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Receiver Performance of Laser Ranging Measurements between the Lunar Observer and a Subsatellite for Lunar Gravity Studies

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Abstract

The optimal receiver for a direct detection laser ranging system for slow Doppler frequency shift measurement is shown to consists of a phase tracking loop which can be implemented approximately as a phase lock loop with a 2nd or 3rd order loop filter. The laser transmitter consists of an AlGaAs laser diode at a wavelength of about 800nm and is intensity modulated by a sinewave. The receiver performance is shown to be limited mainly by the preamplifier thermal noise when a silicon avalanche photodiode is used. A high speed microchannel plate photomultiplier tube is shown to outperform a silicon APD despite its relatively low quantum efficiency at wavelengths near 800 nm. The maximum range between the Lunar Observer and the subsatellite for lunar gravity studies is shown to be about 620 Km when using a state-of-the-art silicon APD and about 1000 Km when using a microchannel plate photomultiplier tube in order to achieve a relative velocity measurement accuracy of 1 millimeter per second. Other parameters such as the receiver time base jitter and drift also limit performance and have to be considered in the design of an actual system.

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1 Introduction

Laser ranging between a host satellite and a subsatellite can provide information about the gravity field distribution of a planet or its moon. The relative velocity between the host satellite and the subsatellite, which varies with the gravity field, can be determined by measuring the Doppler shift of the signal reflected from the subsatellite. Since the divergence angle of a laser transmitter output beam is much smaller than that of a conventional RF Doppler radar, the propagation loss between the transmitter and the receiver of a laser ranging system is much lower than that of a RF system. NASA has proposed to use a laser ranging system in the Lunar Observer which is to be launched in the near future [1]. The Lunar Observer will serve as the host satellite. A subsatellite covered with corner cube retroreflectors will be spring launched from the Lunar Observer. This work studies the design and performance of the receiver of the proposed laser ranging system.

The basic system consists of a laser transmitter, a photodetector, and some signal processing circuitry. An intensity modulated laser beam is retroreflected by the subsatellite. The reflected light is collected by the receiver telescope and focused onto the active area of a photodetector. The received signal is then compared with the transmitted signal to determine the amount of Doppler shift. The hardware required is very simple since it is a direct detection optical ranging system. The average relative velocity is expected to be about 1 m/s and to change very slowly. The required accuracy of the proposed Doppler shift measurements is $\sigma_{v_{re}} \leq 1\ mm/s$. The measurements are to be made ten times each second [1].

The laser source being considered is an AlGaAs laser diode operated at a wavelength near 800nm. The laser diode can be intensity modulated either sinusoidally or pulsed on and off. The receiver for a sinusoidally modulated laser system measures the relative phase delay, as in a conventional Doppler radar. The receiver for

a pulsed laser system measures the round trip time of the reflected pulses, which in turn determines the distance and relative velocity between the host satellite and the subsatellite. However, present day digital electronics cannot determine round trip pulse transit times to accuracies of less than a few picoseconds which is required in order to determine relative velocities to an accuracy of 1 mm/s. Furthermore, the receiver noise bandwidth for detecting rectangular light pulses is much larger than that required for detecting sinusoidal signals. Therefore, sinusoidal modulation is preferred. The modulation frequency should be as high as possible in order to achieve the required accuracy in the Doppler shift measurements.

The photodetectors which render the highest receiver sensitivity are photomultiplier tubes (PMT) and silicon avalanche photodiodes (APD). Both devices can be operated in either photon counting mode or analog mode. In photon counting mode, a photocurrent pulse generated by a single photon absorption is detected by a discriminator as a discrete event. The discriminator is necessary to resolve photocurrent pulses from amplifier thermal noise. However, a discriminator which has a fixed threshold always has some time jitter in the rising edges of its output pulses due to the variation in the photocurrent pulse amplitude caused by randomness in the photodetector gain mechanism. We believe that the amount of jitter will not be acceptable in this application where the resolution of the Doppler shift measurement corresponds to a timing accuracy of a few picoseconds. Therefore a high speed PMT or an APD in analog mode operation is recommended for use as the photodetector in the proposed lunar gravity study instrument.

