Optical Receiver with Photodiode Gain Stabilization

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Abstract: Avalanche photodiodes (APD) used in receivers for Free-Space Optical (FSO) communication operate in a relatively narrow range of avalanche multiplication. The optimal reverse voltage depends on temperature and is subject to manufacturing variance. These effects are usually compensated for in the open-loop mode, where a DC source is programmed with average datasheet temperature coefficients and individually adjusted to compensate for manufacturing variance. The paper describes the design of a receiver with closed-loop stabilization of APD parameters.

Key-Words: Optical Receiver, Free-Space Optics, Avalanche Photodiodes

1 Introduction

The input circuit of an optical receiver consists of an avalanche photodiode and a low-noise transimpedance amplifier (TIA) followed by a filter and decision circuit ( comparator), Fig. 1. Typical reverse operating voltages of avalanche photodiodes range from tens to hundreds of volts.

![Typical receiver configuration.](image)

The photodiode current is [1]

\[ i_D(t) = I_{D1} + M(I_{D2} + R p(t)), \]

where \( M \) is the multiplication coefficient (also referred to as the diode gain), \( I_{D1} \) is the non-multiplied dark current and \( I_{D2} \) is the multiplied one. The diode responsivity \( R \) is wavelength dependent. The symbol \( p(t) \) denotes the instant value of the input optical power.

Fig. 2 shows an example of APD total responsivity \( R_T = M R \) as a function of reverse voltage and temperature for the Elmer-Perkin small-area silicon photodiode C30902. The diode is suitable for data rates of up to 1Gb/s. The APD responsivity also exhibits a manufacturing variance whose typical value is in units of volts.

A stable gain of the APD can be provided either by thermal stabilization of the diode unit or by adjusting the reverse voltage. FSO outdoor units are subject to considerable changes of the ambient temperature (typically from -30°C to +60°C), which complicates thermal stabilization.

![Total responsivity of photodiode C30902.](image)
example, according to the datasheet of C30902 the linear temperature coefficient of reverse bias voltage lies between 0.5 V/°C and 0.9 V/°C, the typical value being 0.7 V/°C.

2 Input Signals

Fig. 3a) shows a typical configuration of terrestrial FSO link. The majority of links use simple on-off keying (OOK) for data transmission, i.e. the symbol “1” is represented by power level $P_{1, \text{RXA}}$ at the receiving aperture and the symbol “0” is represented by power level $P_{0, \text{RXA}} < P_{1, \text{RXA}}$.

There are two main mechanisms influencing the optical beam in the atmosphere – the scattering on particles and the turbulence. The attenuation caused by scattering increases significantly during fog, rain, and snowfall, and may cause an interruption of communication. It is a very slow process, which determines the link overall availability [3]. The atmospheric turbulence affects the optical beam in the millisecond time scale. It increases the bit-error rate and interferes with communication protocols.

| Fig. 3 | a) Propagation of optical beam through atmosphere, b) received optical power influenced by turbulence.

Let us consider the FSO link in Fig. 3a), where the input optical power at receiver aperture RXA is influenced by the atmospheric turbulence. The turbulence just redistributes energy in the beam, i.e. there is no energy loss. If the receiving aperture is smaller than the beam cross-section the received power fluctuates, Fig. 3b). The timescales of data signal and received power are usually characterized by the inequality

$$\tau_{\text{DATA}} \ll \tau_{\text{TURB}}.$$  \hfill (2)

Typical values are $\tau_{\text{DATA}} < 1\mu s$, $\tau_{\text{TURB}} \approx 1\text{ms} - 10\text{ms}$.

Considering the equal probability of the occurrence of symbol “0” and symbol “1” and with respect to (2) we can define the short-time mean power at the receiving aperture as

$$P_{m, \text{RXA}} = \frac{1}{2} P_{1, \text{RXA}}.$$  \hfill (3)

The short-time mean optical power $P_{m, \text{RXA}}$ fluctuates randomly due to atmospheric turbulence. The normalized received power is

$$P_N = \frac{P_{m, \text{RXA}}}{\langle P_{m, \text{RXA}} \rangle}$$  \hfill (4)

where the mean value $\langle P_{m, \text{RXA}} \rangle$ is computed over the time scale of turbulences. The power scintillation index (PSI)

$$\sigma_p^2 = \frac{\langle P_{m, \text{RXA}}^2 \rangle - \langle P_{m, \text{RXA}} \rangle^2}{\langle P_{m, \text{RXA}} \rangle^2} = \langle P_N^2 \rangle - 1$$  \hfill (5)

provides a measure of the scintillation strength, i.e. a measure of how the received power “oscillates” around the mean value. The scintillation index is inversely proportional to the diameter $D$ of receiving aperture due to the effect of averaging [4].

For short-range terrestrial links ($D = 200\text{mm}$, $L_{12} = 1\text{km}$, weak fluctuations) the typical value of PSI is $\sigma_p \leq 0.5$.

Fig. 4 shows a typical power spectrum of short-time mean power. The “bandwidth” is determined mostly by the perpendicular component of wind speed.

| Fig. 4 | Typical power spectral density of power fluctuations at receiver (from [5]).

