Design of Photoelectric Amplification Circuit for Laser Fuze

Yulong Zhang, Yachao Guo, Guanglin He *
School of Mechatronic Engineering Beijing Institute of Technology, Beijing, China

*Corresponding author e-mail: heguanglin@bit.edu.cn

Abstract. A photodetection system based on the PIN photodetector was proposed in order to amplify the weak narrow pulse echo signals received in the laser fuze without distortion. The circuit model of this photodetection system includes a preamplifier circuit, a main amplifier circuit. Computer simulations were carried out in the MULTISIM and TINA environment, and the gain of the photoelectric amplification circuit was 119.9dB and the bandwidth was 89.46MHz. The simulation results show that the optoelectronic receiving system realizes the low-noise and high-gain broadband application, which meets the designed requirements.

Keywords: photoelectric detection; amplifier circuit; low noise; high gain.

1. Introduction
The laser fuze can detect the target distance and position quickly and accurately with narrow detection pulses and multiple modulation modes. It has the advantages of high fuze-warhead coordination efficiency and strong anti-electromagnetic interference ability. In the laser fuze receiving system, the design of the photoelectric conversion system has significant influence on measuring distance and ranging accuracy in [1, 2]. Narrow pulse laser echo signals focusing on the receiving field change rapidly and their power is very low. Laser echo signals focus on the photoelectric sensor and generate microampere-level electrical signals. The pre-amplifier circuit and the main amplifier circuit convert current signals into voltage signals while reducing various noise, and send to the rear level circuit for shaping and conditioning. In addition, narrow pulse width signals contain large amounts of high frequency components. Hence, to restore the target information as much as possible, the photoelectric amplifier circuit needs broad working bandwidth. Therefore, the amplifier must have a large enough slew rate and broad bandwidth and a high signal-to-noise ratio (SNR), so that the echo signal can be amplified without distortion.

The pre-amplifier circuit realizes the pre-processing of the electrical signals output by the photoelectric sensor, amplifies the photoelectric conversion signals with possibly high SNR ratio, and outputs them to the rear level circuit. Reference [3] introduced an optimization design of APD receiving circuit. The working principle and performance of the circuit were verified that the design was valid and feasible. W. Song, E.A. Hasaneen and C. Cabrera respectively studied a weak signal detection preamplifier circuit with low cost, low noise and high SNR ratio in paper [4,5,6,7]. The experimental results and circuit analysis indicated that the circuit could detect and amplify the weak signal. Hojong Choi designed an integrated circuit consisting of a preamplifier with Sallen-Key Butterworth filter in [8]. The above research methods focus on the analysis and design of the preamplifier circuit, ignoring the
influence on the parasitic capacitance of the feedback resistor or the resistor network. Moreover, it is impractical for a single chip capacitor to achieve such a low value. Focusing on a semiconductor photodiode, this article adopts a novel design of the preamplifier circuit and the main amplifying circuit. Both theoretical analysis and empirical simulation results validate that the designed circuit realizes the low-noise broadband high gain amplification of the 10ns-pulse-width echo signal.

2. Design of Photoelectric Circuit

The photoelectric amplifier circuit is generally formed by cascading multi-stage amplifier circuits, which can respond to the rising edge and pulse width of the signal waveform without distortion. In order to get good phase-frequency and amplitude-frequency characteristics of the amplifier circuit and enhance signal-to-noise ratio, amplifier circuit must achieve sufficient wide working bandwidth. As for pulse signals, the bandwidth containing 90% energy is generally regarded as $B_\omega$, and the approximate value of $B_\omega$ is usually $0.89/\tau_0$. If the detected laser echo signal has a pulse width $\tau_0$ of 10ns, then photoelectric amplifier circuit achieves better performance when the amplifier circuit bandwidth is no less than 89MHz.

