Large-Bandwidth High-Gain Low-Noise Transimpedance Amplifier for Scanning Tunneling Microscope Postprint

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Abstract

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Full Text

Preamble

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LABORATORY TECHNIQUES Large-Bandwidth High-Gain Low-Noise Transimpedance Amplifier for Scanning Tunneling Microscope Ying-Xin Liang,* a Anyang Normal Univ, Sch Phys & Elect Engn, Anyang, Henan, 455000 China *e-mail: cryoliang@qq.com Received July 9, 2022; revised September 1, 2022; accepted November 21, 2022

Abstract

In this work, a design of large-bandwidth high-gain low-noise transimpedance amplifier (TIA) for scanning tunneling microscope (STM) is proposed. The simulations show that the proposed TIA has a bandwidth higher than 200 kHz, two orders of magnitude higher than those of conventional commercial TIAs for STM. At low frequencies, the noises of the proposed TIA are almost the same as the conventional commercial ones with the same transimpedance gain. At high frequencies, its calculated input equivalent noise voltage power spectral density (PSD) is 40 nV^2/Hz and its input equivalent noise current PSD is 3.2 fA^2/Hz at 10 kHz. The corresponding values are 23 nV^2/Hz and 88 fA^2/Hz at 100 kHz. The STM with the proposed TIA can meet the needs of fast high-quality STM imaging measurements and fast high-energy-resolution scanning tunneling spectra measurements for low-conducting materials, such as complex organic systems and wide bandgap semiconductors.

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1. Introduction

High-performance transimpedance amplifier (TIA) is a key element in scanning tunneling microscopy/spectroscopy [1, 2]. For the application of scanning tunneling microscopy (STM) to investigate complex organic systems, such as biological macromolecules (DNA [3], RNA [4–6], and proteins [7], etc.) and membranes [8], one of the main problems is that biological samples can sustain only extremely low currents [9]. Furthermore, for many biological samples, due to their variability, real-time observation is required, so fast STM imaging measurements with high quality and fast scanning tunneling spectroscopy (STS) measurements with high energy resolution are necessary. For the application of STM to investigate low-conductivity semiconductor materials, such as wide bandgap semiconductors, the tunnel current is typically less than 100 pA [10]. In the semiconductor industry, fast high-quality STM imaging measurements and fast high-energy-resolution STS measurements of wide bandgap semiconductors are urgently needed to meet high production efficiency requirements. For conventional TIAs in STM with gain higher than 1 GV/A, the typical bandwidth is less than 1 kHz [1, 11], which is too low for fast imaging and fast STS measurements. Therefore, STM with conventional TIA [11] is not suitable for these measurements.

In this work, a TIA for STM with transimpedance gain of 10 GV/A and bandwidth larger than 200 kHz is proposed. Its transient response time is only 3.5 s. Its equivalent input noise voltage PSD is only 23 nV^2/Hz for f $>$ 100 kHz. Its equivalent input noise current PSD is only 4.7 fA^2/Hz at 10 kHz, and 88 fA^2/Hz at 100 kHz. For various low-conductivity materials, fast high-quality STM imaging measurements and fast high-energy-resolution STS measurements can be performed with this apparatus.

2. Design of STM-TIA and Its Electrical Performances

Figure 1 [Figure 1: see original paper] shows the circuit of the proposed TIA. It consists of several components: the pre-amplifier (Pre-Amp) shown in dashed box (A1) of Fig. 1, the post-amplifier (Post-Amp) shown in dashed box (A2), and the compensated feedback network shown in dashed box (B). The two-stage amplifier made of the Pre-Amp and Post-Amp is called the inverting amplifier (Inv-Amp). The Inv-Amp is connected with the feedback network to form the TIA. The circuit shown in dashed box (C) of Fig. 1 is called the signal source circuit. The TIA is connected with the signal source circuit to form the STM-TIA. The parameters of all components of the STM-TIA circuit are listed in Table 1 .