Signal processing will be based on the mathematical form of the optimal estimator for Doppler shift based on photon absorption times. Approximations and modifications will have to be made to arrive at an implementable suboptimal estimation scheme. The performance of the suboptimal estimator will be computed and compared with the theoretical limit achieved by optimal processing. We will first

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assume the photodetector is ideal with an infinitely wide electrical bandwidth and no additive noise. The effects on performance imposed by practical photodetectors will be studied afterwards.

2 Optimal Receiver Structure

The signal output from an ideal photodetector can be modeled as a Poisson random point process with the counting rate given by [2]

$$\lambda(t) = \frac{\eta P_r(t)}{hf} \tag{1}$$

where η is the quantum efficiency, $P_r(t)$ is the total received optical power, and hf is the photon energy. For sinusoidal intensity modulation, the received optical signal power can be written as

$$P_r(t) = LP_{Tx}[1 + m\cos(\omega_{Tx}t + \theta(t))] + P_b, \quad 0 < m < 1 \tag{2}$$

where L represents the sum of the round trip propagation loss from the host satellite to the subsatellite and the total transmission loss of the optics, P_{Tx} represents the average transmitted optical signal power, m is the modulation depth, $\theta(t)$ is the round trip phase delay to be estimated, and P_b is the received background radiation power. Gagliardi and Karp [3] have derived a maximum a posteriori (MAP) estimation scheme for estimating the instantaneous frequency, $d\theta(t)/dt$, of the received optical signal. An upper bound, the Cramer-Rao bound (CRB), for the variance of the estimator was also derived. The MAP frequency estimator consists of a bank of correlators each of which correlates the photodetector output signal with the logarithm of one plus a sinusoidal signal at a different frequency [3]. However, this MAP frequency estimator cannot be easily implemented since there would be too many correlators in order to resolve a 1 mm/s Doppler shift. Alternative suboptimal estimators have to be found.

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The first assumption we made is that the changes in Doppler shift of the received signal encountered in the lunar gravity study are slow enough that the receiver can estimate the instantaneous phase rather than frequency. The round trip Doppler frequency shift can be computed approximately as

$$\Delta f_D = 2\frac{v_{re}}{c}f_{Tx} \tag{3}$$

where v_{re} is the relative velocity, c is the speed of light, and $f_{Tx} = \omega_{Tx}/2\pi$ is the frequency of the intensity modulated signal from the laser transmitter. Since the anticipated relativity velocity between the host satellite and the subsatellite is 1 m/s and changes very slowly, the anticipated Doppler shift can be written as

$$\Delta f_D = 2\frac{f_{Tx}}{c}\left(\bar{v}_{re} + \Delta v_{re}(t)\right). \tag{4}$$

The first term in (4) is a constant equal to $2f_{Tx}\bar{v}_{re}/c = 50/3 \approx 16.7$ Hz for a laser intensity modulation frequency of 2.5 GHz. Only the second term in (4), which varies with the gravity field, needs to be estimated. Since the second term is much smaller than the first term, the amount of frequency shift to be determined is therefore much less than 1 Hz. This frequency shift can be modeled as a slowly varying phase shift in the received signal. We further assume that this phase shift is so slow that we can considered it as constant over the 0.1 second measurement interval. The received optical signal power (2) can now be written as

$$P_r(t) = LP_{Tx}[1 + m\cos(\,(1 + 2\frac{\bar{v}_{re}}{c})\omega_{Tx}t + \theta_{t_i})\,] + P_b, \quad 0 < m < 1 \tag{5}$$

where θ_{t_i} is the phase shift at the ith measurement time. The receiver in this case should be implemented as a phase estimator. The rate of the relative velocity between the two satellites is proportional to the rate of the Doppler frequency shift which can be obtained approximately from the successive phase measurements, as

