B the voltage is 1 V.

Fig. 4 shows the electrical peak voltage $V_{\text{el}}$ (solid line) and optical peak voltage $V_{\text{opt}}$ (dashed line is for capacitance of 10 fF and 2500 $\Omega$ and dotted line is for 1 fF and 25 k$\Omega$) versus SNR at a noise bandwidth of $B = 10$ GHz. At points A and B the voltage is 1 V.

It is also interesting to calculate what voltages we are dealing with in the two cases.

It is seen from Fig. 4 that the voltages involved are indeed quite small, as an example, for $B = 10$ GHz and SNR = 30 dB, we get 2 mV in the electrical case and in the optical case for $C = 10$ fF and $R_2 = 2500$ $\Omega$ it is 14 mV. It can also be seen that the optical case with high SNR’s and low capacitances will lead to unrealistically high voltages, which is further commented on in Section III.
Fig. 5. Equivalent circuit of an amplifier stage. The voltage over the input capacitance $C_a$ is amplified via the voltage controlled current source with transconductance $g_m$. $R_a$ is the load resistance.

The origin of the marked difference between the results of this paper and those of [4] is how the two cases are compared. In [4], a voltage level of 1 V was chosen, corresponding to a power of 20 mW in the electrical case for a 50 Ω system, this is point A in Fig. 3. For the optical case, a voltage level of 1 V corresponds to a power of 1.4 mW, point B, where we assume a capacitance of 10 fF, and a bandwidth of 10 GHz and otherwise using the equations earlier in this section. Thus, [4] finds that the electrical case dissipates more signal power than the optical case. We instead compare the signal powers at the same signal to noise ratio, leading to an opposite result.

III. POWER DISSIPATION IN THE RECEIVER

The role of a complete (3R) receiver is to detect the signal (only in the optical case), amplify (1R), reshape (2R), and retiming (3R) the signal and finally deliver it at a logical level. Not all of 3R functions are always necessary, but all of them dissipate power and at least the amplification affects the SNR because the amplifier is the first stage and it is not noise free. In this paper, we restrict the analysis only to the 1R functionality only and analyze its contribution to noise and dissipated power. This is justified because the 2R and 3R functions will affect the optical and electrical case similarly. The amplification (1R) can be needed to raise the voltage level suitable for reshaping and retiming functions. The voltage level in the electrical case is lower than in the optical case and it can be raised to the optical level either by amplification or by raising the signal power, in the latter case more than necessary for a given SNR. Below we investigate the SNR and total dissipated power in the following three cases: The electrical case with and without an amplification stage and the optical case without an amplification stage under the constraint of equal output voltage $V_{out}$. For simplicity we assume that the amplifier is a single stage bipolar transistor amplifier in common emitter configuration.

Fig. 5 shows the equivalent small signal circuit. The input capacitance $C_a$ is a parasitic capacitance and its value depends much upon the technology used and can be of the order from 10 to 100 fF. The current source is voltage controlled and is described by the transconductance $g_m$ which equals $I_{bias}kT/q$, where $I_{bias}$ is the bias current. $R_a$ is the load resistance. The SNR at the output of the amplifier stage is

$$\text{SNR}_{\text{amp}} = \frac{V_{\text{out}}^2}{2kTR_a q (g_m R_a + g_m + 2/R_a)}$$

(9)

where the three terms in the denominator correspond to the amplified noise from $R_1$, shot noise of the the bias current and thermal noise of the load $R_a$ respectively. In order to minimize the degradation of the SNR in comparison with the SNR without amplification $R_a$ should be as large as possible. It can not be infinite because it limits the bandwidth due to RC-constants of the order $R_a C_a$ where we assume that $C_a$ is representative also for the input capacitance of the stage following the first amplifier stage. (The time constant $R_a C_a$ is neglected in comparison with $R_2 C_2$.) To this end we let the load resistance be equal to the load resistance in the optical detector $R_a = R_2$. We are still free to choose $g_m$. An optimal $g_m$ is found by minimizing the total dissipated power in the receiver under the constraint of a given $V_{out}$. The total dissipated power $P_{\text{diss,el}}$ using the small signal model of the transistor is

$$P_{\text{diss,el}} = P_{\text{el}} + I_{bias} V_{\text{supply}} = \frac{V_{\text{supply}}^2}{R_1} + V_{\text{supply}} \frac{kT}{q} g_m$$

(10)

where $V_{\text{supply}}$ is the supply voltage and is chosen to be 1 V in this paper.

