A low noise single-transistor transimpedance preamplifier for Fourier-transform mass spectrometry using a T feedback network

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A novel single-transistor transimpedance preamplifier has been introduced for improving performance in Fourier-transform ion cyclotron resonance (FT-ICR) mass spectrometry. A low noise junction field-effect transistor (JFET), BF862, is used as the main amplification stage of this transimpedance preamplifier, and a T-shaped feedback network is introduced as both the feedback and the gate biasing solutions. The T feedback network has been studied using an operational amplifier (Op Amp), AD8099. Such a feedback system allows ~ 100 -fold less feedback resistance at a given transimpedance, hence preserving bandwidth, which is beneficial to applications demanding high gain. The single-transistor preamplifier yields a tested transimpedance of $\sim 10^4 \Omega$ (80 dB Ω) in the frequency range between 1 kHz and 1 MHz (mass-to-charge ratio, m/z , of around 180–180k for a 12-T FT-ICR system), with a low power consumption of ~ 6 mW, which implies that this preamplifier is well suited to a 12-T FT-ICR mass spectrometer. In trading noise performance for higher transimpedance, an alternative preamplifier design, an AD8099 preamplifier with the T feedback network, has also been studied with a capability of $\sim 10^6 \Omega$ (120 dB Ω) transimpedance in the same frequency range. The resistive components in the T feedback network reported here can be replaced by complex impedances, which allows adaptation of this feedback system to other frequency, transimpedance, and noise characteristics for applications not only in other mass spectrometers, such as Orbitrap, time-of-flight (TOF), and ion trap systems, but also in other charge/current detecting systems such as spectroscopy systems, microscopy systems, optical communication systems, or charge-coupled devices (CCDs). © 2012 American Institute of Physics. [<http://dx.doi.org/10.1063/1.4751851>]

I. INTRODUCTION

Fourier-transform mass spectrometry (FTMS) is an essential analytical instrument for detailed analysis of chemical compounds. As the demand grows for the highest mass accuracy and the highest mass spectral resolution in the areas of proteomics,¹ petroleum,^{2,3} biological imaging,⁴ and others, where macromolecule/complex mixture analysis is involved, Fourier-transform ion cyclotron resonance (FT-ICR) MS⁵ remains the optimum analytical tool of choice.

Recently, efforts have been made toward the enhancement of the FT-ICR MS performance. In terms of sensitivity, the front-end electronics, especially the preamplifier, in a FT-ICR system plays a crucial role. It was discussed that in a signal processing chain, the noise performance can be dominated by the noise from the signal source and from the first stage “signal processor,” the preamplifier, if the preamplifier is designed with significant gain.⁶ The potential improvement scope for such preamplifiers remains significant, as the conventional preamplifiers, which use 25-year-old designs⁷ updated with modern components,^{8,9} provide a signal-to-noise limitation that requires at least 30–100 ions to achieve a signal-to-noise of 3.^{10,11} To obtain the best system performance, the design goal for a preamplifier is generally to boost its gain to the maximum and to limit its noise to the minimum. New preamplifier designs may allow single charge detection,

which would maximize the potential dynamic range of FTMS instruments.

Similar to the photodiodes in optical communication systems, the signal analyzer of a FT-ICR mass spectrometer, an ICR cell,¹² is a capacitive device. Electrostatically induced image currents are expected at the input of the preamplifier. The parasitic capacitance from the cell can vary from around 10 pF to over 100 pF,¹³ depending on the cell dimensions, feedthroughs, and cabling. A high capacitance at the preamplifier input limits the bandwidth, causing potential signal intensity loss. As a result, a transimpedance amplifier, with its ability of effective input capacitance reduction^{6,14} over an existing voltage amplifier (VA) design,⁸ can potentially provide an ideal preamplifier solution for many mass spectrometry systems, including FT-ICR MS.