2.1. Design of Pre-Amp

The Pre-Amp is a noninverting amplifier consisting of a commercial operational amplifier (OPA) OPA657 [12] and two resistors Rf1 and R1 [13]. Cf1 is the parasitic capacitance of Rf1. The loop gain of the Pre-Amp is TL1(f) = aa1(f)\$ \$1(f), where aa1(f) is the open-loop voltage gain of OPA657 and \$ \$1(f) is the feedback factor of the noninverting amplifier; aa1(f) aa10/(1 + jf/f1b) and \$ $1(f) \approx R1/(R1 + Rf1 + 1/(j2\ fC1c) + 1/(j2\ fC1) + R1a).Here, aa10istheopen - loopvoltagegainofOPA657forf \rightarrow$\$0, f1b is the upper cut-off frequency of OPA657, C1c is the common-mode input capacitance of OPA657, C1 is its differential input capacitance, and R1a is the input resistance of OPA657.

TL1(f) is simulated by TINA-TI [14], and the simulation results are shown in Fig. 2 [Figure 2: see original paper]. At f1 where the modulus |TL1(f1)|dB = 0 dB, the argument (TL1(f1)) = −82°, meaning its phase margin of stability of the Pre-Amp is 98°. |TL1(f)|dB decreases monotonously with the increase of f, so its gain margin of stability is much greater than 6 dB. The voltage gain of the Pre-Amp is Av1(f) = (Rf1 + R1)/R1 × TL1(f)/(1 + TL1(f)). |Av1(f)|dB and (Av1(f)) are simulated by TINA-TI (see Fig. 3 [Figure 3: see original paper]), and Av1(f) is identical with the calculated value obtained with Eq. (1). Figure 3 shows that the upper cut-off frequency (3 dB drop of frequency) for Av1(f) is fh1 = 3.56 MHz, and (Av1(fh1)) = 43.6°. There is no "gain peaking" on the curve of |Av1(f)|dB, so the stability of the circuit is verified [13].

2.2. Design of Post-Amp

The Post-Amp is an inverting amplifier consisting of a commercial OPA THS4021 [15] and two resistors Rf2 and R2 [13]. Cf2 is the parasitic capacitance of Rf2. The loop gain of the Post-Amp is TL2(f) = aa2(f)\$ \$2(f), where aa2(f) is the open-loop voltage gain of THS4021 and \$ \$2(f) is the feedback factor of the inverting amplifier; aa2(f) aa20/(1 + jf/f2b) and \$ $2(f) \approx R2/(R2 + Rf2 + 1/(j2$ \$fC2) + R2a). Here, aa20 is the open-loop voltage gain of THS4021 for f → 0, f2b is the upper cut-off frequency of

THS4021, C2 is the input capacitance of THS4021, and R2a is the input resistance of THS4021.

TL2(f) is simulated by TINA-TI, and the simulation results are shown in Fig. 4 [Figure 4: see original paper]. At f2 where the modulus |TL2(f2)|dB = 0 dB, the argument (TL2(f2)) = −64°, meaning the phase margin of stability of the Post-Amp is 116°. |TL2(f)|dB decreases monotonously with the increase of f, so its gain margin of stability is much greater than 6 dB. The voltage gain of the Post-Amp is Av2(f) = −Rf2/R2 × TL2(f)/(1 + TL2(f)). |Av2(f)|dB and (Av2(f)) shown in Fig. 5 [Figure 5: see original paper] are simulated by TINA-TI; Av2(f) is identical with the calculated results obtained with Eq. (2). As shown in Fig. 5, the upper cut-off frequency for Av2(f) is fh2 = 2.12 MHz and (Av2(fh2)) = 134°. There is no "gain peaking" on the curve of |Av2(f)|dB, so the stability of the circuit is verified [13].

2.3. Inv-Amp Performances

The Pre-Amp and Post-Amp are cascaded to form the Inv-Amp. Both the Pre-Amp and Post-Amp are stable. The output resistance of the Pre-Amp is less than 150 Ω, and the input resistance of the Post-Amp is no less than 500 Ω. Therefore, the Inv-Amp is also stable. The open loop gain of the Inv-Amp is aA(f) = Av1(f) × Av2(f). Figure 6 [Figure 6: see original paper] shows the TINA-TI simulation results for the open loop gain of the Inv-Amp aA(f). The frequency corresponding to (aA) = 135° is fInv = 1.47 MHz. There is no "gain peaking" on the curve of |aA(f)|dB, so the stability of the circuit is verified.

