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Note: A temperature-stable low-noise transimpedance amplifier for microcurrent measurement

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Temperature stability and noise characteristics often run contradictory in microcurrent (e.g., pA-scale) measurement instruments because low-noise performance requires high-value resistors with relatively poor temperature coefficients. A low-noise transimpedance amplifier with high-temperature stability, which involves an active compensation mechanism to overcome the temperature drift mainly caused by high-value resistors, is presented. The implementation uses a specially designed R-2R compensating network to provide programmable current gain with extra-fine trimming resolution. The temperature drifts of all components (e.g., feedback resistors, operational amplifiers, and the R-2R network itself) are compensated simultaneously. Therefore, both low-temperature drift and ultra-low-noise performance can be achieved. With a current gain of 10^{11} V/A, the internal current noise density was about 0.4 fA/$\sqrt{\text{Hz}}$, and the average temperature coefficient was 4.3 ppm/K at 0–50 °C. The amplifier module maintains accuracy across a wide temperature range without additional thermal stabilization, and its compact size makes it especially suitable for high-precision, low-current measurement in outdoor environments for applications such as electrochemical emission supervision, air pollution particles analysis, radiation monitoring, and bioelectricity. *Published by AIP Publishing.* [http://dx.doi.org/10.1063/1.4974741]

Accurate measurement of microcurrent currents is of interest in both fundamental science and applied engineering because many physical phenomena are accompanied by fA–pA-scale current changes.[1–8] For the microcurrent measurement, the commonly used methods are current-voltage conversion (transimpedance amplifier), current-frequency conversion, and capacitance integral.[7,8] The transimpedance amplifier is the most commonly used circuit for the input stage in continuous microcurrent measurement;[1] it converts weak current signals to voltage outputs with measurable amplitude. A high-value feedback resistor is a critical component of the transimpedance amplifier because the internal current noise is mainly due to the Johnson effect brought by it. According to the Johnson noise theory, the internal current noise reference to the input[4] is expressed as

$$I_{nf} = \sqrt{4kTB/R_f}, \tag{1}$$

where k is the Boltzmann constant, B is the bandwidth, and T is the Kelvin temperature. Eq. (1) indicates that the current noise I_{nf} decreases as R_{f} increases. Therefore, a goal in low-noise microcurrent amplifier design is to increase the feedback resistance as much as possible under the amplifier's permitted dynamic range. However, unfortunately, the temperature stability of high-value resistors is far worse than that of ordinary ones:[9] the resistivity of manufacturing materials with general temperature stability (i.e., metal alloys or metal oxide films) is too low for high-value resistors, and high-resistivity materials (e.g., ruthenium compounds) are more temperature sensitive. The typical temperature coefficients of resistivity (TCR) are often up to 200–1000 ppm/K for commercially available high-value resistors above 1 GΩ,[10] and in general, resistivity is negatively associated with temperature stability. To obtain high measuring accuracy while maintaining low-current noise, the ambient temperature range must be limited or an extra thermal stabilization system is required to reduce the temperature drift. These two methods can only be used in the laboratory, but many microcurrent measurements are conducted outdoors.[6] Another method is to use serially connected high-stability low-value resistors instead of a high-value resistor.[11] However, large system size and complexity are the bottlenecks of these methods.

Many microcurrent measurements require continuous operation in outdoor environments.[6] These applications may experience temperature variations of ≥50 K, and they need to be low-noise, small-sized, and relatively low-cost. For these measurement applications, we introduced an active compensation method implemented by a temperature sensor and a programmable gain amplifier (PGA) (Fig. 1). The first stage is regarded as a temperature-sensitive amplifier, and the gain drift is compensated in the second stage, which consists of a gain-compensated PGA connected to a compensation unit and controlled by a temperature sensor. All circuits are arranged on the same high-thermal-conductivity substrate, and the temperature drifts of R_{f} and all other components are corrected

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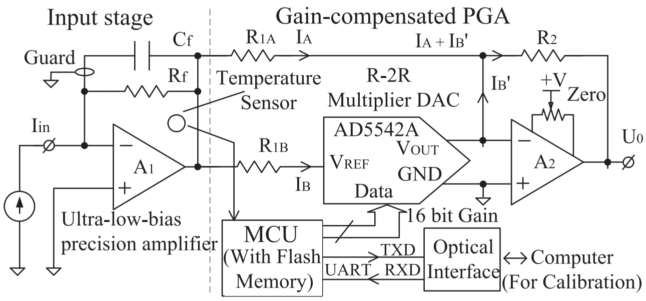

FIG. 1. Schematic diagram of temperature-stable low-noise transimpedance amplifier.

simultaneously. It should be pointed out that even a high-value resistor with a worse temperature coefficient may be employed. Therefore it reduces the costs while maintaining relatively high-temperature stability.