II. THEORY

A. Transimpedance amplifier transfer function

A basic transimpedance amplifier is constructed out of an operational amplifier (Op Amp) with a feedback resistor, as shown in Fig. 1(a). An Op Amp senses the voltage difference between its non-inverting input (V_+) and inverting input (V_-), and amplifies it with its open-loop gain (A). Consequently, the voltage at output becomes

$$V_{out} = A \times (V_+ - V_-). \quad (1)$$

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FIG. 1. Transimpedance amplifiers with different feedback arrangements (I_{in} indicates an ideal current signal source). (a) Transimpedance amplifier with a single-resistor feedback. (b) Transimpedance amplifier with a T-shaped feedback network.

An ideal Op Amp has a few particular characteristics, including infinite input impedance and infinite open-loop gain. In reality, the open-loop gain, A , is a finite large value of typically larger than 10^4 . When analyzing the Op Amp circuit, it is common to assume that the input impedance is large so that the current flowing into either the inverting or non-inverting input is negligible. With the hypotheses mentioned above, the transfer function, or the transimpedance, $T_{single-resistor}$ of a negative-feedback Op Amp circuit (as shown in Fig. 1(a)) can be written as

$$T_{single-resistor} = \frac{V_{out}}{I_{in}} = \frac{-R_f}{1 + \frac{1}{A}} \simeq -R_f. \quad (2)$$

Thus, the closed-loop gain of this transimpedance amplifier system is independent of the Op Amp characteristics (its open-loop gain A), but is a function of the characteristics of the passive components (the feedback resistance R_f in this case) used to “close” the loop.

The single feedback resistor can be replaced by a three-resistor T-shaped feedback network,^{15,16} consisting of two resistors connected in series, and a third resistor coupled be-

tween the junction node of the two series resistors and a reference potential (ground in this case), as shown in Fig. 1(b). If it is chosen that $R_2 = 100R_3$, the equivalent resistance of $R_2 \parallel R_3$ will be about the same as R_3 ($[R_2 \parallel R_3] = [100R_3 \parallel R_3] = \frac{100}{101}R_3 \simeq R_3$). Since typically $A > 10^4$,

$$\frac{A}{R_2}(R_2 \parallel R_3) \simeq A \frac{R_3}{R_2} = \frac{A}{100} \gg 1, \quad (3)$$

the transimpedance of the system shown in Fig. 1(b) becomes

$$T_{T-network} = \frac{V_{out}}{I_{in}} = -\frac{1 + \frac{R_1}{R_3} + \frac{R_1}{R_2}}{\frac{1}{A(R_2 \parallel R_3)} + \frac{1}{R_2}}$$

$$= -\frac{R_1 + R_2 + \frac{R_1 R_2}{R_3}}{1 + \frac{1}{\frac{A}{R_2}(R_2 \parallel R_3)}} \quad (4)$$

$$\simeq -\left(R_1 + R_2 + \frac{R_1 R_2}{R_3}\right). \quad (5)$$

As a result, by replacing the single feedback resistor with the T feedback network, the resistance of the feedback system can be reduced without sacrificing the overall transimpedance in a negative-feedback transimpedance amplifier system. For instance, using the designed resistance values of 47 k Ω , 18 k Ω , and 180 Ω for R_1 , R_2 , and R_3 , respectively, the transimpedance becomes about 4.8 M Ω . Thus, with roughly the same theoretical transimpedance, the resistance values of the feedback resistors can be dropped by about 100-fold. Since the bandwidth of a transimpedance amplifier is a function of the feedback impedance, reducing the feedback resistance increases the bandwidth. Despite different technologies being introduced for amplifier bandwidth extension,^{17–20} the T feedback network is undoubtedly another simple approach to preserve the bandwidth without sacrificing the transimpedance. Note that here, the resistors R_1 , R_2 , and R_3 can be replaced by complex impedances (such as Z_1 , Z_2 , and Z_3) in order to obtain transimpedance characteristics which vary with frequency according to particular requirements in a given application.