$P_{\text{diss,el}}$ can then be expressed as

$$P_{\text{diss,el}} = \frac{V_{\text{out}}^2}{R_1 x^2} + \frac{V_{\text{supply}} kT x}{q R_2}$$

(11)

where $x = g_m R_2$ is the voltage gain. Differentiating $P_{\text{diss,el}}$ with respect to $x$ gives the optimal gain

$$x_{\text{optimal}} = \left( \frac{2V_{\text{out}}^2}{V_{\text{supply}} kT R_1} \right)^{1/3}$$

(12)

and the minimum dissipated power

$$P_{\text{diss,el, min}} = \frac{3}{2} \frac{V_{\text{supply}}^2 kT x_{\text{optimal}}}{q R_2}$$

(13)

of which one-third is signal power. Fig. 6 shows the resulting SNR as function of output voltage $V_{\text{out}}$ at a bit rate of 10 Gb/s for the electrical case with amplification. Also shown are the
SNR for the electrical and optical cases without amplification according to (3) and (7). Notice that the SNR of the electrical case with amplification and optical case without amplification are very similar and around 20 dB worse than the electrical case without amplification. The models are not valid when the output voltages approach $V_{\text{supply}}$. In the optical case the model assumes that the pin-detector is reversed biased by the supply voltage and hence the voltage $V_{\text{opt}}$ must not exceed $V_{\text{supply}}$. In the electrical case the validity of the small signal model of the transistor limits the voltage swing to much lower than the supply voltage. Of course the supply voltage can be raised but then the dissipated power also raises.

Fig. 7 shows the dissipated power in the electrical case according to (13). Also illustrated is the total optical dissipated power

$$P_{\text{diss,OPT}} = P_{\text{opt}} \left( 1 + \frac{V_{\text{supply}}}{q I_o} \right) \quad (14)$$

where the second term stems from that the supply voltage drives the photo induced current and this power, which is roughly as large as the signal power, is dissipated in the pin-diode and load. This term was not included in [4] and corresponds to the second term in (10). Also shown is the power dissipated for the electrical case without amplification according to (1) with $V_{\text{el}} = V_{\text{opt}}$. The figure shows that the power dissipation is smaller for electrical case without amplification than the amplified case if $V_{\text{opt}} < 36 \text{ mV}$ at 10 GHz and as the SNR in former case is much better, there is no need for amplification. The figure also shows that the amplified electrical case dissipates less than the optical case at output voltages larger than around 20 mV. In all cases we can choose an electrical solution with lower dissipation and the same or better SNR than the optical case. As the three cases have equal output voltage the following stages can be assumed to be identical, e.g., a further amplification stage or a decision circuit voltage.

Finally, we point out the similarity between the equivalent circuit of the optical detector and load and the electrical case with one transistor stage. In the first case there is an optical power controlled transistor current source and in the second case there is an electrical voltage controlled transistor current source. In [4], it is claimed that the photodetector works as an impedance converter and therefore works better in the sense as discussed in Section II. If so the electrical receiver with one transistor also can be called an impedance converter, which dissipates roughly the same power as the optical impedance converter. Actually the similarity between the cases is more striking than the dissimilarities. A possible physical explanation is that the current sources in both cases stem from reversed biased diodes, where carriers are injected differently. In the optical case the carriers are created in the intrinsic region whereas in the electrical case minority carriers are injected from the emitter-base diode into the reversed biased collector-base diode.

**IV. INTERCONNECT LENGTHS**

One way to design a long fiber optical link is to first calculate the minimum power needed in the receiver to obtain a sufficient SNR or BER and then to allow that the eye opening can be reduced to 0.8, which corresponds to 1 dB optical power penalty, for distorsion caused by fiber dispersion. The length of the link is then set by the dispersion limit and the transmitter power is set to compensate for the attenuation and the power penalty. In an optical interconnect the length of the fiber is short such that neither dispersion nor attenuation need to be taken into account even at high Gbit-rates (In a standard fiber the dispersion limit is 600 m at 100 Gb/s and the attenuation is 0.2 dB/km at 1.55 $\mu$m. Planar waveguides, in silica on silicon technology, the attenuation is today 0.1–0.2 dB/cm with the same dispersion as in the fiber).

