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Note: A high dynamic range, linear response transimpedance amplifier

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We have built a high dynamic range (nine decade) transimpedance amplifier with a linear response. The amplifier uses junction-gate field effect transistors (JFETs) to switch between three different resistors in the feedback of a low input bias current operational amplifier. This allows for the creation of multiple outputs, each with a linear response and a different transimpedance gain. The overall bandwidth of the transimpedance amplifier is set by the bandwidth of the most sensitive range. For our application, we demonstrate a three-stage amplifier with transimpedance gains of approximately $10^9 \Omega$, $3 \times 10^7 \Omega$, and $10^4 \Omega$ with a bandwidth of 100 Hz. © *2012 American Institute of Physics.*
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High dynamic range transimpedance amplifiers with large bandwidth are useful for multiple applications, including scanning tunneling microscopy,[1] conductive atomic force microscopy,[2] photodiode readouts,[3] and other various instruments. In our application, which is a search for the electron's electric dipole moment using $(Eu,Ba)TiO_3$,[4] we require a high dynamic range transimpedance amplifier to monitor the wide range of currents during a polarization reversal of a ferroelectric sample. In addition, to prevent systematics due to electrical currents after the sample polarization is reversed, we also require a sensitive ammeter capable of detecting currents on the 1 pA to 10 pA level.

A standard way to increase the dynamic range of a transimpedance amplifier is to make a logarithmic amplifier by using a diode or FET as the feedback impedance. This makes the output voltage proportional to the logarithm of the input current.[1,5,6] In these amplifiers, the effective feedback impedance changes as a function of output voltage, creating the nonlinear response.

Using FETs or diodes in the feedback allows for simple designs; however, they can suffer from two problems. First, because the output voltage is proportional to the logarithm of the input current, simply averaging or filtering the output voltage to reduce noise will yield the incorrect current. One must first convert the voltage to current and subsequently average or filter, which prevents the use of filters, averaging, and Fourier transforms on modern oscilloscopes. Second, the user will have difficulty making accurate readings of the current by examining simple, laboratory voltage readouts because of the nonlinear response.

Ideally, to guarantee both linear response and high dynamic range, one would desire a transimpedance amplifier with multiple, linear outputs working in unison, with each output having a transimpedance gain that is two to three orders of magnitude smaller than that of the adjacent output. Our design has precisely this, as shown in Fig. 1. Here, there are three outputs, or "ranges," each with a transimpedance gain of approximately 300 times that of the adjacent range.

The fundamental issue that must be solved to guarantee proper operation for such a transimpedance amplifier is how to prevent saturation of a more sensitive range from adversely affecting a less sensitive range. Our design, which is similar to a design found in Wandlowski and Meszaros[7] and is shown in Fig. 2, prevents saturation of the amplifier's more sensitive ranges by shorting the appropriate feedback impedance. When the output of the OP1 is not saturated, the inverting input will be held at ground. In this case, the outputs of all three ranges are iGR_3, $iG(R_2 + R_3)$, and $iG(R_3 + R_2 + R_1)$ for ranges 3, 2, and 1, respectively. In this case, $G = (1 + R_4/R_5)$ which is the additional gain on the output buffer amplifier and $R_3 \ll R_2 \ll R_1$ to ensure sufficient dynamic range. Because of the presence of the JFETs, if the output voltage of the primary op-amp (OP1) reaches either $\pm(V_b + V_{gs})$, where V_{gs} is the gate source turn on voltage for the JFET, the respective JFET will begin to conduct and bypass current around R_1. This prevents saturation of OP1. Similarly, as the voltage at the point between R_2 and R_1 with respect to ground grows close to $\pm(V_b + V_{gs})$, the JFETs in parallel with R_2 conduct current around that resistor, again preventing saturation of OP1 and OP2. Therefore, the saturation problem is avoided, and three independent ranges work in unison to produce linear responses with three different transimpedance gains.