Fig.1 shows the schematic circuit diagram of the preamplifier circuit and the main amplifier circuit. Active voltage biasing circuit for the PIN photodiode operates in the photoconductive mode. The photoconductive mode enables the photodiode to response faster, which is more suitable for pulse signal detection. In this mode, the photocurrent is unrelated to the load, and the photocurrent is proportional to the light radiation intensity. The selected PIN photodiode is the GT101 photodiode produced by China Electronics Technology Group, and its wavelength response range is from 400nm to 1000nm. Under the condition of 905nm incident wavelength and 15V reverse-biased voltage, the responsivity of PIN photodetector can achieved 0.01 A/W, the response time and the junction capacitance are 5ns, 2.5pF, respectively. In the pre-amplification circuit: TIA amplifier OPA818 is a new 13V supply, 2.7GHz gain bandwidth product(GBP), voltage feedback operational amplifier FET-input operational amplifier produced by TI. The OPA818 showcases low 2.4pF total input capacitance, 2.2nV/√Hz input noise, and fast slew rate of 1400V/µs providing high large-signal bandwidth and low distortion. The resistor R1 is to offset the influence of the bias current, the capacitors C4 and C5 can eliminate the high frequency noise generated by the resistor R1, and C7 is an inter-stage coupling capacitor used to isolate DC signals. C1, C2, C3 form a capacitive tee network to compensate the feedback resistance. C6, C8 and C9, C10 are power supply filter capacitors to suppress noise.

3. Noise and Stability Analysis of Preamplifier Circuit

3.1. Stability analysis of preamplifier circuit

The principle of the PIN photoelectric detection circuit can be simplified as displayed in Fig. 2.
An amplifier's stability depends on its loop gain, which is defined as a product of the open loop gain \( A_{OL} \) and the feedback factor \( \beta \). The analysis of related literature demonstrates that the transfer function of the amplifying circuit system is:

\[
\frac{V_{out}}{I_s} = \frac{Z_F}{1 + 1/[\beta \cdot A_{OL}(j\omega)]}
\]  

(1)

The noise gain of an op amp is just the inverse of its feedback factor \( \beta \):

\[
\frac{1}{\beta} = 1 + \frac{Z_s}{Z_F}
\]  

(2)

\( Z_s \) is the network input impedance.

\[
Z_s = \frac{1}{j\omega C_s}
\]  

(3)

Where \( C_s = C_{CM} + C_{DIFF} + C_p + C_{ST} \).

\( Z_F \) is the feedback network impedance.

\[
Z_F = R_F \parallel \frac{1}{j\omega(C_F + C_{RF})}
\]  

(4)

Where \( C_{RF} \) is the parasitic capacitance of the feedback resistor.

The noise gain is plotted in dotted line in Fig. 3 alongside the single pole, open-loop gain model of the amplifier. The open-loop gain \( A_{OL} \) curve starts to reduce at the dominant pole at a rate of 20dB per decade and passes through the 0dB line at the GBW frequency. At low frequencies, the impedance of the total input capacitance \( C_s \) will be very high and will thus act as an open circuit. The low frequency curve maintains a stable unity gain and is unity gain resistor feedback, thus the amplitude of noise gain \( 1/\beta \) curve is zero dB in the low frequency part. The gain of the amplifier system is determined by the feedback resistor \( R_f \). As the frequency increases, the impedance of the source capacitor \( C_s \) begins to decrease, and
$R_f$ and $C_i$ generate a zero point $f_z$ in the noise gain curve. The impedance of the capacitor at this frequency is equal to the value of the feedback capacitor. Starting from the zero point $f_z$, the noise gain curve starts to increase at the rate of 20 dB per decade. The total phase change due to the zero point is going to be 90 degrees. Similarly, the open loop gain curve starts to reduce at the rate of 20 dB per decade.

The point of intersection, which is the loop-gain crossover, will be the geometric mean of the zero frequency at zero point $f_z$ and the amplifier’s gain bandwidth product. This loop-gain crossover $f_0$ will thus occur. The Loop Gain is equivalent to the subtraction of the open loop gain $A_{OL}$ and the noise gain $1/\beta$, thus the resultant Loop Gain does have the two-pole response. The dominant pole affects a total of 90 degrees of phase shift. However, due to the zero at $f_z$, the phase continues to reduce. And at loop-gain crossover frequency, the total phase shift around the feedback loop is 180 degrees, which means the resultant phase margin is 0 degrees. Actually, insufficient phase margin cause $A_{\beta} \geq 1$ and self-oscillation. If the two frequency response curves approach at a rate of 40dB per decade, the TIA circuit will be unstable.