2.4. Frequency Compensation of Feedback Loop

The Inv-Amp is connected with the feedback network to form the TIA. In order to increase the bandwidth of the TIA with the large feedback resistor RF, frequency compensation must be used in the feedback loop, since the effect from the parasitic capacitance CF of the feedback resistor RF cannot be ignored at high frequencies. In Ref. [16], a very simple design of a feedback circuit with bandwidth compensation has been proposed. As shown in dashed box (B) in Fig. 1, taking Cc kCF, where k = 10^4, adjust Rk RF/k to achieve RkCc = RFCF. Supposing the output voltage of the Inv-Amp is vO and the current flowing through the feedback loop to the input N of the Inv-Amp is iF, then ZF(f) = vO/iF = RF/(1 + j2πfRkCk). In this work, RF = 10 GΩ, CF 0.3 pF, Rk 1 MΩ and Ck 0.2 pF, so that fF = 1/(2πfRkCk) is about 800 kHz. In the (0, 100 kHz) range, |ZF(f)| RF/|1 + j2πfRkCk| and |ZF(f)| $\leq$ RF, so ZF(f) can be considered equal to RF in the (0, 100 kHz] range.

2.5. Circuit Stability of STM-TIA

The differential resistance of the tip-sample tunnel junction (TJ) in the signal source circuit is RJ, and its parallel capacitance is CJ. In this work, RJ $\geq$ 10 MΩ and CJ is usually of the order of fF [17]. BMS in Fig. 1 represents a voltage source, providing the applied DC bias and modulating signal voltage. CI is the capacitance of the cable connecting the tip to the TIA input N. If the cable is 1 cm long, the value of CI is about 0.5 pF.

The loop gain of the STM-TIA (see Fig. 1) is TL(f) = aA(f)β(f), where β(f) is the feedback factor of the TIA, and its reciprocal is 1/β(f) = 1 + ZF(f)/ZJ(f). Here ZF(f) = RF/(1 + j2πfRkCk) and ZJ(f) = RJ/(1 + j2πfRJC). C = CA + CI + CJ. CA is the input capacitance of the Inv-Amp, CA = C1c + C1 = 5.5 pF. RA is the input resistance of the Inv-Amp, and RA = R1a = 10 TΩ.

1/β(f) with CI = 0 for RJ = 10 MΩ is calculated from Eq. (6), and the results are shown in Fig. 6. The frequency where |1/β|dB is equal to |aA|dB is denoted as fc = 235 kHz, and fc « fInv; (1/β(f)) is always smaller than 90°. Therefore, as |TL(f)|dB > 0 dB, (TL(f)) > –135°. Figure 7 [Figure 7: see original paper] shows the simulated |TL(f)|dB and argument (TL(f)); as modulus |TL(f)|dB > 0 dB, (TL(f)) > –95°. Since |TL(f)|dB is monotone decreasing with the increase of f, the Inv-Amp circuit is stable [13]. The above conclusions for circuit stability are still valid for RJ > 10 MΩ with CI ≥ 0 pF.

2.6. Transimpedance Gain of TIA

Applying a sinusoidal current source iT in parallel with TJ, the generated output voltage of the STM-TIA is vO, and AiT(f) = vO/iT is called the transimpedance gain of the STM-TIA. In the (0, 300 kHz] range, AiT is approximately constant. Disconnecting the TIA from the signal source circuit and applying a sinusoidal current source iT into the input of the TIA, the output voltage is vO, and Ai(f) = vO/iT is called the transimpedance gain of the TIA. In the (0, 300 kHz] range, Ai is approximately constant.