For our application, we wanted range 1, the most sensitive range, to be approximately 1 nA/V sensitivity. With a $R_1 = 100$ MΩ and R_4 and R_5 chosen such that the second stage amplifiers yield a gain of 11, that sensitivity is achieved and currents down to approximately 1 pA to 10 pA can be easily recorded with a 16-bit analog-to-digital converter. The largest current we expected to measure in our application is 1 mA. Therefore, we chose $R_3 = 1$ kΩ such that the range 3, the least sensitive range, would correspond to 100 μA/V, making 10 V correspond to 1 mA. Lastly, we chose $R_2 = 300$ kΩ to cover the middle current decade with a sensitivity of approximately 300 nA/V. The output voltage is shown in Fig. 1, which verifies the linear behavior for all three ranges.

Because the lowest range of the amplifier is capable of reading such small currents, care must be taken to select components that will prevent leakage of current out of the feedback chain to ground. We initially selected and designed the

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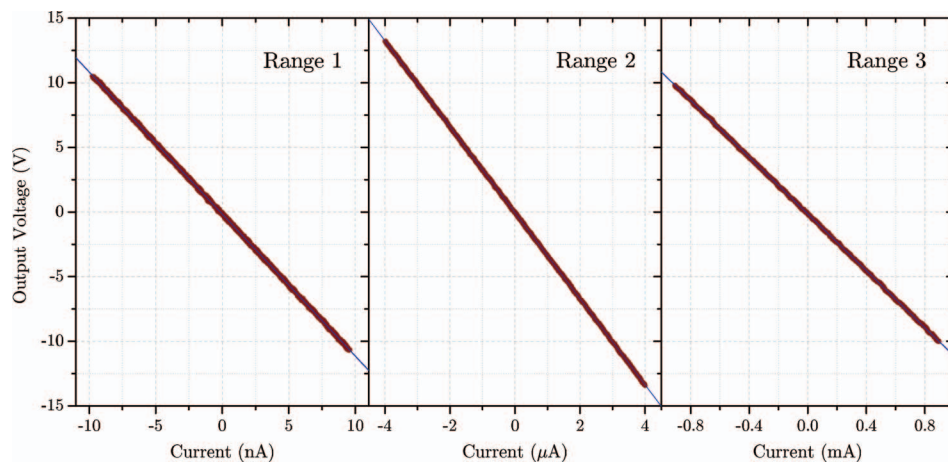
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FIG. 1. (Color online) Output voltage vs. input current for (left) range 1, the most sensitive range, (middle) range 2, and (right) range 3, the least sensitive range.

amplifier with the AD549 as the op-amp of choice for the primary op-amp (OP1) and the follower op-amps (OP2 and OP3) since they have ultralow input bias JFET input stages (around 10–100 fA, depending on the model). However, with a 100 MΩ feedback resistor, a 1pA input bias current amplifier, such as the OPA124 or AD795, would be sufficient to maintain accuracy. For the JFETs, we desired low I_{gss} and low I_{ds} and selected the 2N5460 P-type and the J202 N-type which have approximately 100 pA and 2 pA I_{gss}, respectively. However, we note that the 2N5457 N-type is a more complimentary part to the 2N5460 and easier to obtain. We set the bias, V_b, to be one-half the supply voltage for OP1 to maximize its voltage output range. For the OP4 op-amps, which serve as a buffer to protect the AD549s and increase the output voltage range from $\pm(V_b + V_{gs})$ to the full voltage swing of the amplifier, we chose the OP27 for its low input offset voltage.

Lastly, it was empirically determined that the additional capacitance contained in the JFETs would cause some undamped oscillation of the amplifier if the feedback resistance in parallel was too high. Some additional capacitance (C_1, C_2 ~ 20 pF) was added in parallel to the feedback resistors in order to prevent oscillations of the amplifier.