B. Noise

Reducing the noise from a preamplifier can be a major difficulty for a circuit designer. Typically, noise consists of all of the voltages and currents which accompany a signal of interest, and includes (a) Johnson-Nyquist noise (thermal noise), (b) the noise from active components or amplifiers (such as shot noise and flicker noise), (c) environmental noise, and (d) the noise of natural quantizing fluctuation.²¹ As suggested by Letzter and Webster,²¹ the first two types of noise, thermal noise and amplifier noise, are the major noise sources to be minimized for improving noise performance in a given amplifier circuit design.

FIG. 2. The noise model of an amplifier suggested by Letzter and Webster,²¹ including a noiseless amplifier with gain A , an equivalent amplifier noise current generator i_{na} , an equivalent amplifier noise voltage generator e_{na} , and a noiseless source (the input signal source) resistor R with its noise voltage generator e_n .

Caused by the thermal agitation of electrons inside conductors, thermal noise can be described as the noise power P_n ,

$$P_n = k_B T \Delta f, \quad (6)$$

which is a function of Boltzmann's constant k_B , the absolute temperature T , and the bandwidth Δf . To describe the thermal noise output from a resistor, the equivalent circuit of a noisy resistor can be modeled by a noiseless resistor R , either coupling in series with a noise voltage source e_n , or shunting a noise current source i_n , where

$$e_n = \sqrt{4k_B R T \Delta f}, \quad (7)$$

and

$$i_n = \sqrt{\frac{4k_B T \Delta f}{R}}. \quad (8)$$

Note that here the e_n and i_n are the root mean square voltage and current, respectively. Shot noise and flicker noise are considered electronic noise sources and are commonly seen in an active device. Usually, flicker noise dominates thermal and shot noise in the frequency range from direct current (DC) to ~ 100 Hz.²¹ To characterize the noise performance of an amplifier, Letzter and Webster²¹ suggested a model comprising a noiseless amplifier with both voltage and current noise sources, e_{na} and i_{na} , respectively, connected to the input, and a noiseless source resistor, R , with its noise voltage source, e_n , coupled in series to the input, as shown in Fig. 2.

It can be known from Eq. (6) that to design a preamplifier working in cryogenic temperatures^{9,22} can significantly reduce the thermal noise by about 10-fold. At a given temperature (such as at room temperature), to change the resistance value in an amplifier system can be an approach to alter the thermal noise voltage or current. To characterize such a change in an amplifier system illustrated by Fig. 2, the corresponding modification to either the noise current generator i_{na} , or the noise voltage generator e_{na} , or both, need to be evaluated carefully.

On the other hand, to limit the noise generated by the active components can be another approach to improve the signal-to-noise performance of a preamplifier. It can be done by reducing the number of active components being used. This work presents a new single-transistor transimpedance preamplifier, in which a common-source amplifier is used

as the main amplification stage, replacing the main amplification Op Amp, the AD8099 (Analog Devices, Norwood, Massachusetts),²³ and the input buffer, the junction field-effect transistor (JFET) BF862 (NXP Semiconductors, Eindhoven, The Netherlands),²⁴ in the previous design.⁶ The noise generated by the active components (both AD8099 and BF862) in the previous design is now limited. Only one transistor, which is the low noise JFET, BF862, with the equivalent noise input voltage of $0.8 \text{ nV}/\sqrt{\text{Hz}}$ specified on the data sheet,²⁴ in the new design is responsible for such noise. The T feedback network is also introduced to increase the bandwidth flexibility, and to avoid a large biasing resistor (for biasing both the gate terminal of the transistor and the detection plates of the ICR cell).