Figure 8 [Figure 8: see original paper] shows the TINA-TI simulation results for AiT(f)/RF. There is no "gain peaking" on the curve of |AiT(f)/RF|dB, so the stability of the circuit is verified. As shown in Fig. 8, the upper cut-off frequency for AiT(f)/RF is fhT = 216 kHz and (AiT(fhT)/RF) = 127°. Therefore, for the proposed TIA, high gain and large bandwidth are achieved. According to Eq. (7), with the increase of the cable length between the tip and the TIA input (i.e., the increase of CI), the bandwidth for the STM-TIA is reduced. Therefore, this cable should be as short as possible. Comparing Eq. (7) and Eq. (8), as RJ ≥ 0.001RF and CI « CA, the upper cut-off frequency of the STM-TIA is approximately equal to fhT, so a high gain is achieved at large bandwidth for the proposed STM-TIA.

2.7. Transient Response of STM-TIA

Figure 9 [Figure 9: see original paper] shows TINA-TI simulation results for the transient response performance of the STM-TIA with RJ = 10 GΩ. The dashed curve is the input up-step signal voltage Vin, and the solid curve is the response output voltage Vo. The time taken from adding the input step signal voltage to the output response is called transient response time tr, and tr < 3.5 s as shown in Fig. 9. Furthermore, there are no "ringing" and "overshoot" characteristics on the output response curve, which also verifies the stability of the circuit.

3.1. Equivalent Input Voltage Noise and Equivalent Input Current Noise of TIA

For the TIA, its equivalent input noise voltage is denoted as eT, its equivalent input noise current is iT, and their harmonic components of frequency f are ET and IT, respectively. The circuit containing all noise sources for the TIA is shown in Fig. 10 [Figure 10: see original paper]. The equivalent input noise voltages of OPA657 for its two inputs are denoted as e1p and e1n, respectively, and their harmonic components of f are denoted as E1p and E1n, respectively. The equivalent input noise currents of OPA657 for its two inputs are denoted as i1p and i1n, respectively, and their harmonic components of f are denoted as I1p and I1n, respectively. The noise voltages of the resistor R1 and the feedback resistor Rf1 are denoted as e1 and ef1 respectively, and the harmonic components of f are denoted as E1 and Ef1, respectively. The equivalent input noise voltage of THS4021 is denoted as e2a, and its harmonic component of f is denoted as E2a. The equivalent input noise current of THS4021 is denoted as i2a, and its harmonic component of f is denoted as I2a. The noise voltages of the resistor R2 and the feedback resistor Rf2 are denoted as e2 and ef2, respectively, and their harmonic components of f are denoted as E2 and Ef2, respectively. For noise estimation, the compensated feedback network in dashed box (B) of Fig. 1 can be simplified to a resistor of RF. The noise voltage of RF is denoted as eF, and the harmonic components of f are denoted as EF.

As input N is grounded, the output noise of the TIA is denoted as eov. The equivalent input noise voltage of the noiseless circuit of the TIA eT as the input signal is added on input N, and the output noise is eove. For calculating eT, the equation is established on eov = eove. Therefore, by the nodal analysis method:

$$eT^2 = e1p^2 + (e1n^2 + ef1^2 + e1^2)/|Av1|^2 + (e2a^2 + ef2^2 + e2^2)/|Av1 \times Av2|^2 + eF^2/|Av1 \times Av2|^2$$

As input N is open-circuit, the output noise of the TIA is denoted as eoi. The equivalent input noise voltage of the noiseless circuit of the TIA iT as the input signal is added on input N, and the output noise is eoie. For calculating iT, the equations are established on eoi = eoie. Therefore, by nodal analysis method:

$$iT^2 = i1p^2 + (i1n^2 + i2a^2)/|Av1|^2 + iF^2$$

By Wiener–Khintchine theorem, the power spectral densities can be obtained [18, 19]. The small quantities in the Pre-Amp, such as the thermal noise of R1 and Rf1 can be ignored. All noises produced in the Post-Amp can also be ignored [19]. Therefore:

$$eT^2 \quad e1p^2 + eF^2/|Av1 \times Av2|^2 \; iT^2 \quad i1p^2 + iF^2$$

Here, $e1p^2 = 4kTR1c + eC^2$, where eC^2 is the thermal noise of R1c, and $iF^2 =$