The accuracy of the amplifier will be determined primarily by the offset voltages. In particular, the offset of range 1 will be set primarily by the input bias current of OP1, the input offset voltages of OP1, and the connected buffer amplifier OP4. In our application, these offsets can easily be of order 1 mV which corresponds to 1 pA. Likewise, the higher ranges' accuracy will be set by the offset voltages of their respective op-amps which, in our application, are of order 0.3 mV. With a factor of 300 between the ranges and with a more sensitive range saturating at 10 V, the user would begin reading the next range when its output is ~30 mV, and therefore having an
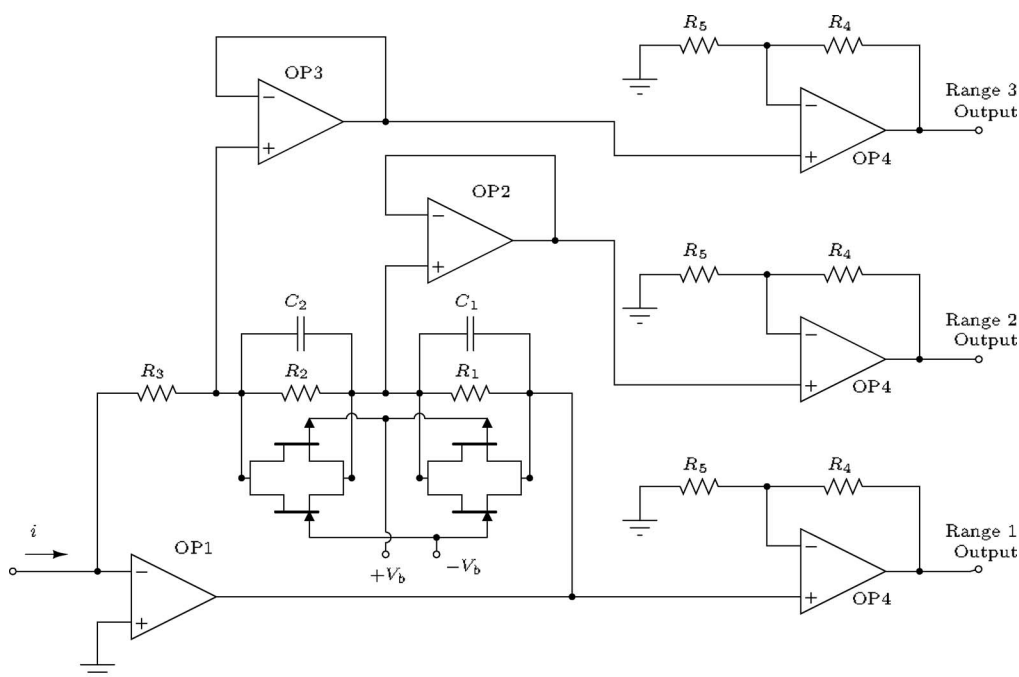

FIG. 2. A schematic diagram of the linear transimpedance amplifier. See text for description.
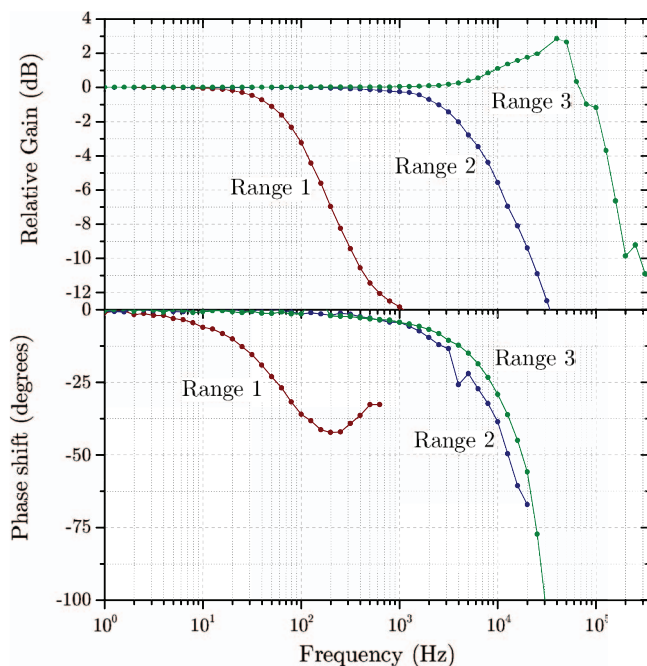
FIG. 3. (Color online) Frequency response (both phase shift and gain) for the various ranges of the transimpedance amplifier described. The red lines represent range 1, the highest sensitivity range; the blue lines represent range 2; and the green lines represent range 3, the least sensitive range. The points mark the frequencies at which data were taken.

error of order 1%. In principle, these offsets of the higher ranges can be measured and corrected when the most sensitive range reads near zero current. Alternatively, the user can choose the feedback resistances and operational amplifiers carefully to reduce these errors to acceptable levels depending on the application.