III. CIRCUIT DESIGN

A. Common-source amplifier design

A common-source JFET amplifier with source degeneration is usually formed by a single JFET, a drain resistor R_D , a gate resistor R_G , and a source resistor R_S (Q1, R4, R3, and R6, respectively, in Fig. 3(a)). The drain current is modulated by the JFET, according to the input voltage signal applied to the gate, and voltage output is expected after such modulated current is converted into voltage by the drain resistor (or the load resistor), R_D . The voltage gain A_v of a common-source amplifier can be written as

$$A_v = -\frac{g_m R_{out}}{1 + g_m R_S} \simeq -\frac{g_m R_D}{1 + g_m R_S}, \quad (9)$$

where g_m is the transconductance of the transistor, and R_{out} is the equivalent output resistance, which can be approximated as R_D , when any resistance shunting the drain resistor R_D is significantly larger than the resistance of R_D .

The first attempt to design a single-transistor transimpedance preamplifier has been made by adjusting the impedance values of passive components, to adjust the DC conditions and to fulfill the requirement of having a large transimpedance between the frequency of 1 kHz and 1 MHz, which corresponds to the mass-to-charge ratio, m/z , of ~ 180 – $180k$ in a 12-T FT-ICR system. The designed circuit is shown in Fig. 3(a).

In particular, the forward transfer admittance, y_{fs} ($= dI_d/dV_{gs}$, the ratio of the drain current change over the change of the gate-source voltage), of the JFET, BF862, increases with its drain current, as suggested by the data sheet.²⁴ Reduction of the resistance of the resistor R6 results in increase of the drain current, causing a larger transfer admittance. Although this larger transfer admittance leads to larger gain, it sacrifices the advantages of not only the low power consumption characteristic (due to the low drain current) to lower the heat dissipation, but also the large voltage gain variation between the DC and alternative current (AC) signals to make this design a better high-pass filter. Figure 3(b) shows the computer simulated correlations between the voltage gain, drain current (I_D), and power consumption of this common-source JFET amplifier (the circuit of Fig. 3(a) without the feedback loop) when five different source resistors (R6 in

FIG. 3. Common-source JFET amplifier with source degeneration and its performance. (a) Schematic of a single-transistor transimpedance preamplifier comprising a feedback loop (consisting of a feedback resistor R_1 and a DC blocking capacitor C_1) and a common-source JFET amplifier with source degeneration. (b) Computer simulated BF862 common-source amplifier (as shown in Fig. 3(a), but without the feedback loop) voltage gains and drain currents I_D with different source resistors R_s (R_6 in Fig. 3(a)). (c) Schematic of the single-transistor transimpedance preamplifier with a T feedback network. (d) Measured transimpedance of the single-transistor transimpedance preamplifier with a T feedback network (schematic shown in Fig. 3(c)), where two resistance values of the source resistors R_6 (47 Ω and 470 Ω) were tested to show the variation of the transimpedance according to the biasing condition change.

Fig. 3(a) are selected. As expected, the AC voltage gain is higher when a smaller source resistor is selected, at the cost of both higher DC drain current, and less DC and AC voltage gain difference.

B. Feedback loop

When a feedback resistor (R_1 in Fig. 3(a)) is added into this common-source amplifier to modify this circuit into a transimpedance amplifier, not only the DC blocking capacitor C_1 is needed to separate the DC potentials of the gate and the drain nodes, but there are loading issues to be carefully considered. It can be seen clearly from Eq. (9) that the drain resistance R_D plays a role when determining the voltage gain A_v . With the feedback loop coupled between the input and the

output, the feedback resistor R_1 becomes part of shunting resistance to R_D . As a result, there will be a minimum threshold resistance for R_1 to keep the voltage gain A_v at a reasonable scale for a given application.

Meanwhile, at the input node (the gate of the JFET), it is expected that the majority of the input signal current flows into the feedback resistor R_1 instead of the gate biasing resistor R_3 . Assuming that the input resistance of the JFET is too large (in comparison to R_1 and R_3) to be considered, the ratio of R_3 over R_1 should be so large that the input current grounded via R_3 can be negligible. For a larger transimpedance, a larger feedback resistor R_1 is desired, resulting in a desired much larger R_3 to push the signal into the feedback loop. Thus, there is also a maximum threshold resistance of R_1 or R_3 , such that the JFET input resistance can be neglected. Therefore, to push the transimpedance of this

single-transistor preamplifier, namely, to further increase the feedback resistance value, such gate biasing resistor has to be removed.

