A gain and bandwidth enhanced transimpedance preamplifier for Fourier-transform ion cyclotron resonance mass spectrometry

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The nature of the ion signal from a 12-T Fourier-transform ion cyclotron resonance mass spectrometer and the electronic noise were studied to further understand the electronic detection limit. At minimal cost, a new transimpedance preamplifier was designed, computer simulated, built, and tested. The preamplifier design pushes the electronic signal-to-noise performance at room temperature to the limit, because of its enhanced tolerance of the capacitance of the detection device, lower intrinsic noise, and larger flat mid-band gain (input current noise spectral density of around 1 pA/ $\sqrt{\text{Hz}}$ when the transimpedance is about 85 dB Ω). The designed preamplifier has a bandwidth of ~ 3 kHz to 10 MHz, which corresponds to the mass-to-charge ratio, m/z , of approximately 18 to 61 k at 12 T. The transimpedance and the bandwidth can be easily adjusted by changing the value of passive components. The feedback limitation of the circuit is discussed. With the maximum possible transimpedance of 5.3 M Ω when using an 0402 surface mount resistor, the preamplifier was estimated to be able to detect ~ 110 charges in a single scan. © 2011 American Institute of Physics. [doi:10.1063/1.3660778]

I. INTRODUCTION

Mass spectrometry (MS) is generally recognized as one of the essential analytical tools in modern chemical and biological research. Due to its flexibility, and because it provides the highest resolving power and mass accuracy currently possible with mass spectra, Fourier-transform ion cyclotron resonance (FT-ICR) MS (Refs. 1 and 2) has become the instrument of choice for proteomics,³ biological imaging,^{4–6} and other research involving complex mixtures⁷ or macromolecules.^{8–10}

Positioned within an ultra high vacuum chamber ($<10^{-9}$ mbar), an ion cyclotron resonance (ICR) cell is the main analyzing component of FT-ICR MS. The ICR cell detects ion motion in the magnetic field by electrostatic induction. The induced current is Fourier transformed into the frequency domain, and such ICR frequencies are used to determine the masses of the ions. However, ICR frequencies are a function not only of the magnetic force, but also of the electrostatic trapping field, which is ideally hyperbolic. In reality, with the ICR cell typically used in commercial instruments, the electrostatic trapping field is a good approximation of a hyperbolic field only near the center of the cell. Thus, with the imperfect electric fields typically used, ICR frequencies are modulated as the ions orbit in the cell.

In the 90s, different ICR cell designs were conceived to reduce the perturbation caused by the imperfect electrostatic trapping potential to improve performance.¹¹ In recent years, cell design research remains active; a few modern designs have been reported to optimize the in-cell electric field^{12–17} to push the FT-ICR MS performance close to its theoretical limit. Ion packets experience an imperfect, non-hyperbolic electric field and axial magnetic field inhomogeneity when

rotating within an ICR cell at a “higher” orbital radius. Furthermore, the space-charge forces arising from the Coulombic interaction between ions affect both frequency and stability (peak width) of the peaks in the spectra, thus limiting the mass accuracy. These effects can also result in a rapid loss of coherence in the transient which limits the resolving power.¹⁸ To avoid such phenomena, fewer ions are sent into the cell and excited to a “lower” rotating orbit for detection. As a consequence, at the detection plates of the cell, the induced current due to the cyclotron motion of the ions is smaller, resulting in a weaker signal being presented to the front-end electronics.