$$\frac{d(\Delta f_D)}{dt} = \frac{1}{2\pi}\frac{d^2\theta(t)}{dt^2} \approx \frac{1}{2\pi}\frac{\frac{\Delta\theta_{t_i}}{\Delta t} - \frac{\Delta\theta_{t_{i-1}}}{\Delta t}}{\Delta t} = \frac{1}{2\pi}\frac{\theta_{t_i} - \theta_{t_{i-2}}}{\Delta t^2} \tag{6}$$
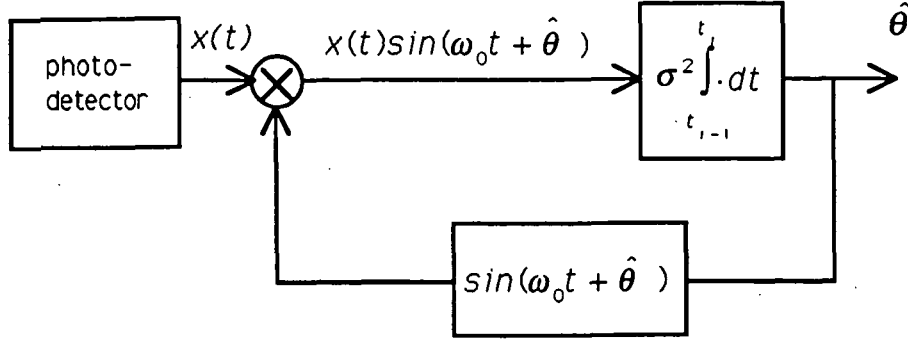
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Figure 1: Phase tracking loop as the MAP phase estimator.

where Δt is the measurement interval.

The second assumption in our analysis is that the total effective background radiation noise, which includes the contribution from the photodetector dark current and the amplifier thermal noise, is large compared to the sinusoidal component at frequency f_{Tx}. This assumption is valid when the distance between the host satellite and subsatellite is the greatest and the received optical signal is the weakest. It is only under this worst case situation that the receiver needs to be optimized.

The MAP phase estimator under the above conditions is the solution to the equation [3]

$$\hat{\theta} = \sigma^2 m \frac{L P_{Tx}}{P_b} \int_{t_i}^{t_i + \Delta t} sin(\omega_0 t + \hat{\theta}) dt \qquad (7)$$

where ω_0 is the average frequency given by $\omega_0 = \omega_{Tx}(1 + \bar{v}_{re}/c)$, and $\hat{\theta}$ is the MAP estimate of the phase which is assumed to be a zero mean Gaussian random variable with variance σ^2. An explicit solution to (7) is not immediately available. However, $\hat{\theta}$ can be interpreted as the output of a phase tracking loop shown in Figure 1 when the loop reaches its steady state. Such a phase tracking loop can be implemented approximately as an ordinary first order phase lock loop (PLL), shown in Figure 2, with the open loop gain, g, given by $g = \sigma^2 m L P_{Tx}/P_b$. The voltage controlled oscillator (VCO) acts as both the integrator and the sinewave generator since it is
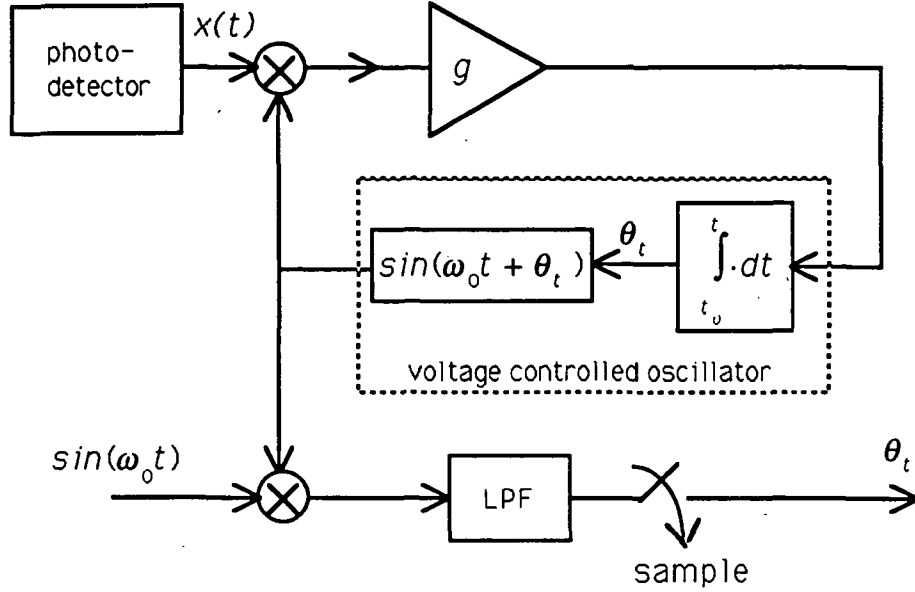
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Figure 2: First order phase look loop as an approximated MAP phase estimator.