In all cases, the phase compensation circuit can be used to stabilize the circuit. Generally, a capacitor parallel with the feedback resistor can provide the necessary compensation to ensure sufficient phase margin. As shown in solid line of Fig.3, in addition to shifting the pole, the capacitor $C_f$ also introduces a pole $f_p$ in the noise gain curve; for the loop gain, $R_f$ and $C_f$ introduce a zero in the Loop Gain curve to compensate for the feedback network. A better compensation scheme is to introduce 45 degrees of phase margin at the loop-gain crossover, where $f_p=f_0$. Theoretically, the correct phase margin for a bipolar system should be at least 45 degrees. The step response of the phase margin circuit is 22.5%. In fact, a stable circuit design can be achieved when the phase margin ranges from 45 degrees to 70 degrees. Considering the amplifier bandwidth, resistance, capacitance and parasitic capacitance, slight change of the above components will have great disadvantages to the circuit with 45 degrees of phase margin, so 60 degrees or a slightly larger phase margin can be appropriately selected.

Fig. 4 shows the phase frequency characteristic curve of different phase compensation capacitance C. Setting the feedback capacitance much smaller than this value of $C_f$ will push the pole frequency further out. And this will result in a less-than-ideal phase margin. And the closed-loop gain amplitude-frequency characteristic curve will have a peak, causing obvious ringing in the impulse response. Similarly, increasing the feedback capacitance to be much greater than $C_f$ will pull pole frequency closer to the zero. This will result in an over-damped response. However, the closed-loop bandwidth is going to be a lot lower than before. In such applications, it is necessary to find the minimum value of the feedback compensation capacitor $C_f$ to eliminate self-oscillation and minimize ringing in the impulse response. Besides, choosing a slightly larger compensation capacitor can provide enough protection bandwidth. Under the premise of ensuring sufficient bandwidth, it is recommended to use a slightly larger capacitor for compensation.

When the TIA amplifier is selected, the gain bandwidth product and the input parasitic capacitance are also determined accordingly. After setting the feedback resistance $R_f$, the closed-loop $f_{3dB}$ bandwidth can be expressed as

$$f_{3dB} = \frac{GBP}{\sqrt{2\pi R_f \left( C_i + C_f \right)}}$$

The feedback capacitor $C_f$ makes the operational amplifier work stably, and reach the best second-order Butterworth response. The closed-loop system operating in a critically damped state, the amplifying circuits have the maximum flat frequency response, and the phase margin is 65.5 degrees. The value of $C_f$ must meet the following conditions.

$$\frac{GBP}{\sqrt{4\pi R_f C_i}} = \frac{1}{\frac{2\pi R_f}{C_f}}$$
Equations (5) and (6) indicate that the higher the closed-loop bandwidth, the smaller the value of the feedback capacitance required to obtain the Butterworth response, and it may even be tens of femtofarads. In such cases, the feedback capacitance value may be difficult to implement in hardware. Achieving extremely low values of feedback capacitance may be impractical for the lack of availability of chip capacitors less than 0.2pF. Due to the PCB parasitic effects, external capacitors are inferior. In addition, the feedback resistor itself has a parasitic capacitance, and this capacitance value should be added, which increases the difficulty for designers to achieve extremely low feedback capacitance values. The parasitic effect can be reduced by removing the amplifier, reverse input terminal, output terminal, and the ground plane and power plane under the feedback path. A capacitor T-type network is introduced to achieve extremely small capacitor values, as shown in the circuit above. As shown in Fig.1, the effective capacitance $C_{eq}$ is given by the following equation.

$$C_{eq} = C_1C_2/(C_1+C_2+C_3)$$  \hspace{1cm} (7)

