Energy-per-Bit Limits in Plasmonic Integrated Photodetectors

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Abstract—The energy consumption per transmitted bit is becoming a crucial figure of merit for communication channels. In this paper, we study the design tradeoffs in photodetectors, utilizing the energy per bit as a benchmark. We propose a generic model for a photodetector that takes optical and electrical properties into account. Using our formalism, we show how the parasitic capacitance of photodetectors can drastically alter the parameter values that lead to the optimal design. Finally, we apply our theory to a practical case study for an integrated plasmonic photodetector, showing that energies per bit below 100 attojoules are feasible despite metallic losses and within noise limitations without the introduction of an optical cavity or voltage amplifying receiver circuits.

Index Terms—Bit error rate, nanophotonics, optical losses, photodetectors.

I. INTRODUCTION

URING last decade, there has been a growing discrepancy between the increase in processing performance per chip and the improvement of memory access bandwidth and delay. At the same time, the power consumption of electrical interconnects has become the limiting factor in the overall system performance [1]. For longer distances, optical fibers have alleviated the electrical interconnect bottleneck and optical interconnects are now ubiquitous for wide and metropolitan area networks. Also, on shorter distances, such as for local area networks, optical interconnects are replacing copper cables while offering increased bandwidth and less power consumption [1]. For the even shorter distances, between chips and on chips, the power consumption of electrical counterparts because it is not necessary to electrically charge the entire interconnect wire (at least over the pulse length) to the switching voltage of the receiver. Instead, only the electrical capacitance of the optoelectronic components, such as the modulator and the photodetector, need to be charged up to the switching voltage [9]. For photodetectors, the received optical energy $E_p$ needed per bit is given by [3]

$$E_p = \frac{C_t V_d h\omega}{e\Gamma}$$

where $C_t$ is the total capacitance of the detector (including parasitic and receiver front-end capacitance), $V_d$ is the switching voltage swing needed on the front-end amplifier, $h$ is Planck’s constant, $\omega$ is the angular frequency of the optical beam, and $e$ is the elementary electron charge. $\Gamma \leq 1$ is the external quantum efficiency—the fraction of the incident photons that are absorbed by the device to generate electron–hole pairs—a factor that encompasses many nonidealities that increase the energy consumed per bit. For a fixed switching voltage and frequency, it is clear from (1) that the energy per bit in a photodetector is minimized by minimizing the ratio $C_t/\Gamma$. The parameters that determine $C_t$ and $\Gamma$ depend on the specific detector design bottlenecks [2]. Hence, as pointed out in [3], optical interconnects can only be a viable alternative to electrical interconnects if they consume significantly less energy per transmitted bit. Much research has been carried out in the area of low-power photodetectors that report different performance metrics such as active volume [4] or gain bandwidth [5], [6] or which introduce a new photodetector design with better performance than other designs [7]. While all those performance metrics are valid, little attention has been paid to using energy consumption per bit as a specific design goal in photodetectors. Since this is one of the two principal bottlenecks in interconnects, we will, in this paper, investigate the design tradeoffs in photodetectors that arise when we choose to optimize the energy per bit.

This paper is structured as follows: first we briefly introduce the calculation of the energy per bit in a photodetector and a generic model for a photodetector; second, the tradeoffs that arise in each optimization are illustrated both analytically and by example using the simulated properties of a plasmonic germanium photodetector. We choose germanium because it is a very promising material for photodetectors given its CMOS compatibility and its high absorption coefficient up to the band edge at 1550 nm [8]. Finally, we draw conclusions.

II. ENERGY CONSUMPTION IN PHOTODETECTORS

Optical interconnects can be more energy efficient than their electrical counterparts because it is not necessary to electrically charge the entire interconnect wire (at least over the pulse length) to the switching voltage of the receiver. Instead, only the electrical capacitance of the optoelectronic components, such as the modulator and the photodetector, need to be charged up to the switching voltage [9]. For photodetectors, the received optical energy $E_p$ needed per bit is given by [3]
and include attributes such as material choice, size, electrical contacting and/or doping structure, and optical aspects such as front surface reflections, the choice of whether to use a resonator, and mode matching into resonators or waveguides. To study the main design tradeoffs, we propose a generic detector model that allows us to consider the influence of many of these design parameters on $C_l/\Gamma$, at least in the simplest useful approximations.