the frequency, not phase, which is proportional to the control voltage. The phase of the VCO output is obtained by mixing the VCO output with the sinusoidal signal which modulates the laser transmitter, as shown in Figure 2.

Unfortunately, the performance of a first order PLL becomes very poor when the input signal phase is a function of time or the VCO free running frequency drifts [4]. A second order PLL is often preferred in practice because of its much superior tracking performance. A third order PLL may even be required if the time derivative of the Doppler shift of the received signal is large. The resultant PLL is no longer the same MAP estimator which satisfies (7). Nevertheless, it is probably among the best of the approximations which can be realized in practice.

3 Receiver Performance Analysis

The tracking error of a PLL is defined as the phase difference between the received signal and the VCO output. If the input to the PLL consists of a constant phase

sinusoidal signal and white noise, the variance of the tracking error can be computed as [4]

$$\sigma_\theta^2 = \frac{N_0 B_L}{P_s} \qquad (8)$$

where N_o is the one sided noise power spectral density, B_L is the one sided noise bandwidth of the PLL, and P_s is the power of the sinusoidal signal component at the PLL input. The PLL signal to noise ratio is often defined as $SNR_L = P_s/2N_o B_L$ [4]. The tracking error variance becomes $\sigma_\theta^2 = 1/2SNR_L$.

The signal output from the photodetector in our case consists of a sinusoidal signal and shot noise which is neither wide sense stationary nor exactly white. Nevertheless, equation (8) can still be used to a good approximation if the SNR_L is large and the loop bandwidth is small compared with the tracking frequency [5]. These two conditions are always satisfied in our situation. The average noise spectral density, $N_{oi} \doteq \lim_{T\to\infty}(1/T) \int_0^T N_o(t)$, should be used in (8) in this case.

We now compute the sinusoidal signal power and noise power spectrum at the input of the PLL. We can compute these equivalently by computing the signal and noise currents at the output of the photodetector assuming the amplifiers are noiseless. The thermal noise of the amplifiers can be treated as an equivalent current noise source which appears as additive to the photocurrent. The average photocurrent output from a PMT or an APD is given by

$$\overline{i_{ph}} = qG\lambda(t) + I_d = qG\frac{\eta}{hf}[LP_{Tx}(1 + m\cos(\omega_0 t + \theta_{t_i})) + P_b] + GI_{dbk} + I_{dsf} \qquad (9)$$

where q is the electron charge, G is the average gain of the photodetector, I_{dbk} is the bulk leakage dark current which is multiplied by the photodetector gain, and I_{dsf} is the surface leakage dark current which is not multiplied by the gain. The amplitude of the sinusoidal signal current is equal to

$$I_{sin} = qG\frac{\eta}{hf}LP_{Tx}m. \qquad (10)$$

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The one sided noise current spectral density is given by [6]

$$N_{oi}(t) = 2q^2G^2F\lambda(t) + 2qG^2FI_{dbk} + 2qI_{dsf} + \frac{d<i_{amp}^2>}{df} \qquad (11)$$

where F is the excess noise factor, defined as the ratio of the mean square to the square of the mean of the multiplication gain of the photodetector. The last term in (11) is the additive amplifier thermal noise current. If a PMT is used, the photoelectron multiplication gain is usually so large that the amplifier thermal noise can be neglected. If an APD is used, a preamplifier becomes essential and the amplifier thermal noise has to be considered. When a transimpedance preamplifier with an FET front end is used, the equivalent noise current spectrum may be written approximately as [6]