The T-type network forms a capacitive attenuator from input to output through $C_1$ and $C_3$, and forms a capacitive attenuator from output to input through $C_2$ and $C_3$. When the value of $C_3$ is higher than $C_1$ or $C_2$, only part of the signal is output through $C_1$. This results in a very small shunt current input through $C_1$, and this reduced shunt current affects the performance equivalent to a very small capacitor. In order to find a suitable capacitance value for the T-type network, the capacitors $C_1$ and $C_2$ can be an equal and achievable smaller value. Let $C_{eq} = C_T$, then the value of $C_3$ can be calculated as follows

$$C_3 = [C_1C_2(C_1+C_2)C_{eq}] / C_{eq}$$  \hspace{1cm} (8)

3.2. Noise analysis of preamplifier circuit

Microampere electric current signal will be completely submerged in noise, so it is necessary to design a low-noise and high-gain amplifier circuit. According to the calculation formula of the total noise figure of the multi-stage amplifier (the Forex formula), the noise figure can be derived as

$$F_n = F_1 + F_2 - \frac{1}{G_1} + \frac{F_1 - 1}{G_1G_2} + \ldots + \frac{F_n - 1}{G_1G_2\ldots G_n}$$  \hspace{1cm} (9)

$F_1, F_2, \ldots, F_n$ are the noise coefficients of each level of amplifying circuit, and $G_1, G_2, \ldots, G_n$ are the gains of each level of amplifying circuit, respectively.

Equation (9) shows that if the gain of the first stage amplifying circuit is large enough, the influence of the subsequent circuits can be ignored. The noise of the preamplifier circuit is the main factor affecting the amplifying circuits’ noise performance.
For DC-coupled, pulse-oriented applications, there are three parts of noise affecting the circuit performance. They are shot noise from the photodiode, thermal noise and amplifier's noise in preamplifier circuit, and environmental noise. Ambient noise can be reduced or eliminated by shading, electrical shielding, filtering circuit, power supply filtering, and other methods [7]. Working in photoconductive mode, the photodiode has a small amount of current even without light irradiation, which is dark current \( I_D \). Thus shot noise can be represented by the square root of noise current, that is \( I_n = \sqrt{2q(I_D + I_p)\Delta f} \). But its amplitude will not change versus frequency. As for this part, we can choose the appropriate type of photodiode and limit the working bandwidth to reduce it. The dark current of the Si-PIN tube can be less than 1nA. The noise model of the preamplifier circuit is displayed in Fig.5. Four contributing factors are the noise contributed by the voltage noise \( e_n \), the noise contributed by the current noise of the op amp \( i_n \), the noise contributed by the feedback resistor element \( E_{nR} \), and there is the total output voltage noise due to the amplifier's voltage noise and the noise gain shape. \( R_d \), the internal resistance of the diode, can reach the magnitude of several or even hundreds of megohm. Hence, the current flowing through \( R_d \) is such small that this loop circuit can be regarded as open circuit. The equivalent noise voltage and noise current at the input end can be equivalent to an ideal noise-free resistor as a noise model in series with a voltage source. The root mean square of the noise voltage source is \( E_{nR} = \sqrt{4kTR\Delta f} \), where \( k = 1.38 \times 10^{-23} \text{J/K} \) is the Boltzmann's constant; \( T \) is the thermodynamic temperature (K); \( \Delta f \) is the noise bandwidth and consistent with the \( f_{3\text{dB}} \) bandwidth.

The equivalent input current noise is the sum of root mean squares of these factors. The total input current noise \( I_n \) can be expressed as follows

\[
I_n = \sqrt{\frac{E_n^2}{R_f^2} + \frac{I_n^2}{R_f} + \frac{\left(E_n2\pi C_f\Delta f\right)^2}{3}} \tag{10}
\]

And total output voltage noise \( V_N \) equals to \( I_nR_f \). The output current of the photodiode is amplified by the TIA amplifier and the output voltage is \( i_p R_f \). Thus the signal-to-noise ratio of amplifier circuit can be given as

\[
\text{SNR} = \frac{i_p R_f}{e_{ab}} = \frac{R_f}{4kT} \sqrt{\frac{i}{E_{nR}}} \tag{11}
\]

As seen in (11), increasing \( R_f \) helps to improve SNR. However, due to the limited GBP of the operational amplifier, increasing \( R_f \) will lead to decrease in working bandwidth. Therefore, Appropriate \( R_f \) is a way to optimize SNR while maintaining the operating bandwidth.