A preamplifier, which is usually mounted inside the ultra high vacuum chamber as closely as possible to the ICR cell, is the key front-end electronic component in FT-ICR MS. The preamplifier detects the image current¹⁹ induced by the excited ion packets rotating at a coherent orbit in the cell. These currents are generally on the scale of tens of femtoamps to a few hundred picoamps within the frequency range of 10 kHz to 1 MHz. The amplified signal will be further processed in a signal processing chain involving second stage amplifiers, band pass filters, and an analog-to-digital converter (ADC), in air outside the vacuum system. The digitized signal then can be analyzed by a computer. The total gain within such a signal chain should be carefully designed so that the amplified signal fits in the dynamic range of the ADC. Insufficient gain sacrifices the detection sensitivity, whereas excessive gain causes saturation of the ADC, which results in the generation of artifact peaks by the Fourier transform.²⁰

A. Signal and gain

The root mean square (RMS) signal current¹⁹ from an ICR cell can be given by Eq. (1),

$$I_s(r.m.s) = \frac{Nq^2B}{\sqrt{2m}} \times \frac{r}{d}, \quad (1)$$

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where N is the number of excited ions, q is the ionic charge, B is the magnetic field strength, m is the ion mass, r is the ion rotation orbital radius, and d is the cell diameter (spacing). For a singly-charged, 1000-Dalton ion rotating at the orbit radius of $1/4$ the cell diameter in a 12-T magnetic field, namely, $N = 1$, $B = 12$ T, $m = 1000$, and $r/d = 1/4$, Eq. (1) calculates a RMS current signal of about 33 fA/charge.

Considering the worst-case scenario of single-charge detection, 1-bit change at the least significant bit of the ADC output is needed to detect this single ion. If the ADC has a resolution of 16 bits with an input range of ± 1 V, the minimum total gain, which is the transimpedance A_T in this case, of the full amplifier must be

$$A_T = \frac{2 \times \frac{1}{2^{16}}}{3.27 \times 10^{-14}} = 9.3 \times 10^8 \text{ (V/A)}. \quad (2)$$

Note this calculation assumes a noiseless signal, but upon digitization, noise allows detection of periodic signals less than 1 bit change.²¹ Instead of having only one amplifier, such a large gain can be achieved by introducing several amplifying stages in series, including a preamplifier at the first stage, and several instrumentation amplifiers at the later stages. Figure 1 shows such a signal chain with three amplifying stages. The first stage amplifier provides transimpedance A_T for the input current signal s and the noise current n_0 coming from the ICR cell. The transimpedance preamplifier converts the input current into a voltage output for the following second and third stage voltage amplifiers. After the preamplifier, the signal is further amplified by voltage gains G_2 and G_3 at the second and third stage, respectively.

However, each amplifying stage adds its own intrinsic noise onto the signal. In Fig. 1 such noises are referred back to the inputs of the first, second, and third stage amplifiers and are represented by n_1 , n_2 , and n_3 , respectively. As a result, the output contains not only the amplified signal, S , where

$$S = s A_T G_2 G_3, \quad (3)$$

but also the noises, N , where

$$N = (n_0 + n_1) A_T G_2 G_3 + n_2 G_2 G_3 + n_3 G_3. \quad (4)$$

Here, the gains G_2 and G_3 are generally greater than 1. With large transimpedance A_T , where $A_T \gg n_2$ and $A_T \gg n_3$, the signal-to-noise ratio, S/N , at the output can be obtained by dividing Eq. (3) by Eq. (4), and can be approximated as

$$S/N = \frac{s}{n_0 + n_1 + \frac{n_2}{A_T} + \frac{n_3}{A_T G_2}} \simeq \frac{s}{n_0 + n_1}. \quad (5)$$

To conclude, by having significant gain at the first amplifying stage, the contribution of the electronic noise is limited to the first stage, namely, the intrinsic noise of the preamplifier and detection components. To electronically improve the

FIG. 1. (Color online) Gain and noise distribution in a three stage amplifier system.

signal-to-noise performance, it is required that the first stage amplifier has not only as much gain as possible within the bandwidth of interest, but also minimal noise, n_1 , so that the preamplifier design is crucial for best performance.