$$\frac{d<i_{amp}^2>}{df} = \frac{4KT_a}{R_f} + 2qI_g + \frac{4KT_a}{g_mR_f^2} + \frac{4KT_a}{g_m}(\omega_0C_i)^2 \qquad (12)$$

where K is Boltzmann's constant, T_a is the ambient temperature in Kelvin, R_f is the feed back resistance, I_g is the FET gate leakage current, g_m is the transconductance of the FET, and C_i is the input capacitance. Since the last term in (12) usually becomes dominant as the frequency exceeds several hundred megahertz, it is important to keep the input capacitance C_i as low as possible. The total average noise spectral density is equal to

$$N_{oi} = 2q^2G^2F\frac{\eta}{hf}LP_{Tx}m + 2qG^2FI_{dbk} + 2qI_{dsf}$$
$$+\frac{4KT_a}{R_f} + 2qI_g + \frac{4KT_a}{g_mR_f^2} + \frac{4KT_a}{g_m}(\omega_0C_i)^2. \qquad (13)$$

The phase variance of the tracking loop can then be obtained by substituting (10) and (13) into (8). The noise bandwidth of the PLL, B_L, should be close to, but no smaller than, the sampling rate of the Doppler shift measurements.

4   Results

We now compute the receiver performance based on the mathematical models given in the previous section. The laser source is assumed to be an AlGaAs laser diode with wavelength $\lambda = 805 \ nm$. The average transmitted optical signal power is $P_{Tx} = 200 \ mW$. The laser intensity modulation frequency is $f_{Tx} = 2.5 \ GHz$ and the modulation depth is $m = 1$. The maximum signal transmission loss between the transmitter and the receiver is $L = 10^{-12}$ [1] which includes all the losses due to the transmitter and receiver optics, laser beam propagation diffraction, and the fraction of the received signal split into the tracking subsystem.

4.1   Receiver Performance When using a *Slik* APD

The photodetector being considered here is a *Slik* silicon APD [7] which has a quantum efficiency of $\eta = 80\%$ at wavelengths near 800 nm. The hole to electron ionization coefficient ratio is $k_{eff} = 0.004$. The capacitance across the APD is about $C_{APD} = 0.25 \ pf$. Since the received optical signal level is very low, the optimal value of the average APD gain is expected to be larger than what the APD can provide under normal linear operation conditions. We used $G = 600$ in our computation which is about the highest value a *Slik* APD can provide without using a very complicated temperature controlled biasing subsystem. The excess noise factor is given by [8]

$$F = k_{eff}G + (2 - \frac{1}{G})(1 - k_{eff}). \tag{14}$$

The APD is assumed to be cooled to $-20°C$ in order to reduce the dark current. The surface leakage current is $I_{dsf} = 8 \ nA$. The bulk leakage current is assumed to be $I_{dbk} = 0.025 \ pA$. This value of the bulk leakage current is based on the value measured at room temperature and assuming a decrease by a factor of 40 when cooled to $-20°C$ [9].

A state of the art transimpedance preamplifier [10] is assumed which has a band-

width of over 2 GHz with an effective transimpedance of $R_f = 4.7\ K\Omega$. We assume the advanced FET described in [11] will be used as the front end of the transimpedance amplifier. The transconductance of the FET is $g_m = 65\ mS$. The FET gate leakage current is $I_g = 10\ nA$. The input capacitance of the amplifier itself is $C_{amp} = 0.6\ pf$. The total input capacitance is therefore $C_i = C_{APD} + C_{amp} = 0.85\ pf$. The ambient temperature is $T_a = 300°K$. The noise bandwidth of the PLL is assumed to be $B_L = 20\ Hz$ which is twice the sampling rate of the measurements. Table 1 contains a complete list of all the parameter values used in the computation.