III. GENERIC CAVITY-LESS PHOTODETECTOR MODEL

In this paper, we assume that all the light that reaches the photodetector is contained within a single optical mode. This framework includes waveguide-based photodetector designs [10], [11], where the light is coupled in through a single mode waveguide, as well as surface-normal designs [12], where the incoming mode can be approximated by a plane wave. In devices without a resonant cavity, detector operation is based on the absorption of light as it travels through an absorbing medium for a certain length. Given a particular lateral structure, this length determines the size of the detector. The longer this interaction distance is, the more light is absorbed in a single pass. To formalize this intuitive definition, we use a detector model as depicted in Fig. 1, where only one propagating optical mode with propagation constant $\tilde{k} = k_r - i\gamma$ is assumed in the absorbing region. This principle has often been implemented in waveguide-based devices [10], [13] and shows no optical bandwidth limitations beyond the material absorption bandwidth.

In addition to cavity-less photodetectors, a resonance could not be formed by a ring resonator [14] or integrated waveguide resonators [15], [16] for waveguide-coupled devices. Being resonant, these devices always have some limit to their optical bandwidth, and will be somewhat harder to make because of the necessity of tuning the resonance to the desired operating wavelength. As will become clear in this paper, plasmonic photodetectors can be made very short without the introduction of a cavity. Therefore, we will restrict this paper to the simplest structures, considering only cavity-less devices.

We assume a mode coupler that couples the light from the incoming single mode source to the absorbing mode without reflections and independent of wavelength. After propagation over a distance $d$ in the absorber, the remaining light of the optical mode is presumed to be perfectly coupled out without any back reflections. Though not all detectors will be so ideal, finite coupling losses could be added in later, at least as long as the detector is strongly enough absorbing that back reflections do not lead to resonances inside the detector. The fraction $A$ of the optical power that is absorbed within the detector (either in the semiconductor or the metal) is, therefore, given by

$$A = 1 - e^{-2\gamma d} < 1.\quad (2)$$

The generated photocarriers are collected by contacts placed parallel to the propagation direction of the optical mode that needs to be absorbed. The generated current is then amplified by a front-end amplifier to generate an isolated output voltage that will drive subsequent electrical stages. The contacts are assumed to have a capacitance per unit of length $C_l$ while the front-end amplifier contributes to the parasitic capacitance $C_p$ that is fixed and does not depend on the length of the detector. The total capacitance at the front end is given by $C_l = C_l d + C_p$. In fact, we assume that, when the size of the detector is increased, the total capacitance increases linearly with $d$ while the nonabsorbed power decreases exponentially with the detector length. We believe that this trend holds in most nonresonator photodetector designs.

In photodetectors, only a fraction of the absorbed power leads to the creation of electron–hole pairs in the semiconductor. Some of the power is, for example, optically absorbed by metallic contacts if the optical field of the mode reaches them (as it may well do in plasmonic designs). To account for those losses, we define the semiconductor absorption fraction $\eta_{\text{abs}}$ as the ratio of the useful power absorbed in the semiconductor to the total absorbed power. Calculating this ratio from first principles yields

$$\eta_{\text{abs}} = \frac{\int_{\text{SC}} \text{Im}(\varepsilon_{\text{SC}})|E|^2 dS}{\int_{\text{TOT}} \text{Im}(\varepsilon)|E|^2 dS}\quad (3)$$

where Im denotes the imaginary part of the relative permittivity, SC refers to the area covered by the semiconductor, and TOT refers to the entire mode area. Here, for simplicity, we neglect any absorption in the semiconductor regions that does not lead to photocurrent, such as free-carrier absorption, on the presumption that it will be weak compared to direct optical absorption in the semiconductor.

Assuming the best possible resulting external quantum efficiency ($\Gamma = \eta_{\text{abs}} A$) and utilizing (1) we obtain the following
equation for the energy per bit:

\[ E_p = \frac{(C_1 d + C_p) V_d \hbar \omega}{\eta_{\text{abs}} A} = \frac{(C_1 d + C_p) V_d \hbar \omega}{\eta_{\text{abs}} (1 - e^{-2\gamma d})}. \]  

(4)