B. Transimpedance and bandwidth

Fundamentally, the signal to be detected is a current due to the motion of ions. Thus, a transimpedance amplifier to convert the current input to a voltage output for further amplification is needed. A transimpedance amplifier is a widely used current-voltage converting solution for many applications such as optical communication.^{22–24} The negative-feedback transimpedance technique has the advantages of reducing the effective input load capacitance to the amplifier²⁵ and extending the bandwidth²⁶ by a factor equal to the open-loop gain of the amplifier. Therefore, it appears to be ideal for detecting ion signals from an ICR cell, which in general acts like a current source¹⁹ with capacitance (10–100 pF) depending on cabling and the size of the cell. A typical FT-ICR preamplifier can be modeled from a circuit schematic such as Fig. 2(a), where C is the total source capacitance from the

FIG. 2. (Color online) Two types of preamplifier system schemes. The ideal current source I represents the signal from the ICR cell. The capacitor C is the combination of the parasitic capacitance of the cell and cabling. (a) A unity gain operational amplifier with an input resistor R to convert current signal into voltage. (b) A transimpedance amplifier with a feedback resistor R .

ideal alternating current (AC) current source I , which represents the current signal from the ICR cell. The operational amplifier (Op Amp) with open-loop gain G has an input resistor R , which is responsible for transforming the input current into voltage. The current-to-voltage transfer function $H(\omega)$ can be derived as

$$H(\omega) = \frac{R}{1 + j\omega(RC)}, \quad (6)$$

and the 3-dB bandwidth ω_{3dB} is given by

$$\omega_{3dB} = \frac{1}{RC}. \quad (7)$$

If the configuration of such an amplifier is replaced by a transimpedance amplifier with negative feedback, the circuit becomes what is shown in Fig. 2(b).²⁷ The transfer function becomes

$$H(\omega) = \frac{R}{1 + j\omega(RC/G)}, \quad (8)$$

and the 3-dB frequency will then be extended to

$$\omega_{3dB} = \frac{G}{RC}. \quad (9)$$

With such bandwidth increment, the gain-bandwidth product is further pushed to higher frequency range and therefore higher gain can be introduced to the amplifier system with the same 3-dB bandwidth. Equivalently, the tolerance to the parasitic capacitance of the cell and cabling is stronger. Therefore, the preamplifier can be located further away from the cell to isolate it from any possible electrical perturbation induced by the high magnetic field.

C. Noise

To best isolate the preamplifier system from environmental noise from multiple sources, the grounding and the shielding of the circuitry have to be carefully considered. Even with careful protection, some noise sources, such as the RF noise from the switching power supply,²⁸ unwanted coupling from wiring, and Johnson-Nyquist noise (thermal noise), which is the electronic noise caused by electron's thermal agitation inside conductors,^{29,30} cannot be completely prevented. Thermal noise intrinsically exists in any electrical conductive material, and can be modeled as a current source. The RMS noise current of a given resistor with resistance R is given by

$$i_n(f) = \sqrt{\frac{4k_B T \Delta f}{R}}, \quad (10)$$

where k_B is Boltzmann's constant, T is the absolute temperature in kelvins, and Δf is the bandwidth over which the noise is measured. In such a resistor, the noise power

$$P_n = 4k_B T \Delta f \quad (11)$$

is generated. The only way to reduce such noise power is to lower the temperature. In 2002, O'Connor proposed an idea of holding the vacuum system of the FT-ICR MS inside a helium cooled cold bore at the temperature of 4.2 K.³¹ Significantly reducing the thermal noise, the reported preamplifier design³² showed about 20 times improvement in signal-to-noise ratio.

However, the high cost of maintaining liquid helium to preserve the 4-K environment prevents such a system from being popularized. In 2007, Mathur *et al.* reported a room temperature differential preamplifier,³³ which was further based on Jefferts and Walls' design,³⁴ with similar configuration as shown in Fig. 2(a). In Mathur *et al.*'s design with a total gain of about 3500 (around 25 at the first stage), the signal current was converted into voltage by a large input resistor, which acted as a major noise source at the input node, before being further processed by the input transistor of the first stage preamplifier. Here, this work presents a new circuit design to convert the input current directly to voltage at the output, with negative feedback stabilization, lower intrinsic noise, higher and yet flat gain, and broader bandwidth with better tolerance to the intrinsic capacitance of the ICR cell and cabling, to further improve both the signal sensitivity and the noise performance for FT-ICR MS.