An effort has been made to relocate the biasing gate resistor into the feedback network. Consequently, the single feedback resistor of R_1 in Fig. 3(a) is replaced by a three-resistor, T-shaped feedback system, consisting of R_1 , R_2 , and R_3 shown in Fig. 3(c). The schematic shown in Fig. 3(c) represents the low noise single-transistor transimpedance preamplifier with a T feedback network. Without a gate resistor, the gate can still be properly biased to a fixed potential (ground in this case) through both resistors of R_1 and R_3 . Note that the input of the preamplifier is DC coupled to the detection plates of the ICR cell, and thereby provides the ground potential needed for stable ion trajectories.

Moreover, the output impedance of this single-transistor transimpedance preamplifier needs to be matched to the impedance of a 50- Ω transmission line in a FTMS system. A VA, such as the AD8099 VA shown in Fig. 4(a), can be introduced for this purpose. An overall transimpedance of 4.8 M Ω can be estimated by Eq. (5). However, by using a common-source amplifier instead of an Op Amp as the main amplification stage, the assumption of the open-loop gain $A > 10^4$ is no longer valid, causing the actual transimpedance to be smaller than the estimated value from Eq. (5). Such results can be validated later by the measured transimpedance data.

C. Printed circuit board

The computer-aided design software, Altium Designer Build 8.4 Service Pack 4 (Altium Limited, Sydney, Australia), was used to design a single-layer printed circuit board (PCB). All components were located inside a ground ring to assist shielding from environmental noise. Low noise surface mount thin-film chip resistors from Panasonic Corporation (Osaka, Japan) were selected to limit the intrinsic noise from the passive components. The board manufacturing was carried out in-house, using the PCB prototyping plotter ProtoMat S62 (LPKF Laser & Electronics, Garbsen, Germany).

The populated PCB (sized $\sim 100 \times 60$ mm², where the single-transistor preamplifier occupies the areas of about 20×15 mm²) is shown in Fig. S1 of the supplementary material.²⁹ This PCB is for circuit testing purposes. To test the FT-ICR performance with this newly designed amplifier, a smaller sized board should be designed, so that the stray impedance can be limited. The PCB manufactured for the previous design, an AD8099 preamplifier,⁶ is also used here to test the performance of the T feedback network.

IV. CIRCUIT TESTING

The National Instruments (NI) (Austin, Texas) PXI platform, consisting of a PXI-5122 oscilloscope, a PXI-5421 arbitrary waveform generator, a PXI-6733 analog output card, a PXI-8336 control card, a PXI-1042 chassis, and the software LabVIEW 2009 (Service Pack 1), was utilized for the measurements. The SPICE simulation was carried out using the computer simulation software NI Multisim v11.0.2. The

FIG. 4. The AD8099 preamplifier and the AD8099 voltage amplifier (VA). (a) Schematic of the AD8099 preamplifier with a T feedback network. When the transistor Q1, capacitors C1, C2, and resistors R1–R4 are disconnected, the rest of the circuit forms an AD8099 VA, which can be used to match the output impedance of the single-transistor preamplifier to a 50- Ω transmission line in an FTMS system. (b) Measured transimpedance of the AD8099 preamplifier with three feedback systems: (i) a single 18-k Ω feedback resistor shunting a 0.8-pF capacitor; (ii) a single 4.7-M Ω feedback resistor; (iii) the T feedback network as shown in Fig. 4(a).

circuit was powered by the DC power supply TTI EL301R (Thurlby Thandar Instruments, Huntingdon, UK). Twenty points were measured with a 100 MS/s sampling rate, and then averaged to obtain both the input current flowing into the node “ I_{in} ” and the output voltage at node “ V_{out} ” in the schematic figures. The peak-to-peak amplitude of the testing input sinusoidal current was 120 μ A for circuits with estimated overall transimpedance below 20 k Ω , and 20 μ A for circuits with estimated transimpedance above 1 M Ω .