4kT/RF is the thermal noise of RF, where k is Boltzmann's constant and T is the absolute temperature. eC^2 is only 1.6×10^{-20} V^2/Hz at T = 300 K. $e1p^2$ is the announced equivalent input noise voltage PSD of OPA657, which is 40 nV^2/Hz at f = 1 kHz and 23 nV^2/Hz for f $\geq$ 100 kHz [12]; $i1p^2$ is the announced equivalent input noise current PSD of OPA657, which is only 1.7 fA^2/Hz [12]. Therefore, eT^2 is 40 nV^2/Hz and iT^2 is 3.2 fA^2/Hz at 10 kHz, and eT^2 is 23 nV^2/Hz and iT^2 is 88 fA^2/Hz at 100 kHz [12]. For FEMTO DE-DLPCA-200 with the gain of 1 GV/A, iT^2 is 18.5 fA^2/Hz at 1 kHz. For FEMTO DE-LCA-200-10G with the gain of 10 GV/A, eT^2 is 16 nV^2/Hz and iT^2 is 3.2 fA^2/Hz at 1 kHz, eT^2 is 3.2 fA^2/Hz at 10 kHz, and eT^2 is 40 nV^2/Hz and iT^2 is estimated as 81 nV^2/Hz at 1 kHz and larger than 2.3 fA^2/Hz at 1 kHz. The proposed TIA has a bandwidth higher than 100 kHz, and the noises of the proposed TIA are almost the same as the commercial ones with the same transimpedance gain at 1 kHz.

3.2. Inherent Noise of STM-TIA

The inherent noise of STM-TIA can be described by a virtual noise current source iin in parallel with TJ, whose noise current PSD is called the equivalent input noise current PSD of STM-TIA, denoted as iin^2 [19]. According to the noise model of the amplifier:

$$iin^2 = iT^2 + (2\pi fCIJ)^2 eT^2$$

where CIJ = CI + CJ. Putting Eqs. (11), (12), and (13) into Eq. (14), iin^2 can be calculated.

4. DC Tunnelling Current Measurement Accuracy with the Proposed TIA in STM

The Inv-Amp in the TIA has the input offset voltage VOS, as shown in Fig. 11 [Figure 11: see original paper]. It is easy to prove that VOS Voff, where Voff is the input offset voltage of OPA657. In Fig. 11, as switch K is open, for the TIA, VOS will cause a non-zero DC voltage Vtia at the output, and Vtia can be measured. It is also easy to prove that Vtia −VOS [17]. As switch K is closed, the STM-TIA is formed. For the STM-TIA, applying the input voltage Vi, the potential at input N is VN and at the output O the voltage is Vo which can be measured. The DC tunneling current is I = (Vi − VN)/R, where R is the TJ DC resistance corresponding to Vi. For the STM-TIA, I = (VN − Vo)/RF and Vo = aA0(VN + VOS).

IS = (Vtia − Vo)/RF is the approximate value of the DC tunneling current I. For the STM-TIA, since aA0 −90000, the relative error |I − IS|/|I| for the DC tunneling current I is less than 12 ppm. The bias drop on TJ is V = Vi − VN = Vi − Vtia − (Vo − Vtia)/(1 − aA0); as aA0 $\rightarrow -\infty$, Vi − Vtia can be considered as the approximate value of V and the relative error is less than 1/89.

5. Conclusions

In this work, a large-bandwidth high-gain low-noise transimpedance amplifier (TIA) for STM is proposed to meet the needs of fast high-quality STM imaging measurements and fast high-energy-resolution STS measurements for low-conductivity materials. For the noises of the proposed TIA, as the calculations show, eT^2 is 3.2 fA^2/Hz at 1 kHz, eT^2 is 40 nV^2/Hz and iT^2 is 3.2 fA^2/Hz at 10 kHz, and eT^2 is 23 nV^2/Hz and iT^2 is 88 fA^2/Hz at 100 kHz.

According to the simulation results, the proposed TIA has a bandwidth of over 200 kHz, and the noises of the proposed TIA are lower than the commercial ones with the same transimpedance gain at 1 kHz. This apparatus is very suitable to be equipped in STM for research of complex organic systems and quality tests of wide bandgap semiconductors in the semiconductor industry.

Conflict of Interest

The authors declare that they have no conflicts of interest.

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Note: Figure translations are in progress. See original paper for figures.