II. METHODS

A. Circuit design

The basic design uses a junction field-effect transistor (JFET) front-end (first stage) and an Op Amp main stage for gain. The schematic of the preamplifier circuit is shown in Fig. 3(a). The overall feedback is controlled by a single resistor (R_1), which determines the transimpedance of the preamplifier. The same resistor in the feedback loop also takes the responsibility of biasing the detection plates of the ICR cell, in this case, to a virtual ground potential, which is necessary to preserve ion trajectories in the cell. In order to limit the intrinsic noise from the passive components, low noise surface mount ceramic capacitors and surface mount metal film chip resistors were selected to define the biasing conditions of the preamplifier system and to provide the transimpedance in the feedback loop. The circuit was built on a carefully designed single-layer printed circuit board (PCB), as shown in Fig. 3(b), to limit the noise coupling and parasitic capacitance.

The main stage is formed by an ultralow distortion Op Amp AD8099 from Analog Devices, Inc. (Norwood, Massachusetts),³⁵ biasing resistors R_3 - R_5 , bias adjustment resistors R_7 - R_9 , feedback resistor R_6 and capacitor C_6 , direct current (DC) blocking capacitor C_2 , and bypass capacitors C_3 - C_5 and C_7 - C_{10} . AD8099's ultralow noise input voltage spectral density of $0.95 \text{ nV}/\sqrt{\text{Hz}}$ (specified at 100 kHz), ultralow distortion (-92 dBc at 10 MHz), and wide bandwidth (700 MHz at the gain of $+2$) are specifically utilized in this preamplifier system. Pin 5 of the Op Amp is not connected, as suggested by the data sheet³⁵ when the gain is set to around $+20$. Pin 1 and pin 6 are internally connected inside the surface mount package to increase the routing flexibility on a PCB. Resistors R_4 and R_6 set the AC signal gain of this main stage to 19, while the 3-dB bandwidth f_{3dB} (high frequency cut-off), which can be calculated by Eq. (12), is limited by the capacitor C_6 together with the feedback resistor R_6 to 10 MHz. Bypass capacitors are placed as close as possible to the AD8099 chip, as suggested by the data sheet,³⁵ for optimum distortion and power supply

(a)

(b)

FIG. 3. (Color online) Preamplifier circuit. (a) Schematic of the transimpedance preamplifier. (b) Single layer printed circuit board (sized $\sim 65 \times 60$ mm) with components fully populated.

rejection performance.

$$f_{3dB} = \frac{1}{2\pi(R6)(C6)}. \quad (12)$$

The DC biasing conditions of both AD8099 inputs are balanced by R3 together with R4 and R5, and can be trimmed by the trimmer resistor R8 to maintain the output DC level at 0 V. In this application, pin 8 (the disable pin) is connected to +6V DC power rail so that the Op Amp remains on whenever the power to the preamplifier system is on. Potentially, connecting this disable pin with a control signal may be a good solution when isolation between the ICR cell and the amplifying circuitry is necessary.

The input current biasing requirement (1 μ A in maximum, as specified on the data sheet³⁵) of AD8099 was fulfilled by introducing a front-end buffer. In order to limit the possible intrinsic noise added by this first stage buffer into the preamplifier system, the ultralow noise (noise input voltage spectral density of 0.8 nV/ $\sqrt{\text{Hz}}$ at 100 kHz) JFET BF862 from NXP Semiconductors N.V. (Eindhoven, The Netherlands),³⁶ was chosen. The BF862 front-end can be configured either as a source follower or a common source amplifier. A common source amplifier provides more voltage gain to increase the open-loop gain (G in Eq. (9)) to extend the 3-dB bandwidth, but also results in an unwanted 180-degree phase shift before unity gain, causing the preamplifier to oscillate. Therefore, this BF862 first stage was designed to be a source follower with an 1-k Ω source resistor (R2 in Fig. 3(a)).