The T feedback network was tested using the previously reported AD8099 preamplifier.⁶ Figure 4(a) illustrates the schematic of the AD8099 preamplifier with the T feedback network. A voltage amplifier using the same Op Amp AD8099 is also shown in Fig. 4(a). In this report, when testing the T feedback network using the AD8099 preamplifier, the resistance values were always $R_1 = 47$ k Ω , $R_2 = 18$ k Ω , and $R_3 = 180$ Ω , with a 0.8-pF capacitor shunting R_2 . When testing the newly designed single-transistor transimpedance

preamplifier with a T feedback network, the 0.8-pF capacitor was not connected. Since the AD8099 is a high speed Op Amp with gain-bandwidth product of about 3.8 GHz, the existence of the 0.8-pF capacitor (C1 in Fig. 4(a)) is to limit the bandwidth to around 1 MHz, which corresponds to the mass-to-charge ratio, m/z , of ~ 180 in a 12-T FT-ICR system. For the JFET BF862, its bandwidth characteristic will limit the 3-dB corner to around 1 MHz at a transimpedance of ~ 80 dB Ω . To compare the performance of the T feedback network with a single resistor feedback, a feedback system with only a 4.7-M Ω resistor (in such case, the capacitor C1 and the resistors R1–R3 were replaced by a 4.7 M Ω resistor, coupling the I_{in} and V_{out} nodes in Fig. 4(a)) is also tested.

The transimpedance measured by the NI PXI system is reported using the commonly used form of Bode magnitude plots. The magnitude axis of such a plot is often reported in decibel scale. Since the transimpedance (gain) of a transimpedance amplifier has a unit of $V/A = \Omega$, in a Bode plot, the transimpedance is reported using dB Ω , where dB Ω is defined as $X(\text{dB}\Omega) = 20 \times \log[Y(\Omega)]$, in which X and Y are the transimpedance in dB Ω and Ω , respectively.

V. RESULTS AND DISCUSSIONS

Figure 4(b) shows the frequency response using the AD8099 preamplifier in three feedback conditions: (i) a 18-k Ω resistor in parallel with a 0.8-pF capacitor (previously reported),⁶ (ii) a single 4.7-M Ω feedback resistor, and (iii) the T feedback network as shown in Fig. 4(a). The change of the DC blocking capacitor C2 could be also noticed. C2 was 220 nF in the previous design,⁶ when the 18-k Ω curve was measured. In order to set the low corner frequency at around 1 kHz (m/z of 180k in a 12-T system), the 220-nF C2 is replaced by a 4.7- μ F capacitor. A narrower bandwidth for the single 4.7-M Ω feedback curve can be noticed. It is caused by the parasitic capacitance of this 4.7-M Ω resistor. Despite the bandwidth variation, in Fig. 4(b) the measured transimpedance of the preamplifier using the T feedback network (consisting of 47-k Ω , 18-k Ω , 180- Ω resistors, and a 0.8-pF capacitor) shows a good agreement with the preamplifier using a single 4.7-M Ω feedback resistor.

The transimpedance frequency response of the single-transistor transimpedance preamplifier with the T feedback network is measured when a 470- Ω or a 47- Ω source resistor is used. As shown in Fig. 3(d), the use of a 470- Ω source resistor shows a flat transimpedance of a few dB Ω less than the 47- Ω source-resistance preamplifier. However, there is better filtering at low frequency for the curve with a 470- Ω source resistor. As the DC drain current is about 0.95 mA and 5.1 mA for the source resistance values of 470 Ω and 47 Ω , respectively, with the supplied voltage of 6 V, the power consumption when using a 47- Ω source resistor is about 25 mW more than, or around fivefold of, the power used in a preamplifier with a 470- Ω source resistor. More consumed power means more heat dissipation. Undesired heating can cause instability, so minimizing preamplifier power consumption in MS applications is important, especially when the preamplifier is located in vacuum close to the mass analyzing device for minimal parasitic capacitance from cabling.