---

**Figure 5.** Equivalent noise model of pre-amplifier circuit
3.3. Simulation Results of pre-amplifier circuit with Multisim and Tina Software

Assuming the feedback resistance $10\,\text{k}\Omega$, $C_f$ is calculated to be about $240\,\text{fF}$ by (3)–(6). In the capacitor T-type network, setting $C_1$ and $C_2$ equal to $2.2\,\text{pF}$, $C_3$ calculated by (7) is about $15\,\text{pF}$. Optimize the value of the feedback capacitor through parameter-optimization-sweep function of the TINA-Pspice software. Parameter $C_f$ sweep range from $0.220\,\text{pF}$ to $0.280\,\text{pF}$, the sweep step size is $0.015\,\text{pF}$, and the time domain parameter sweep curve is revealed in Fig.6. As the feedback capacitance increases, the overshoot of the signal output by the TIA op amp gradually decreases. At the $C_f$ value of $0.265\,\text{pF}$, the overshoot almost completely disappears, and the self-excitation of the circuit is basically eliminated. However, the closed-loop gain also decreases. In the AC sweep, the frequency domain sweep curve shows that the $f_{-3\text{dB}}$ bandwidth of the output signal becomes smaller with the increase of feedback capacitor, and the peak gain decreases from nearly $1.2\,\text{dB}$ to $0\,\text{dB}$.

Considering the compromise between the working bandwidth and loop gain, when the bandwidth requirement is met, a larger $C_f$ is selected as $0.258\,\text{pF}$. In this case, the amplitude-frequency response curve of the preamplifier circuit is shown in Fig.7. The closed-loop bandwidth is about $89.19\,\text{MHz}$, which is slightly larger than required $89\,\text{MHz}$. Given the bell-shaped input signal with a bandwidth of $10\,\text{ns}$ and an amplitude of $1\,\mu\text{A}$, the output signal of the pre-amplifier circuit has a pulse width of $17\,\text{ns}$ and an amplitude of $-9.82\,\text{mV}$. The inverting amplification gain of $79.84\,\text{dB}$ is approximately equal to theoretical calculation results. The signal waveform is shown in Fig.7(a), where INPUT is the input current signal, and Out-AMP1 is the output voltage signal of preamplifier circuit.

The input voltage noise $E_n$ and the input current noise density $I_n$ of the transimpedance amplifier OPA818 are $2.2\,\text{nV}/\sqrt{\text{Hz}}$, $2.5\,\text{fA}/\sqrt{\text{Hz}}$ respectively. $C_f$ is $0.258\,\text{pF}$ and $\Delta f$ is $89.31\,\text{MHz}$. We find that $I_N$ is $1.31\,\text{pA}/\sqrt{\text{Hz}}$. The simulation result of total output referred TIA noise of the preamplifier circuit with Tina software is exhibited in Fig.8. The flat-band output voltage noise is $13.09\,\text{nV}/\sqrt{\text{Hz}}$, and that is equivalent to $1.31\,\text{pA}/\sqrt{\text{Hz}}$ of input referred current noise as the gain is $10\,\text{K}\Omega$. The noise in relatively low frequency region where the noise gain of the amplifier is $1\,\text{V/V}$ is dominated by thermal noise of the feedback resistor. At mid frequencies beyond the zero formed by $R_f$ and $C_s$, the noise gain of the amplifier amplifies the voltage noise of the amplifier. The amplifier's noise starts to become the dominant noise contributor from this frequency onwards before the output noise starts to roll off at frequencies beyond $f_{-3\text{dB}}$ bandwidth.
4. Analysis and simulations of main amplifier circuit with Multism and Tina Software