B. Bandwidth and computer simulation

The ICR frequency ω of the excited ions is roughly³⁷

$$\omega = 2\pi f = \frac{qB}{m}. \quad (13)$$

The designed bandwidth was limited to the range of 1 kHz to 10 MHz, which corresponds to an output mass-to-charge ratio, m/z , of roughly 18 to 180 k in a 12-T FT-ICR MS system. Use of a bandpass filter configuration minimizes the artifacts, such as the signal aliasing caused by high frequency signal fold-over, and the offset in the signal,²⁰ in FT-ICR mass spectra. In this design, the low frequency cutoff is defined by the capacitor C2 and the effective resistance (~ 1 k Ω) in series with it, whereas the high frequency 3-dB point is defined by the feedback resistor R1 together with the capacitor C1. Both frequency poles can be estimated by Eq. (12). It is to be noted that the resistance of R1 was chosen to be large due to required large transimpedance. This limits the capacitance of C1 to a range that may be similar to the parasitic capacitance contributed by the resistor R1. In reality, when R1 is large enough, the feedback capacitor C1 may not be necessary to meet the required high frequency cutoff. For example, a ~ 200 k Ω 0805 surface mount package resistor, which can have the parasitic capacitance of about 80 fF, will set the 3-dB point at around 10 MHz. To test the designed bandwidth and noise performance, computer simulation was carried out using the computer simulation tool Multisim v11.0 from National Instruments (Austin, Texas).

C. Printed circuit board design

A single-layer PCB was designed using the computer-aided design software, Altium Designer Build 8.4 (Service Pack 4) from Altium Limited (Sydney, Australia). On the PCB, all of the components were located inside a ground ring to shield them from environmental noise. The board was manufactured in-house using the PCB prototyping plotter ProtoMat S62 from LKPF Laser & Electronics AG (Garbsen, Germany). The populated PCB (sized $\sim 65 \times 60$ mm) is shown in Fig. 3(b).

D. Experiment

The measurements were carried out by using the PXI cards, PXI-5122 oscilloscope, PXI-5421 arbitrary waveform generator, and PXI-6733 analog output card along with PXI-8336 control card and PXI-1042 chassis from National Instruments (Austin, Texas). The software LabVIEW 2009 (Service Pack 1) from National Instruments was used to control and to collect data from the PXI system.

The AC analyses were carried out with 100 MS/s sampling rate. The preamplifier was powered by the DC power supply TTI EL301R from Thurlby Thandar Instruments Ltd. (Huntingdon, UK). 20 points are measured and averaged to obtain both the input current going into the node “Iin” and the output voltage measured at node “Vout” in Fig. 3(a). The peak-to-peak amplitude of the testing input sinusoidal current was 0.12 mA. The standard deviations of the data measured at input were between 0.18 and 0.56 μ A, and, at output, between 0.36 and 3.3 mV.

Noise performance was measured at the output by the spectrum analyzer IFR A-7550 from Aeroflex Inc. (Plainview, New York) when the input was floating and a DC blocking capacitor was connected between the output and the spectrum analyzer. The noise power was measured in dBm. The noise power in dBm, $P_{n(\text{dBm})}$, can be calculated from the noise power in watts, $P_{n(W)}$, by

$$P_{n(\text{dBm})} = 10 \times \log \left(\frac{P_{n(W)}}{1 \times 10^{-3}} \right). \quad (14)$$