The input stage is a classical transimpedance amplifier with a 100 GΩ feedback resistor, expected current gain 10 V/100 pA, and ideal Johnson noise 0.4 fA/$\sqrt{\text{Hz}}$ at 23 °C. The high-value resistor is an ordinary commercial ruthenium-based resistor (Vishay Semiconductor) with a TCR of approximately 1000 ppm. C_f is the stray capacitance of ∼0.1 pF. The high-value resistor together with the stray capacitance limits the bandwidth of the amplifier to ∼15 Hz.

In this proposal, the required gain adjustment range performed by the PGA is only a few percent, but it needs very high gain resolution to track the minor gain drifts. We note that commercial PGA circuits mainly function on the scale of the dynamic range of overall gain (typically dozens of dB), which is not suitable for fine trimming of the gain within a small range. Thus, we developed a fine-trimming PGA (FT-PGA) circuit to compress the PGA adjustment range while providing much finer gain resolution (2nd stage in Fig. 1). The FT-PGA constitutes external bypass resistors R_{1A} and R_{1B} and a 16-bit R-2R network-based multiple digital-to-analog converter (DAC). The equivalent circuit is shown in Fig. 2.

The voltage gain of the FT-PGA is

$$A_2 = \frac{U_o}{U_1} = -\frac{(I_A + I_B')R_2}{U_1} = -\left(\frac{D}{2^N}\frac{R_2}{R_{1B}+R} + \frac{R_2}{R_{1A}}\right), \tag{2}$$

and the step or resolution of the gain is

$$A_{STEP} = \frac{1}{2^N}\frac{R_2}{(R_{1B}+R)}. \tag{3}$$

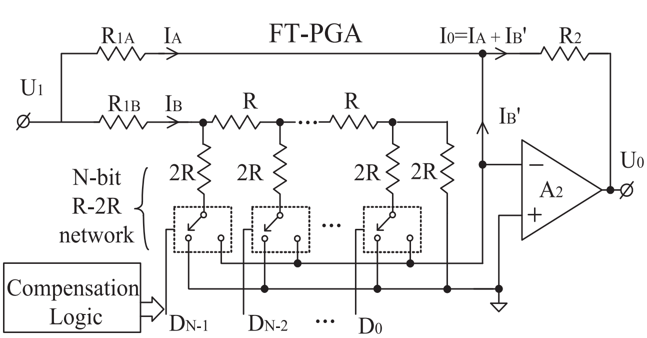

FIG. 2. The schematic structure of the fine-trimming PGA (FT-PGA).

According to Eq. (2), the minimal gain is $|A_{\min}| = R_2/R_{1A}$ when $D = 0$, and the maximum gain is $|A_{\max}| \approx R_2/(R_{1B}+R) + |A_{\min}|$ when $D = 2^N - 1$. Therefore, the gain can be adjusted linearly from $A_{\min}$ to $A_{\max}$ by the variation of D. To balance the adjustable margin between the positive and negative drifts, $|A_{\mathrm{mid}}|$ should be set to 1.0 when $D = 2^{N-1}$, satisfying

$$R_2 = R_{1A} || 2(R_{1B}+R). \tag{4}$$

$A_{\min}$ can be set to slightly less than 1 by choosing an R_{1A} value slightly larger than that of R_2, and $A_{\max}$ can also be set to slightly greater than 1 by selecting a proper value of R_{1B}. This achieves programmable gain modification in a small range near 1.0 with high resolution.

To obtain a gain adjustment range of 1.00% ± 2.50%, we set $A_{\min} = 0.975$ by choosing $R_{1A} = 9.76$ kΩ and $R_2 = 10$ kΩ, and then we set $A_{\max} = 1.025$ by choosing $R_{1B} = 191$ kΩ and R = 10 kΩ. According to Eq. (3), the gain step is $1/(20 \times 2^N)$, and the gain resolution is 20 times that of full-range PGAs. In this case, ∼0.75 ppm trimming gain resolution can be achieved by using a 16-bit R-2R network.