We now calculate the length \( d_m \) that provides the minimum energy per bit, for a given \( C_1, C_p, \gamma, \) and \( V_d \hbar \omega \), by the usual technique of finding where the derivative with respect to length becomes zero, i.e.,

\[ \frac{\partial E_p}{\partial d}(d_m) = \frac{V_d \hbar \omega}{\eta_{\text{abs}} e} \frac{C_1 + 2\gamma(C_1 + C_p d_m)}{e^{\gamma d_m}} = 0. \]  

(5)

To obtain an exact value of \( d_m \), (5) needs to be solved numerically. To obtain insight, we distinguish the regimes in which the parasitic capacitance respectively dominates \( (C_1 \ll C_p \gamma) \) and does not dominate \( (C_1 > C_p \gamma) \). In the latter case, we obtain a solution for \( d_m \) that yields physical insight by making a third-order Taylor expansion of \( \exp(2\gamma d) \) in the numerator of (5) and obtain

\[ d_m \approx \sqrt{\frac{C_p}{C_1 \gamma}} = \sqrt{L_e L_\gamma} \]  

(6)

where \( L_e = C_p / C_1 \) and \( L_\gamma = \frac{1}{\gamma} \). We note that the Taylor series converges for \( C_1 L_\gamma > C_p \). From (6), we observe that \( d_m \) equals the geometrical average of the parasitic capacitance matching length \( L_e \) that corresponds to the detector length needed to have the detector internal capacitance match the parasitic capacitance \( C_p \), and the absorption length \( L_\gamma \). This implies that a shorter optimal length is obtained by reducing the absorption length.

In addition, we note that as the parasitic capacitance tends to zero, the length of the most energy-efficient detector tends to zero \( (d_m \to 0) \), but the energy per bit \( E_{p m} \) still tends to a finite result equal to

\[ E_{p m} = \lim_{C_1 \to 0} E_p = \frac{C_1 V_d \hbar \omega}{2 \eta_{\text{abs}} \gamma}. \]  

(7)

So, as the input capacitance of the front-end transistors is reduced as they are scaled down, it becomes increasingly beneficial to further reduce the size of the detector even without enhancement of its sensitivity and despite the smaller number of photons absorbed. It should therefore be stressed that, when the energy-per-bit figure of merit is applied, the design goal of absorbing nearly 100% of the incoming photons is not crucial. In addition, the minimal achievable energy consumption per bit for a given \( C_1, \gamma, \) and \( \eta_{\text{abs}} \) is given by (7), when \( C_p = 0 \).

In contrast, when the parasitic capacitance dominates \( (C_1 < C_p \gamma) \), the most energy-efficient solution is a long detector, absorbing almost 100% of the light. An accurate result for \( d_m \) can then only be obtained through an exact numerical solution of (5) and the optimal energy per bit can be approximated by

\[ E_p = \frac{C_p V_d \hbar \omega}{\eta_{\text{abs}}}. \]  

(8)

It should be stressed that in this regime \( \eta_{\text{abs}} \) is the only optical property of the photodetector that has a nonnegligible influence on the energy per bit. For a given photodetector topology one could define the equivalent parasitic capacitance \( C_{pe} \)

\[ C_{pe} = C_1 L_\gamma. \]  

(9)

When \( C_p = C_{pe} \) the contribution to the energy per bit of the parasitic capacitance \( E_1 \) [given by (10)] is equal to \( E_2(d_m) \) [given by (11)], the contribution to the energy per bit of the detector capacitance with length \( d_m \). \( E_1 \) and \( E_2 \) add up to the total energy per bit

\[ E_1 = \frac{C_p V_d \hbar \omega}{\eta_{\text{abs}} A}. \]  

(10)

\[ E_2(d) = \frac{C_1 d V_d \hbar \omega}{\eta_{\text{abs}} A}. \]  

(11)

It is, of course, important to minimize the parasitic capacitance overall, and good design here should make sure that the photodetector is integrated very close to the front-end transistors. Wiring will typically have a capacitance \( \sim 100 \text{ fF} \) per micrometer of length [3], so, when working with detector capacitances in the attofarad or low picofarad range, close integration (e.g., \(< 100 \text{ nm} - 1 \text{ μm} \)) is very desirable; such close integration is, however, consistent with transistor and feature sizes in current and future silicon electronic technologies.