Alternatively, the measured noise power in dBm can be converted back to watts, and can be referred to the output voltage V_{out} by

$$P_{n(W)} = 10^{\left(\frac{P_{n(\text{dBm})} - 30}{10}\right)} = \frac{(V_{out})^2}{R_{load}}, \quad (15)$$

where R_{load} is the input resistance of the spectrum analyzer. The transimpedance, A_T , of the preamplifier, and the resolution bandwidth, (Res), of the spectrum analyzer, have to be considered to further correlate the output voltage V_{out} of the preamplifier back to the input current spectral density $i_n(f)$, where

$$i_n = \sqrt{\frac{P_{n(W)} \times R_{load}}{(Res) \times (A_T)^2}} \quad (\text{pA}/\sqrt{\text{Hz}}). \quad (16)$$

By using Eqs. (15) and (16), the input current noise spectral density $i_n(f)$ (in $\text{pA}/\sqrt{\text{Hz}}$) data were calculated from the

measured noise power $P_{n(dBm)}$ (in dBm) and the transimpedance A_T data collected in Fig. 5(b).

III. RESULTS & DISCUSSIONS

The AC analysis simulations were performed to show the transimpedance with different feedback resistance and capacitance values. Two feedback resistors, 180 Ω and 18 k Ω , were chosen to show the gain variance. Figure 4(a) shows the results when feedback capacitor C1 was not connected to the system. In the simulation program, the parasitic capacitance of the 0805 sized surface mount feedback resistor R1 was set to 85 fF. The AC analysis with designed 0.85 pF feedback capacitor is shown in Fig. 4(b). As expected, the low frequency cut-off is independent of the feedback impedance and is slightly below 1 kHz for both cases. When the feedback capacitance of C1 is fixed at 0.85 pF, from Eq. (12) the 3-dB bandwidth occurs at around 10 MHz and 1.0 GHz when R1 is 18 k Ω and 180 Ω , respectively. The simulation results supported such estimations.

FIG. 4. (Color online) Computer simulations of the preamplifier transimpedance. (a) AC analysis without the feedback capacitor C1 in Fig. 3(a). (b) AC analysis with 0.85 pF feedback capacitor.

FIG. 5. (Color online) AC analysis from the measured data. (a) Voltage gain of the main (AD8099) stage. (b) Transimpedance of the entire preamplifier system in three feedback conditions.

In Fig. 4(a), the peakings around 20 MHz and 200 MHz of the 18-k Ω and 180- Ω curves, respectively, are results of the phase shift caused by the LC resonance of the Op Amp output impedance, and can be eliminated by introducing the 3-dB poles before the peaks. As shown by the 18-k Ω curve in Fig. 4(b), the 0.85 pF feedback capacitor and the 18 k Ω feedback resistor cause a pole at around 10 MHz, thus the transimpedance drops before the peak at $\sim$ 20 MHz. For the 180- Ω curve, with the same feedback capacitance of 0.85 pF to have the 3-dB pole located at about 1.0 GHz, the peaking at $\sim$ 200 MHz remains.

The voltage gain of the main stage and the overall transimpedance performance were measured, and the results are shown in Fig. 5. The voltage gain of the Op Amp AD8099 was tested when the first stage (BF862 and its source resistor R2 in Fig. 3(a)) was not mounted to the PCB, to ensure that the performance of this main stage matches the designed conditions of providing +19 voltage gain between 1 kHz and 10 MHz. The testing signal was fed into the system via the DC blocking capacitor (C2), and the output was measured directly from the output of AD8099. The result showed an agreement

with the designed voltage gain of +19 (about 25 dB), and the designed 3-dB bandwidth of 10 MHz. Figure 5(b) shows the transimpedance of the entire preamplifier system in three conditions, when different feedback resistors and capacitor were soldered onto the system: (a) only a 180 Ω resistor, (b) only an 18 k Ω resistor, and (c) an 18 k Ω resistor together with a 0.8 pF capacitor.