The amplitude of the sinusoidal component of the received signal according to (10) is $I_{sin} = 62.3pA$. The resultant equivalent noise current spectral density according to (13) is $N_{oi} = 4.9 \times 10^{-23}\ A^2/Hz = (7.0\ pA/\sqrt{Hz})^2$. The resultant PLL signal to noise ratio is

$$SNR_L = \frac{I_{sin}^2/2}{2N_{oi}B_L} = 0.99 = 0.0\ dB. \tag{15}$$

The resultant phase tracking error according to (8) is $\sigma_\theta = 0.71\ radian = 40.7\ degree$, which is clearly much too large to achieve the 1 mm/s Doppler shift measurement accuracy which requires $\sigma_{\theta_{max}} = 0.105\ radian = 6.0\ degree$.

The Cramer-Rao lower bound for the receiver performance under no background and amplifier noises is

$$\sigma_{CRB} = \left(\frac{1}{\eta L P_{Tx} T_s / hf}\right)^{1/2} = 3.92 \times 10^{-3}\ radian = 0.225\ degree. \tag{16}$$

The major factor which prevents the receiver performance from reaching the theoretical limit is the amplifier thermal noise, especially the noise due to the capacitance at the input of the preamplifier. Of the total noise at the input of the preamplifier, 92.6% is due to the input capacitance and 7.19 % is due to the 4.7 $K\Omega$ feedback resistor. The shot noise generated by the input optical signal contributes only 0.133% of the total noise. The shot noise from the APD bulk leakage current is about one quarter of that generated by the input optical signal. Therefore, the receiver perfor-

mance would also be limited by the APD bulk leakage current if the APD was not cooled to $-20°C$. The contributions to the total noise from the APD surface leakage current, I_{dsf}, and the FET gate leakage current, I_g, are negligibly small.

The PLL signal to noise ratio needed to achieve the required measurement accuracy is $SNR_L \geq 45.4$. The received optical signal power has to be about 7 times larger in order to achieve the desired SNR_L with the present system design. Alternatively, the maximum range between the Lunar Observer and the subsatellite has to be reduced from the proposed 1004 Km to about 620 Km in order to maintain the required 1 mm/s relative velocity measurement accuracy.

4.2   Receiver Performance When Using a PMT

The PMT being considered here is a Hamamatsu R2809U-11, which is similar to a R2809U [12] but has a GaAs photocathod. The anode of the PMT consists of two microchannel plates. The quantum efficiency of the PMT is typically 3.0% but may be increased to 20% in the future. The total anode dark current is 1 nA. The current amplification gain is 5×10^5. Therefore, the bulk leakage current is approximately $I_{dbk} \approx 1nA/5 \times 10^5 = 2.0 \ fA$. The excess noise factor is taken to be $F = 2.0$. A summary of these PMT parameters used in the calculations can be found in Table 1.

Since the gain of the PMT is more than 800 times that of the APD and the shot noise spectral density is proportional to the square of the gain, the receiver in this case is operated in the shot noise limited regime and the amplifier thermal noise becomes negligible. The phase estimation error according to (10), (13), and (8) when neglecting the PMT electron transit time is $\sigma_\theta = 0.0997 \ radian = 5.71$ degree if $\eta = 3.0\%$ and $\sigma_\theta = 0.0326 \ radian = 1.87$ degree if $\eta = 20\%$.

We now consider the effect of jitters in the PMT output pulses due to the electron transit time spread (TTS). It has been shown that the effective signal power under

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this condition becomes [3, p. 137]

$$P_s = |E\{e^{-j\omega_0 \Delta\tau}\}|^2 P_{s_{\Delta\tau=0}} \tag{17}$$

where $\Delta\tau$ represents the electron transit time variation. The shot noise is not affected by the TTS. Since the magnitude of the characteristic function of the jitter $|E\{e^{-j\omega_0 \Delta\tau}\}|$, is always less than or equal to unity, the jitters in the PMT output pulses effectively attenuate the power of the useful signal component and consequently reduce the signal to noise ratio. If the jitter is assumed to be a Gaussian random variable with mean zero and variance $\sigma_{\Delta\tau}$, the characteristic function of the jitter is $E\{e^{-j\omega_0 \Delta\tau}\} = e^{-\omega_0^2 \sigma_{\Delta\tau}^2 / 2}$. The amount of the attenuation is equal to $e^{-\omega_0^2 \sigma_{\Delta\tau}^2}$ according to (17).