IV. NOISE LIMITATIONS AND BIT-ERROR RATE (BER)

So far, we have only focused on optimizing a generic photodetector so that a minimal energy per bit can load the input capacitance \( C_1 \) to the switching voltage \( V_d \). We have done this without taking into account the unavoidable intrinsic fundamental noise components in the photodetector that will limit the achievable BER. While the purpose of this paper remains to minimize the energy per bit we will, in this paragraph, illustrate how noise places a lower limit on the minimally required energy per bit for a given BER and electrical bandwidth, when using a photodetector that has been optimized using the technique described in the previous paragraphs. As detector noise cannot be analyzed without first considering the receiver circuit, we assume the use of a typical “receiverless” receiver architecture as depicted in Fig. 1. This receiver topology has a “high impedance” and should show excellent noise characteristics when the load resistance \( R_l \) is large (this topology could also be seen as a transimpedance amplifier with a very high feedback resistance, essentially eliminating the feedback). However, the large resistance \( R_l \) comes at the expense of the electrical bandwidth \( \Delta f_{RC} \) that depends on the total detector capacitance and the load resistance [17] and is defined as

\[ \Delta f_{RC} = \frac{1}{2\pi R_l C_1}. \]  

(12)

Without avalanching effects, the fundamental relation between the BER and noise is given by

\[ Q = \frac{2RP_{\text{rec}}}{\sqrt{\sum_i \sigma_0^2 + \sqrt{\sum_i \sigma_1^2}}} \]  

(13)

\[ \text{BER} = \frac{1}{2} \text{Erfc} \left( \frac{Q}{\sqrt{2}} \right) \]  

(14)

where \( \sum_i \sigma_0^2 \) and \( \sum_i \sigma_1^2 \) represent the total mean square of the current noise contributions when a “0” and a “1” are received, respectively [18]. \( P_{\text{rec}} \) in (13) is the required averaged input power to achieve a BER set by \( Q \) and under the noise figure set
by \( \sum_i \sigma_{0i}^2 \) and \( \sum_i \sigma_{1i}^2 \), \( R \) is the responsivity of the detector, and \( Q \) is linked to the BER through (14).

In order to calculate \( \sum_i \sigma_{0i}^2 \) and \( \sum_i \sigma_{1i}^2 \), we follow the formalism set by Krishnamoorthy and Miller [19] and consider thermal noise in \( R_t \), the short channel excess noise in the field effect transistor (FET), and the shot noise contributions of the photocurrent, of the dark current and of the gate leakage current through the gate of the front-end transistor. If noise was not included in this analysis as noise contributions from low-frequency components can be avoided by using ac coupling or by using encoding techniques that do not make use of low-frequency components [19]. This yields (15) for \( \sum_i \sigma_{0i}^2 \) and (16) for \( \sum_i \sigma_{1i}^2 \):

\[
\begin{align*}
\sum_i \sigma_{0i}^2 & \approx \frac{4k_b F_n T I_2 \Delta f}{R_l} + 2eI_2 \Delta f(I_d + I_l) + 4k_b T \Gamma_n \frac{(2\pi C_l)^2}{g_m} I_3 \Delta f^3 \\
\sum_i \sigma_{1i}^2 & \approx \frac{4k_b F_n T I_2 \Delta f}{R_l} + 2eI_2 \Delta f(I_d + I_l + 2RP_{rec}) + 4k_b T \Gamma_n \frac{(2\pi C_l)^2}{g_m} I_3 \Delta f^3
\end{align*}
\]

where, \( F_n \) is the amplifier noise factor, \( k_b \) is Boltzmann’s constant, \( T \) is the absolute temperature, \( \Delta f \) is the noise bandwidth, \( I_d \) is the dark current, \( I_l \) is the leakage current through the gate, \( \Gamma_n \) is the short channel excess noise factor, \( g_m \) is the transconductance of the FET, and \( C_l \) is the total capacitance at the front end. \( I_2 \) and \( I_3 \) are Personick integrals that depend on the exact temporal shape of the signal as can be found in [19] and the references therein. Also both shot and thermal noise statistics are assumed to be Gaussian. Note that the only difference between \( \sum_i \sigma_{0i}^2 \) and \( \sum_i \sigma_{1i}^2 \) is \( 2eI_2 \Delta f(2RP_{rec}) \), the photocurrent contribution to the shot noise term.