The main amplifier circuit further amplifies the signal output by the pre-amplifier circuit, as shown in the Fig.1. The main amplifier circuit adopts the TIA amplifier OPA657 with extremely low input current noise density of 1.3fA/\sqrt{Hz} and high bandwidth gain product of 1.6GHz. The signal of pre-amplifier circuit is input through R4, and output after amplification of U2 and U3. Fig.9(a) and (b) display the simulation results in Tina software. The simulation results reveal that working bandwidth of the main amplifier circuit is 91.06MHz, which is slightly larger than that of the preamplifier circuit and meets the design requirements. The amplified output signal of the main amplifier circuit is 979.2mV, and the gain is about 39.97dB, which is approximately equal to the calculated results.
Figure 10. Amplitude-frequency response curve of the amplifying circuit system

Fig.10 illustrates that the working bandwidth and the gain of the amplifying circuit system is 89.46MHz, 119.9dB, respectively.

5. Conclusions
In this paper, a pulsed laser photoelectric detection system is designed with the high-bandwidth, ultra-low noise feedback operational amplifier OPA818 and the low-noise, high-gain amplifier OPA657 as the core components. The parasitic effects in the amplifying circuit are reduced through the capacitive T-network, and the high-gain and low-noise amplification of weak photoelectric signal is achieved. Theoretical analysis and optimized simulations prove that the design can improve the signal-to-noise ratio of the detection system, which provides a reference for the research of photoelectric detection technology.

Future work will further focus on the experimental implementation of detection and amplification of weak photoelectric signals. And optimize the design method of amplifying circuit in receiving circuits of laser fuze.

Acknowledgments
We gratefully acknowledge the valuable cooperation of Dr. Guangsong Ma (National Key Laboratory of Science And Technology Fuze Dynamic Characteristics) for his consultations and useful literature support.

References
[1] B. Tian, T. Li, T. Li and W. Li, "An Adaptive Precision Array Laser Fuze Detection Simulation Method," 2018 12th International Symposium on Antennas, Propagation and EM Theory (ISAPE), Hangzhou, China, 2018, pp. 1-6, doi: 10.1109/ISAPE.2018.8634117.
[2] Chao Han, Jia-Hao Deng and Jing Gao, "Design of Simulation Software for Laser Fuze Target Detection and Recognition", Transactions of Beijing Institute of Technology, vol. 35, no. 2, pp. 149-153, 2015.
[3] W. Song, S. Dong and Z. Jia, "The design and realization of APD receiving circuit used in M-ary VLC", 2015 IEEE 4th Global Conference on Consumer Electronics (GCCE), pp. 543-544, 2015.
[4] P. Bartona, M. Ammana, R. MartinaK. "Ultra-low noise mechanically cooled germanium detector." Nuclear Instruments and Methods in Physics Research Section A: Accelerators, Spectrometers, Detectors and Associated Equipment , vol. 23, pp.17-23,2016
[5] L. G. El-Fadali, E. A. Hasaneen, A. I. Galal and H. F. Hamed, "An Ultra-Low-Power, Low-Noise, Linear Preamplifier with Wide Dynamic Range for Electret Microphones," 2018 30th
International Conference on Microelectronics (ICM), Sousse, Tunisia, 2018, pp. 104-107, doi: 10.1109/ICM.2018.8704107.

[6] C. Cabrera, R. Caballero, M. C. Costa-Rauschert, C. Rossi-Aicardi and J. Oreggioni, "Low-Voltage Low-Noise High-CMRR Biopotential Integrated Preamplifier," in IEEE Transactions on Circuit and Systems I: Regular Papers, doi: 10.1109/TCSI.2020.3035357.

[7] X. Sun, "Optimization of detector arrays and circuits targeted for precision calculation in infrared laser interferometer," 2012 12th International Conference on Numerical Simulation of Optoelectronic Devices (NUSOD), Shanghai, 2012, pp. 75-76, doi: 10.1109/NUSOD.2012.6316537.

[8] H. Choi, X. Li, S. Lau, C. Hu, Q. Zhou and K. K. Shung, "Development of integrated preamplifier for high frequency ultrasonic transducer," 2010 IEEE International Ultrasonics Symposium, San Diego, CA, 2010, pp. 1964-1967, doi: 10.1109/ULTSYM.2010.5935544.