The feedback resistor sets the transimpedance of the preamplifier system. Nevertheless, the parasitic capacitance within the feedback resistor itself limits the maximum value of the feedback resistance. With the 0805 surface mount package used in this design, the parasitic capacitance of a resistor can be as large as 80 fF. According to Eq. (12), with the designed 10 MHz bandwidth, the resistance is calculated to be about 200 k Ω . When a smaller package is used, such parasitic capacitance can be limited to a lower value. For instance, a 0402 surface mount resistor has a typical parasitic capacitance value of around 30 fF. As a result, by introducing a resistor with 0402 package, the transimpedance of this system can be pushed to 530 k Ω for 10-MHz bandwidth, or to 5.3 M Ω for 1.0 MHz.

The noise performance was tested when the input node was floating. The total output power was recorded, converted into input current noise spectral density by using Eqs. (15) and (16), and then plotted in Fig. 6. The measured noise performance of this preamplifier system agrees with the general characteristic of an Op Amp, where the noise spectral density curve is the combination of low frequency *1/f noise* and also *white noise*.

One thing to notice is that the measured noise power is independent of the feedback resistance, namely, the transimpedance of this preamplifier. If a 5.3 M Ω 0402 surface mount resistor is used as the feedback while narrowing the bandwidth to 1 MHz, the equivalent input current noise spectral density will become 3.7 fA/ $\sqrt{\text{Hz}}$, rather than the reported ~ 1.1 pA/ $\sqrt{\text{Hz}}$ (shown in Fig. 6) at 1 MHz. Meanwhile, the bandwidth can be easily adjusted by changing the capacitance of C1 (for high frequency cut-off) and C2 (for low frequency cut-off) in Fig. 3(a). When the bandwidth is narrowed to

FIG. 6. (Color online) Measured input current noise spectral density with 18 k Ω transimpedance.

1 MHz (commonly used m/z region of having a low mass cut-off at ~ 180 m/z for a 12-T FT-ICR MS system), with a 5.3 M Ω feedback resistor the electronic noise generated by this preamplifier referred back to the input current will become roughly 3.7 pA.

The detection bandwidth can be traded for sensitivity by narrowing the detection window, which can be done by introducing a filtering technique or by changing the preamplifier bandwidth. At threshold, the signal current has to be at least equal to the noise level in order to be detected, which means from Eq. (1) around 110 singly-charged, 1000-Dalton ions (assuming the rotating orbit radius of 1/4 the cell diameter, in a 12-T magnetic field) in the ICR cell for detection. In lieu of detecting the 1-MHz bandwidth in one detection using a 1-MHz detection window, a narrower 100-kHz window can be introduced to finish the same task in 10 detections. In which case, with a 5.3 M Ω feedback resistor ~ 35 charges ($= 3.7 \times 10^{-15} \times \sqrt{100 \times 10^3} \div 33 \times 10^{-15}$) can be detected in one scan. In reality, techniques such as multiple acquisition can be introduced to allow detection of a signal below the noise threshold.

IV. CONCLUSION

As the key front-end electronic component, the preamplifier plays a critical role in pushing the unmatched FT-ICR MS performance to the limit. Following the studies of the ICR signal model from Comisarow,¹⁹ the gain and noise distribution in a multistage amplifier system, and the electronic noise, this work reports a preamplifier with transimpedance configuration to have not only stronger tolerance to the intrinsic capacitance of the cell, but also higher designed gain as a result of the bandwidth increase.

At minimal cost, the improved preamplifier has excellent flexibility in adjustment of its transimpedance and bandwidth, while maintaining flat mid-band gain at the frequency of interest. With the chosen 18 k Ω feedback resistor, 0.8 pF feedback capacitor, and 220 nF DC blocking capacitor, the transimpedance of the preamplifier is around 85 dB Ω between ~ 3 kHz and 10 MHz; the input current spectral density is about 1.1 pA/ $\sqrt{\text{Hz}}$ at 1 MHz. With a 5.3 M Ω feedback resistor, this preamplifier has the evaluated capability of detecting ~ 110 singly-charged ions, or ~ 35 charges when the bandwidth is restricted to 100 kHz. In the near future, this preamplifier will be further mounted onto a FT-ICR MS system to test the performance.

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