For a given BER and bandwidth we substitute (15) and (16) in (13) and solve for \( P_{rec} \) and obtain (17), shown at the bottom of the page, where \( E_{pn} \) is the noise limited minimum energy per bit.

\[
P_{rec} = E_{pn} \Delta f \approx \frac{Q}{R} \left( Qe \Delta f + \frac{4k_b F_n T I_2 \Delta f}{R_l} + 2eI_2 \Delta f(I_d + I_l) + 4k_b T \Gamma_n \frac{(2\pi C_l)^2}{g_m} I_3 \Delta f^3 \right)^{1/2}
\]

\[
E_{pn} \approx \frac{Qh\omega}{e\eta_{abs}(1 - e^{-2\gamma d})} \left( Qe + 8\pi k_b T F_n C_l + \frac{2eI_2(I_d + I_l)}{\Delta f} + 4k_b T \Gamma_n \frac{(2\pi C_l)^2}{g_m} I_3 \Delta f \right)^{1/2}
\]
per unit length $C_l$ between the contacts. The refractive indices at 1300 nm for gold $n_{\text{gold}} = 0.4080 - 8.3028 \ i$, germanium, $n_{\text{Ge}} = 4.3500 - 0.1i$, and oxide $n_{\text{ox}} = 1.44$ were taken from [8]. For the calculation of $C_l$, the d.c. relative permittivities $\varepsilon_{\text{Ge,DC}} = 16$ for germanium, and $\varepsilon_{\text{ox,DC}} = 2.07$ for silicon dioxide (the oxide in Fig. 2) were used. For gold’s d.c. electrical conductivity (but not the optical or plasmonic propagation) we assume a highly conducting metal. Both simulations are performed using the finite-element-method-based tool COMSOL Multiphysics.

First we show the influence of the detector length $d$ and the parasitic capacitance $C_p$ on the energy per bit for a fixed photodetector topology with $w_g = t_w = 50 \ \text{nm}$. In Fig. 4(a)–(c), the energy per bit is plotted for different parasitic capacitances as a function of the detector length. The optimal detector length $d_m$ given by (6) is denoted by a triangle in each plot. The contributions to the energy per bit of the parasitic capacitance $E_1$ and the detector capacitance $E_2$ are plotted separately. Their sum adds up to the total energy per bit. Also, the minimum achievable energy per bit $E_{\text{pm}}$ is indicated by a cross (X). For calculating these energies, we presume that the switching voltage swing $V_d$ needed on the front-end amplifier is 1 V. For a different required $V_d$, the energy will scale proportionally. Finally, the noise-limited minimum energy per bit $E_{\text{pm}}$ is plotted assuming $Q = 8$, which sets the BER to $10^{-15}$. The amplifier noise factor was set to $F_n = 3 \ \text{dB}$, the temperature was set to $T = 300 \ \text{K}$, and the bandwidth was chosen as $\Delta f = 50 \ \text{GHz}$. The dark current $I_d$ was approximated as the dark current in a reverse biased metal semiconductor metal junction. For $I_d$, we used the regular expression for the current density across a Schottky barrier of height $\phi$

$$I_d = t_w d_m R_d T^2 \exp(-c_\phi/(k_b T)) \quad (19)$$

where $R_d = 120 \ \text{A/cm}^2\text{K}^2$ and where we took $\phi = 0.25 \ \text{eV}$ [23]. Realistic dark currents for $t_w = w_m = 50 \ \text{nm}$ are about 25 nA. For $I_d = j_i A_g$, we used a current density $j_i = 0.13^3 \text{A/m}^2$ as was found in the international technology roadmap for semiconductors (ITRS) roadmap for “low standby power logic” [24] and a corresponding gate surface $A_g = 22 \ \text{nm} \times 200 \ \text{nm}$. Low standby power transistors tend to have thicker gate oxides to reduce leakage that reduce the gate capacitance and the transconductance but increases the threshold voltage [24]. This increase is acceptable because of the chosen 1-V swing on the detector. For the Personick integrals, we chose $I_2 = 0.6$ and $I_1 = 0.09$ as is chosen in [19]. Finally, for the excess short channel noise, we estimated the transconductance $g_{m0} = 140 \ \mu\text{S}$ by using the linear model [25] and by using the maximum drain source current found in the ITRS roadmap for the 22-nm technology [24]. Also, we used an excess noise factor $\Gamma_n = 3$, which seems to be in the range of what has been presented in literature until gate lengths of 60 nm [26]. In our example, we found that the noise figure of our detectors was mainly dominated by the finite responsivity $R$ for small detectors lengths $d$ and by thermal noise that is about five times greater than other contributions in this example.