Fig. 5 shows the situation at the decision circuit.
in the case of DC coupling. All signal and noise levels are referred to the TIA input current \( i(t) \). Current \( S_1 \) corresponds to the reception of symbol “1” (light source is on). Noise current \( N_1 \) (for “1”) is generally greater than noise \( N_0 \) (for “0”) since the APD noise is signal-dependent [2].

\[
\begin{align*}
\text{Fig. 5} & \quad \text{Signals at decision circuit.}
\end{align*}
\]

On the assumption of the Gaussian noise and the equal probability of the occurrence of symbols “0” and “1”, the probability of a single-bit error can be expressed as

\[
P_e = \frac{1}{2} \Phi \left( -\frac{c_{th}}{\sqrt{\langle N_0^2 \rangle}} \right) + \frac{1}{2} \Phi \left( \frac{c_{th} - S_1}{\sqrt{\langle N_1^2 \rangle}} \right), \tag{6}
\]

where \( \Phi \) is the normalized Gaussian probability distribution function and \( c_{th} \) is the threshold.

The optimum value of \( c_{th} \) minimizing the probability of a bit error can be found using rather complex formulae [6]. Nevertheless, the most common solution is to use approximation (7), i.e. a simple AC coupling between the receiver front-end and the decision circuit

\[
c_{th,AC} = \frac{S_1}{2}. \tag{7}
\]

Thus, the filter in Fig. 1 is a combination of Low-Pass Filter (LPF) and High-Pass Filter (HPF). The cut-off frequency of HPF should be set as a trade-off between filtering-out the power fluctuations and the requirements of the data signal.

The effect of unfiltered fluctuations can be modeled by deviation \( \Delta c_{th} \) of the threshold from its “optimum” position. For simplicity, equal noise contributions \( N_0 \) and \( N_1 \) were used in Fig. 6. The symbol \( P_{e\theta} \) denotes the probability of a single bit error for threshold \( c_{th} \) calculated using (7).

Taking into account the typical value of PSI \( (\sigma_p \leq 0.5) \) and the curves in Fig. 6, we need approximately 40dB (electrical) attenuation of fluctuations. For the 1\(^{st}\) -order HPF this gives

\[
f_{HPF} / f_{urb} > 100, \tag{8}
\]

where \( f_{HPF} \) is the cut-off frequency and \( f_{urb} \) is the “bandwidth” of fluctuations, see Fig. 4. For example, for a 1Gb/s receiver with MAX3793 the cut-off frequency of HPF implemented directly in TIA is 70kHz.

\[
\begin{align*}
\text{Fig. 6} & \quad \text{Increase of bit-error rate caused by deviation of } c_{TH} \text{ from the value of (7).}
\end{align*}
\]

3 APD Gain Stabilization

The proposed APD stabilization method consists in injecting a harmonic pilot signal to the optical input, Fig. 7. The pilot signal is generated with a current-driven LED suitably coupled with APD and then extracted at the TIA output with a narrow band-pass filter. The reverse bias voltage of APD is adjusted by the microcontroller to maintain a constant level of the extracted pilot signal. This simply compensates for temperature and manufacturing variance of APD parameters.

\[
\begin{align*}
\text{Fig. 7} & \quad \text{Receiver with APD stabilization loop.}
\end{align*}
\]

The main source of temperature dependence is the LED. Its efficiency decreases with temperature. Fig. 8 shows the typical dependence for an 850nm infra LED. For the temperature range of interest (-30°C to +60°C) the expected variation is ±0.5dB
(optical), i.e. ±12% of the APD gain.

Fig. 8 Relative radiant output of infrared LED.

The stabilization loop should not impair the receiver sensitivity. The amplitude of the pilot signal should be as low as possible.

Let $P_{m,TH}$ be the short-time mean power of signal corresponding to a bit-error rate of $10^{-6}$. For the Elmer Perkin APD C30902, the threshold is $P_{m,TH} = -38\text{dBm}$ [2]. With respect to (8), the level of pilot signal at the APD input should be comparable to $P_{m,TH}$ at the frequency of the pilot signal of 1kHz. The attenuation of the LED output power is attained by a suitable mechanical design, Fig. 7.

The pilot signal is extracted from the TIA output by means of the lock-in amplifier, which is a technique used to separate a small, narrow-band signal from interfering noise. The lock-in amplifier acts as a detector and narrow-band filter combined. Very small signals can be detected in the presence of large amounts of uncorrelated noise when the frequency and phase of the desired signal are known. The lock-in amplifier is basically a synchronous demodulator followed by a low-pass filter.

Fig. 9 Spectrum of input signals.

Using Analog Device’s balanced modulator AD630, the lock-in amplifier provides more than 100dB (electrical) dynamic range, i.e. the pilot signal can be recovered from an uncorrelated noise signal approximately 100,000 times larger [8]. This provides enough room for the whole dynamical range of receiver (at most 50dB – optical).

4 Conclusions

The paper describes the design of an FSO receiver with active stabilization of APD parameters. The ADP gain is guaranteed within ±12% interval over temperatures from -30°C to +60°C without any compensation for LED parameters. With temperature compensation by a linear approximation, the gain stability can be improved in the order of magnitude.

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