The lowest TTS ever achieved with a Hamamatsu R2809U PMT according to the data sheet is 22.5 ps which is measured at the full width half maximum (FWHM) points [13]. The standard deviation of the TTS should be approximately $\sigma_{\Delta\tau} = 10$ ps for the Gaussian distribution. The amount of the attenuation at 2.5 GHz is $e^{-\omega_0^2 \sigma_{\Delta\tau}^2}$ = 0.976 = 0.1 dB. The typical value of TTS is 55 ps at FWHM according to the data sheet [12, 13]. The corresponding TTS standard deviation is $\sigma_{\Delta\tau} = 27$ ps. The resultant attenuation becomes $e^{-\omega_0^2 \sigma_{\Delta\tau}^2}$ = 0.835 = -0.78 dB. Therefore, the TTS of these PMTs will not be a major factor which limits the receiver performance. The phase estimation error calculated using the typical PMT parameter values (55 ps TTS at FWHM and $\eta = 3.0\%$) is $\sigma_\theta = 0.109$ radian = 6.27 degree. This performance is very close to what is required in order to achieve the 1 mm/s measurement accuracy ($\sigma_{\theta_{max}} = 6.0$ degree). The maximum ranging capability of this system will be close to the proposed 1004 Km. The microchannel PMTs will outperform the silicon APDs according to the above analysis.

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5 Comments

Other parameters in practice may also limit the performance of the proposed laser ranging system. One issue to be considered is the jitter and drift of the time base of the ranging system which we assumed as perfect in the analysis. The random jitter of the master clock oscillator which modulates the laser has a similar effect as the transit time spread of PMTs. The rms jitters of the master clock oscillator and the VCO of the phase lock loop should not exceed 45 ps or the receiver performance will deteriorate very rapidly. The gradual frequency drift of the master clock oscillator will manifest itself as a the Doppler shift measurement error. Therefore, the clock oscillator frequency drift must be highly stable over the measurement interval, typically drifting by no more than 0.018 cycle per 100ms to ensure a $1mm/s$ Doppler shift measurement accuracy. Variations in phase delay with temperature due to the amplifiers and other circuit components in the system appear as phase estimation errors and therefore have to be tightly controlled.

Table 1: Parameter values in the calculation of the receiver performance.

AlGaAs Laser Transmitter
wavelength	$\lambda = 805$ nm
average output power	$P_{Tx} = 200$ mW
modulation frequency	$f_{Tx} = 2.5$ GHz
modulation depth	$m = 1.0$

Transmission Channel
total losses	$L = 10^{-12} = -120$ dB
received background radiation power	$P_b = 0$ W
average Doppler shift	$f_{Tx}\bar{v}_{re}/c = 50/3$ Hz

APD Photodetector (Slik)
quantum efficiency at $\lambda = 805$ nm	$\eta = 80\%$
average gain	$G = 600$
ionization coefficient ratio	$k_{eff} = 0.004$
excess noise factor	$F = 4.4$
bulk leakage current	$I_{dbk} = 25$ fA
surface leakage current	$I_{dsf} = 8$ nA
capacitance	$C_{APD} = 0.25$ pf

Microchannel PMT
quantum efficiency at $\lambda = 805$ nm	$\eta = 3.0\%$ typical, 20% projected
average gain	$G = 5 \times 10^5$
excess noise factor	$F = 2.0$
bulk leakage current	$I_{dbk} = 2.0$ fA
surface leakage current	$I_{dsf} \approx 0$
transit time spread at FWHM	22.5 ps min., 55 ps typical

Transimpedance Preamplifier
feedback resistance	$R_f = 4.7$ KΩ
1st stage FET transconductance	$g_m = 65$ ms
FET gate leakage current	$I_g = 10$ nA
FET input capacitance	$C_{amp} = 0.6$ pf
total input capacitance	$C_i = C_{APD} + C_{amp} = 0.85$ pf
ambient temperature	$T_a = 300°K$

Receiver Phase Lock Loop
loop filter type	2nd order active
noise bandwidth	$B_L = 20$ Hz

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