When $C_p$ is 10 aF, which is extremely small, we can see from Fig. 4(a) that the contribution of the parasitic capacitance is almost negligible. The ideal detector length $d_m$ is 90 nm, which is very close to zero, and the optimal energy per bit is roughly $E_{\text{pm}}$ as given by (7). Here, we are clearly in the regime where $(C_l > C_p \gamma)$. Also we have $E_{\text{pm}} = 70 \ \text{aJ}$, while $E_p(d_m) = 75 \ \text{aJ}$, so we remain above the noise limit as calculated under our assumptions. As $C_p$ is increased to 100 aF, we observe from Fig. 4(b) that the parasitic contribution roughly equals the detector contribution at $d = d_m$ and that we are well above the noise limit. The equivalent parasitic capacitance is $C_{pe} \approx 25 \ \text{aF}$ for this detector topology. In addition, (6) provides an accurate approximation for $d_m$. As $C_p$ is increased even further to 1000 aF in Fig. 4(c), the parasitic capacitance entirely dominates $(C_l \ll C_p \gamma)$ and the energy per bit is given by (8).

Also the detector capacitance $C_l$ and the absorption length $L_s$ then have a negligible influence on the energetic performance of the photodetector. The fact that we predict quite low energies even with such a relatively large $V_d$ is important in itself, and we will discuss this point more later.

We now calculate the $C_l$, $L_s$, $n_{\text{eff}}$, $\eta_{\text{abs}}$, $E_{\text{pm}}$, and $C_{pe}$ of the integrated plasmonic photodetector as we vary the width $w_g$ and the thickness $t_w$.

From Fig. 5 we observe that the absorption length of the plasmonic photodetector can be extremely short. Regardless of the thickness, $L_s$ becomes as short as 50 nm when $w_g = 10 \ \text{nm}$. For larger $w_g$, $L_s$ remains smaller than 200 nm. As can be seen in Fig. 6, $\eta_{\text{abs}}$ remains above 40% even for very small $t_w$ and $w_g$, because of the strong absorption coefficient of germanium.

In Fig. 7, the capacitance per unit of length $C_l$ is depicted. As would be expected from the a parallel plate approximation, $C_l$ is approximately proportional to $t_w/w_g$. A very low $C_l$ can still be reached for small $w_g$, as long as the thickness $t_w$ is reduced proportionally. The ability to reduce the thickness of the contacts to minute dimensions is a property of plasmonic photodetectors

Fig. 4. Plots of the energy per bit and the noise limit $E_{\text{pm}}$ as a function of the device length $d$. In each figure, the energy per bit is plotted for a specific parasitic capacitance (a) $C_p = 10 \ \text{aF}$, (b) 100 aF, and (c) 1000 aF and for $w_g = 50 \ \text{nm}$ and $t_w = 50 \ \text{nm}$. Also, the black triangles denote the energy per bit for $d = d_m$ given by (6).

...
that cannot be replicated by a nonplasmonic equivalent; a nonplasmonic device at such small scales, would no longer be able to confine an optical mode substantially within the structure.

As shown in Fig. 8, a minimum achievable energy per bit $E_{\text{pm}}$ between 10 and 100 aJ is reachable with this topology with $V_{\text{th}} = 1$ V, despite the metallic losses. The lowest energy per bit is obtained when a very thin metal layer is used. However, to achieve this low level of energy per bit, the parasitic capacitance has to be reduced as well, below the equivalent parasitic capacitance $C_{\text{pe}}$ depicted in Fig. 9, which lies between 20 and 100 aF. Currently, the best conventional front-end amplifiers still have an input capacitance well above this limit, for example, around $C_p = 30$ fF [27]. Lowering the parasitic capacitances of the front end to 10s to 100s of attofarads remains a necessary condition to make full use of the low-energy potential of a plasmonic photodetector. Otherwise, plasmonic photodetectors will likely remain less energy efficient than their nonplasmonic counterparts as the lower detector capacitance can no longer compensate for the metallic losses. The capacitance of the front-end amplifier can drop with downscaling of the transistors and by using more energy-efficient amplification schemes not based on the transimpedance amplifier [28], but the improvements needed are very substantial.