Since each feedback resistance (R_1, R_2, and R_3) is tied to a different capacitance, each range has a different frequency response, with less sensitive, or lower transimpedance gain ranges having a larger frequency response at high frequencies, as shown in Fig. 3. When signals are applied which both saturate high sensitivity ranges and are at a frequency greater than the high sensitivity ranges' bandwidth, the output waveform can be distorted. When a signal sweeps through zero current quickly, the higher sensitivity range may have a delayed response relative to the lower sensitivity ranges causing a momentary waveform distortion as the higher sensitivity range sweeps through zero. This leads to an effective increase in the RMS output voltage, which leads to the gain observed in range 3 at high frequencies. This anomalous gain is not seen if the amplifier is driven with a small, ac signal on top of a larger, dc offset which has no polarity switching.

In order to prevent this unwanted gain from affecting the waveform shape, we added third-order, low-pass Bessel filters using the voltage-controlled voltage source design to the outputs to limit all ranges to the bandwidth of highest sensitivity range. Using Bessel filters provides better waveform preservation than Butterworth filters; however, this comes at the expense of a more rounded knee in the transition between the passband and stopband.

The effects due to temperature variations are suppressed because the components most sensitive to temperature variations, the JFETs, are only used to change the currently active range of the amplifier. Therefore, any temperature variations will then only affect the voltage at which the switch from range to range will occur. Since the next range will typically be read before saturation of the previous range occurs, this produces no real effect on the performance of the ammeter. Moreover, the calibration of each range will remain virtually unchanged, where the dominant temperature shift will be caused by the shift in the feedback resistance. We found that this amplifier, with its temperature drift dominated by the resistors in the feedback, can be of order 100 ppm/°C with appropriately chosen resistors.

Unlike logarithmic amplifiers, whose noise goes as $1/i$, where i is the current, the noise function for this amplifier is a step-like function, since each range has its own level of noise. For currents that are near the small signal end of a particular range, this could lead to a relative $\delta i/i$ which can be larger than that of a logarithmic amplifier. However, for currents that are near the high end of a given range, the relative error $\delta i/i$ would be correspondingly lower than that of a logarithmic amplifier. If the user requires high precision for a certain range of currents, the feedback resistors can be chosen to optimize the recording of those ranges of currents.

In conclusion, we have designed a high dynamic range, linear response transimpedance amplifier. The amplifier has three different outputs, each with a linear response and different transimpedance gain, which work in unison to measure a single current. In order to achieve this behavior, JFETs are combined with resistors which give linear responses and prevent saturation of the operational amplifiers in the circuit.

[1]U. Durig, L. Novotny, B. Michel, and A. Stalder, Rev. Sci. Instrum. **68**, 3814 (1997).

[2]L. Aguilera, M. Lanza, M. Porti, J. Grifoll, M. Nafria, and X. Aymerich, Rev. Sci. Instrum. **79**, 073701 (2008).

[3]Y. B. Acharya and P. D. Vyavahare, Rev. Sci. Instrum. **69**, 595 (1998).

[4]A. O. Sushkov, S. Eckel, and S. K. Lamoreaux, Phys. Rev. A **81**, 022104 (2010).

[5]C. A. Grimbergen and G. H. P. Kohnke, Rev. Sci. Instrum. **47**, 854 (1976).

[6]R. Ahrenkiel, Rev. Sci. Instrum. **40**, 78 (1969).

[7]T. Wandlowski and G. Meszaros, "High-dynamic ammeter with small time constant," WIPO patent application WO/2009/000236 (December 31, 2008), see http://www.sumobrain.com/patents/WO2009000236.html.