The noise performance of the newly designed single-transistor transimpedance preamplifier with the T feedback network is not reported, since the noise spectrum is below the sensitivity of the spectrum analyzer, IFR A-7550 (Aeroflex, Plainview, New York), utilized to measure the noise performance of the preamplifier reported previously.⁶ In such a single-transistor preamplifier design, this T feedback network replaces not only the feedback resistor, but also the biasing resistor (for biasing both the input (gate) terminal of the transistor and the detection plates of the ICR cell). Thus, the amplifier noise cannot be simply modeled with a simple current or voltage noise source following a standard noise analysis procedures. Instead, the method suggested by Letzter and Webster²¹ must be employed for each resistor in the network, but calculating the “referred input noise current” and “referred input noise voltage” will be difficult, as both the noise and the gain are affected. To predict the noise behavior in this case calls for further careful research. To test the resulting signal-to-noise behavior, a noise analysis system with sensitivity better than the spectrum analyzer used for this report would be necessary.

To conclude, the T feedback network provides more bandwidth for a given transimpedance, since the parasitic capacitance shunts a reduced resistance in this feedback system. The T feedback network also increases the transimpedance and the flexibility to obtain transimpedance characteristics for a given application. The single-transistor transimpedance preamplifier with a T feedback network provides a very low noise preamplification solution, but less gain in comparison to an Op Amp based preamplifier, such as AD8099 preamplifier. Recall that the signal-to-noise performance of the complete system can be maximized electronically by not only reducing the noise, but also increasing the gain of a preamplifier. An optimized balance between the preamplifier gain and noise characteristics needs to be obtained for the best signal-to-noise performance in a given system.

VI. CONCLUSIONS

A new low noise preamplifier for an FT-ICR mass spectrometer using a single-transistor transimpedance preamplifier with a T feedback network is designed, manufactured, and tested. The characteristics of using a resistive T feedback network is studied by using an Op Amp. The introduction of the T feedback network in an Op Amp based transimpedance system allows ~ 100 -fold less resistance for a given transimpedance, hence more bandwidth can be preserved.

In response to the need of an ultra low noise preamplifier, a very low noise single-transistor common-source JFET amplifier is used as the main amplification stage. The very large open-loop gain offered by an Op Amp is traded with a better noise performance and a lower power consumption. In such a case, the measured transimpedance of around 80 dB Ω between the frequencies of about 1 kHz and 1 MHz satisfies the need of a 12-T FT-ICR mass spectrometer with corresponding mass-to-charge ratio, m/z , of approximately 180 to 180k. Alternatively, using the AD8099 preamplifier with the T feedback network, the transimpedance can be around 120 dB Ω with roughly the same bandwidth. To use the

single-transistor transimpedance preamplifier with the T feedback network as the front-end electronics for the FT-ICR MS, the reported AD8099 VA can be a buffer candidate to match the output impedance to the impedance of a signal transmission cable.

Although here the preamplifier is designed for the application of FT-ICR MS, a similar technique can be introduced to other mass spectrometers such as the Orbitrap,²⁵ which has a similar frequency range of interest to the FT-ICR MS, time-of-flight (TOF), and ion trap systems. Each of the three constructing elements in a T feedback network can be in any form of the combinations of resistive, capacitive, or inductive elements to create any complex impedance to fit the frequency and transimpedance requirements of any particular application. Since this design can be an ideal front-end amplification solution to be applied to any charge/current detecting device, it can be introduced to other areas of applications, such as nuclear magnetic resonance (NMR) spectroscopy²⁶ or other spectroscopy systems, microscopy systems,²⁷ optical communication systems,²⁸ or charge-coupled devices (CCDs).

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