The basic design of such low-capacitance receiver circuits has been considered in principle before [19], [29], [30]. Traditional receiver circuits for telecommunications are often designed for the minimum received optical energy regardless of electrical power dissipation in the receiver circuit—not minimum total energy per bit—and are therefore designed to be limited by noise considerations. For conventional photodetectors with large capacitances, such noise optimization generally leads to wide input transistors with large $g_m$ to minimize excess Johnson noise from the transistor channel [last term in (18)], which in turn leads to large receiver dissipation from the standing current in the transistor. Designing for minimum total energy per bit instead leads to much simpler receiver circuits; one optimum is to have a single CMOS transimpedance stage with near minimum-size transistors [19]. As previously mentioned, our very simple design that may be close to optimal in total energy
is to eliminate voltage amplification altogether in the “receiverless” design; in such a design, the optically induced voltage change on the photodetector corresponds directly to the logic level swings needed in electronic logic circuits [19], [29], [30]. Our choice of $V_d = 1$ V here is sufficiently large to run in this receiverless mode, in which case there is no electrical dissipation in voltage amplifiers and allowing for the use of high-threshold transistors with a thick gate oxide. Despite this large voltage swing, very low energy per bit within the noise limit is possible, as illustrated in our calculations, because of the use of very small, low capacitance detectors. With such approaches and with the continuing reduction in size and capacitance of future transistors, the necessary low parasitic capacitances that our detector approach here demands from the receiver circuit front end may be achievable.

VI. ENERGY PER BIT AND ELECTRICAL BANDWIDTH

The electrical bandwidth of a photodetector is limited both by the time it takes for the minority carriers to reach the contacts, referred to as the transit-time bandwidth [25], and by the RC constant of the electrical front end, referred to as the RC bandwidth. As the RC-bandwidth may be set by the electrical amplification scheme and by noise requirements, the transit-time bandwidth still is an upper limit on the achievable bandwidth, regardless of the amplification scheme used, and is found to be the limiting factor in many practical detectors [31]. The transit-time bandwidth $f_t$ is proportional to the saturated drift velocity $v_c$ in the semiconductor and inversely proportional to the width of the depleted semiconductor region (i.e., the distance between the contacts) [25]. Hence, $f_t$ is given by

$$f_t = \frac{0.44 \, v_c}{w_g}.$$  \hspace{1cm} (20)

Using (20) and $v_{c,Ge} = 6.0 \times 10^4$ m/s [31], the transit-time electrical bandwidth of the plasmonic (and nonplasmonic) detector is plotted in Fig. 10 as a function of the waveguide width $w_g$. From Fig. 10, we can deduce that for an electrical bandwidth higher than 100 GHz a waveguide width smaller than 200 nm is required, and this is at the limit of what can be achieved with regular dielectric waveguides. Extremely high bandwidths of 1 THz are achievable for $w_g = 30$ nm that can be reached through plasmonics. (Though such bandwidths exceed those of current electronic amplifiers, such fast detectors could be used directly in applications such as photoconductive switches, e.g., for sampling for analog to digital conversion [32]).

From Fig. 8, we observe that for $w_g > 20$ nm the minimal achievable energy per bit increases as $w_g$ is reduced. This can be attributed to an increasing $C_l$ as $w_g$ is reduced. In that regime, the electrical bandwidth comes at an energy cost per bit. However, for very narrow gaps $w_g < 20$ nm a smaller width $w_g$ yields a slight reduction in minimal achievable energy per bit $E_{pm}$, despite the increase in $C_l$ and lower $\eta_{abs}$. This is because the increase in $C_l$ is compensated by the reduction of the absorption length $L_{\gamma}$, yielding more electrical bandwidth at a lower energy cost per bit, in spite of the increased metallic losses. This benefit may be partly due to an increase in the effective index of the mode at these small size scales, which in turn will lead to greater absorption in a given length of Ge. To show that, in our topology, the energy per bit is mainly determined by $C_l/\gamma$ the minimal energy per bit with metal losses neglected $E_{pm}\eta_{abs}$ is plotted in Fig. 11. One can see that Fig. 11 is essentially only a scaled version of Fig. 8.