In the case of electrical interconnect both attenuation and distorsion occur even at short lengths. The distorsion is now caused not by dispersion but by a frequency dependent attenuation. The attenuation coefficient $\alpha$ scales inversely to the square root of frequency $f$ for a given geometry of the waveguide, $\alpha \propto f^{-1/2}$ due to skin effect loss (dielectric loss is neglected). This leads to a length $L$ of the interconnect that is limited to $\alpha L \approx 0.2$ in agreement with [3] at $f = \frac{1}{2}$ the bit rate, for a reduced eye opening of 0.8 caused by distorsion. The transmitted signal voltage is increased by $1/0.8$ to compensate for the attenuation, and hence the transmitted signal power is increased by 2 dB.

Fig. 8 shows the length of an electrical interconnect as a function of bit rate for a coaxial geometry and with the diameter of the inner conductor as a parameter. Observe that the attenuation coefficient $\alpha$ also scales inversely with the geometrical dimension. Fig. 8 also shows another limitation. In order to ensure single mode operation the diameter must be smaller when the bit rate grows (in the coaxial case the next mode can appear when the mean circumference is one wavelength, in the microstrip case there are also higher order modes and also substrate modes).

As a conclusion the electrical interconnect length is limited in comparison with the optical interconnect length, but can still be several centimeters even at very high bit rates.
short distance interconnect in two cases:

- comparison of required signal powers for receivers without amplification to logic levels, where the requirement was equal SNR;
- comparison of total receiver power dissipation with and without amplification which is also noisy, where the requirement was equal output voltage.

In the first case, the electrical case is superior in terms of signal power, in the second case, comparable power dissipation results in optimized cases. Thus, the main conclusion from the above analysis is that electrical interconnects compare favorably with optical ones when total dissipated power in the receiver and transmitter are compared under the assumption of the same output voltage, and under the assumption that the length of the electrical interconnect is not longer than the order of cms at bit rates of 100 Gb/s and above.

In the analyses, many simplifications have been made, e.g., we have not considered the ZERO pulse case. Most of the simplifications are justified as the optical and electrical cases are compared on equal footing leading to meaningful comparisons even if the absolute numbers are not precise.

Another question is how low the logical swing can be. We have only considered shot noise and thermal noise, however the main noise or disturbance which we may meet is switching noise (noise in ground and supply) and crosstalk [7]. This electrical noise or crosstalk is normally mastered by using differential communication and low impedance levels. Differential communication is compatible with the above analysis (although we need two wires instead of one for each channel) and our proposed impedance level of 50 \( \Omega \) is low. Also, there is a strong trend today toward smaller voltage swings on interconnects, 50–400 mV [5], [6] and also toward lower supply voltages. If this noise still is a problem, we may improve the situation by using more efficient modulation or coding methods.

The claimed advantage with the optical interconnect is the impedance transforming action of the photodiode, but this can also be done in the electrical case with a transistor stage. In principle, we could of course use an ordinary transformer or another passive transformation device, to perform corresponding transformation in the electrical case. This is however considered unrealistic for the very large bandwidths and frequencies discussed here.

In summary, it appears that a general statement regarding the superiority of optical interconnects in all cases cannot be made, rather the results of this paper point to the fact that electronic interconnects indeed perform comparably in most or all respects, if transmission lengths of the order of cms are considered, even for very high bit rates, which is sufficient for interchips or intrachip connection. This paper has elucidated power dissipation and SNR in optical and electrical interconnect. A full and total comparison would have to involve entire circuits, including crosstalk aspects.

VI. DISCUSSION

This paper has basically compared optical and electrical short distance interconnect in two cases:

- comparison of required signal powers for receivers without amplification to logic levels, where the requirement was equal SNR;
- comparison of total receiver power dissipation with and without amplification which is also noisy, where the requirement was equal output voltage.

In the first case, the electrical case is superior in terms of signal power, in the second case, comparable power dissipation results in optimized cases. Thus, the main conclusion from the above analysis is that electrical interconnects compare favorably with optical ones when total dissipated power in the receiver and transmitter are compared under the assumption of the same output voltage, and under the assumption that the length of the electrical interconnect is not longer than the order of cms at bit rates of 100 Gb/s and above.

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REFERENCES


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