The fact that it is possible to achieve more bandwidth at lower cost per bit in our topology is merely a consequence of the fact the $\gamma$ increases faster than $C_l$ as $w_g$ is lowered below 20 nm. In addition, we need to underline that the minimal energy per bit assumes a detector capacitance that is linearly dependent on the detector length and assumes no parasitic capacitance. As $\gamma$ gets extremely large this assumption will break down as 3-D end effects will start to dominate the detector capacitance. These effects could be modeled as an intrinsic parasitic capacitance but the minimal energy per bit would rather be given by (8) than by (7). For our topology, solutions of the 3-D Poisson equation for a detector of finite length showed us that the detector capacitance is well approximated by the $C_l d$ until $d = l_w$ as can be seen in Fig. 12.

The relation between the energy per bit and the electrical bandwidth is illustrated in Fig. 13, where the minimal energy per bit $E_{pm}$ of the analyzed plasmonic germanium photodetector is plotted for each electrical bandwidth $f_t$ as calculated from (20). At lower bandwidths and large widths $w_g$, there is a tradeoff...
between speed and energy per bit. At higher bandwidths and narrow widths this tradeoff no longer exists as narrower devices yield a lower energy per bit due to the dramatic reduction in absorption length. For the sake of comparison, we performed the same analysis for a nonplasmonic integrated germanium photodetector described in [31], as it perfectly matches our formalism and uses germanium as a photoabsorber with high quantum efficiency. A minimal achievable energy per bit $E_{\text{pm}} = 700 \text{ aJ}$ was obtained, which is approximately two orders of magnitude higher than what could potentially be achieved with the plasmonic topology, despite the losses.

VII. CONCLUSION

When the energy-per-bit metric is applied to a cavity-less waveguide photodetector, where the total capacitance consists of the sum of a capacitance that varies linearly with the detector’s length and a fixed parasitic capacitance, we draw the following conclusions. The value of the parasitic capacitance distinguishes two regimes. First, when the parasitic capacitance dominates, the energy per bit is proportional to $C_p/\eta_{\text{abs}}$. This means that the semiconductor absorption fraction $\eta_{\text{abs}}$ is the only optical property of the photodetector that has an influence on the energy per bit. In this regime, plasmonic photodetectors will remain less energy efficient than regular larger dielectric solutions because of their metallic losses. For a given topology the equivalent parasitic capacitance limits this regime. This limit lies between 10 and 100 aF for the plasmonic detector we analyzed here. As this is much lower than the input capacitance of today’s front-end amplifiers, current technology is still well within this parasitic-dominated regime. In addition, to keep $C_p$ at these levels, the front-end transistor would presumably have to be placed very near the photodetector as the parasitic capacitance of their interconnection could be large. Second, when the parasitic capacitance drops below the equivalent parasitic capacitance, the ideal detector length that optimizes the energy per bit is equal to the geometric average of the absorption length and parasitic capacitance matching length. This length tends to zero as the parasitic capacitance is made zero while the optimal energy per bit converges to a finite minimal achievable energy per bit for a given photodetector topology. This minimum is proportional to $C_i/((\eta_{\text{abs}}\gamma))$. In plasmonic detectors, the lower absorption fraction $\eta_{\text{abs}}$ can, therefore, still be compensated by a lower $C_i$ and an increased absorption constant $\gamma$ as long as the parasitic capacitance is negligible. We calculated that a minimal energy per bit $E_{\text{pm}}$ between 10 and 100 aJ in germanium-based plasmonic photodetectors, which is much lower than the $E_{\text{pm}}$ we estimated for the dielectric equivalent found in [31]. We also point out that the noise limited energy per bit $E_{\text{pm}}$ for such photodetectors is on the order of $70 \text{ aJ/bit}$ at 50 GHz for a BER of $10^{-15}$. Furthermore, we emphasize that the goal of this paper is minimize the energy per bit without optimizing for bandwidth nor for BER. Detectors whose minimal energy per bit are below the noise limit may still function at that energy per bit but with a higher BER. The found energies per bit presume, conservatively, that we are inducing 1 V of swing in the photodetector with the absorbed light. Such low energies per bit would, however, require very small receiver circuits with correspondingly low input capacitance, though such receivers appear to be possible in principle using CMOS amplifiers or by operating “receiverless,” i.e., without any receiver voltage amplification, as is possible with our presumed 1-V swing. The lowest $E_{\text{pm}}$ are achieved when very thin metals are used. We emphasize that, while substantial efforts have been delivered to reduce the waveguide gap, significant gains in energy per bit can also be achieved by reducing the metal thickness.

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REFERENCES


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