The integral nonlinearity error of the DAC multiplier is as low as 1 LSB (Analog Device AD5542A), which ensures linearity and monotonic gain trimming. The temperature is acquired by a digital thermal sensor with 16-bit resolution (Analog Device ADT7310, with 0.0078 °C thermal resolution), and then the compensated gain for the second stage is calculated and executed by the FT-PGA circuit in real time. The gain compensation lookup table was prestored in the flash memory of the micro-controller unit (MCU). An optical interface connects the MCU to the computer, to transfer the commands and data during the calibration and isolate the ground current interference from computer.

The internal noise performance was investigated in the frequency range 10 mHz–10 Hz. The amplifier's input terminal was open-circuited and covered by a metal cap to minimize the external interference. Output voltage sequence $\mathbf{V_{nk}}$ was recorded by a voltage meter (Keysight DMM-4050) with a sampling rate $F_s = 23.6$ sps and a sampling length $N = 7200$, then the total sampling time is $T = (N-1)/F_s = 305$ s. The current noise sequence is $\mathbf{I_{nk}} = \mathbf{V_{nk}}/R_f$, and then the current noise density is obtained via Fourier transform,

$$\mathbf{I_{nf}} = \frac{2T}{N}\cdot|\mathrm{FFT}[\mathbf{I_{nk}}]|. \tag{5}$$

The result shown in Fig. 3 indicates that a low white noise level of 0.4 fA/$\sqrt{\text{Hz}}$ was achieved above the corner frequency of ∼2.5 Hz. At lower frequencies, noise in excess of 1/f was observed, and it was mainly contributed by pink noise from the front stage of the amplifier and the high-value resistor.

The amplifier's transfer gain was calibrated throughout the range of 0–50 °C at 5 °C intervals before temperature dependence testing. Then, the amplifier was placed in a stabilized air bath, and the ambient temperature was varied from 0 to 50 °C with a ramp rate of ∼5 °C/h. The input current was kept constant at 100.00 pA, and then the transfer gain drift was calculated from the output voltage deviations during the temperature variation. The results, shown in Fig. 4, were compared with those of the noncompensated case.

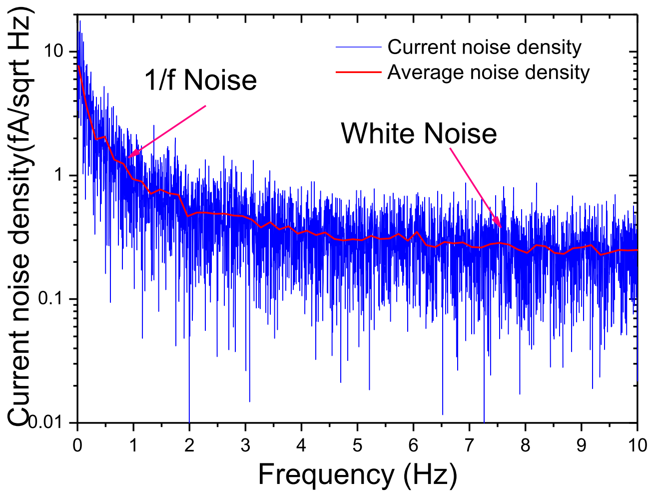

FIG. 3. Input current noise density.

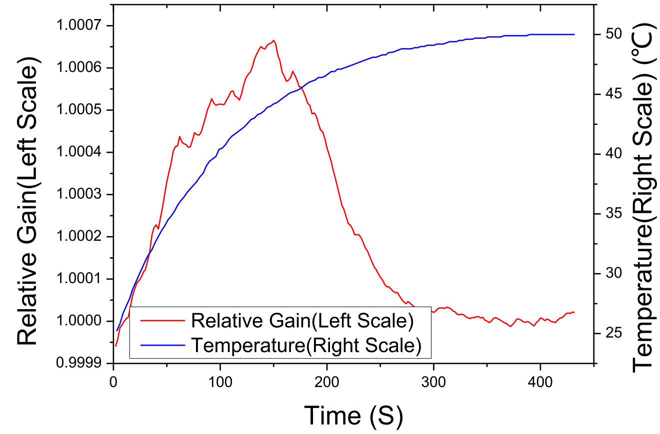

FIG. 5. Gain drift after temperature transits from 25 °C to 50 °C.

The noncompensated circuit presented 5% (50 000 ppm) of gain drift and a temperature coefficient of up to –910 ppm/K at 0–50 °C, mainly caused by the high-value resistor's temperature drift. The introduction of the compensation mechanism reduced the average temperature coefficient to 4.3 ppm/K; the maximum drift is not more than 300 ppm over the entire operating temperature range and less than the uncompensated drift by a factor of 150. The nonmonotonicity of the drift curve is caused by the unexpected residual error in interpolation of the compensation coefficient.

Another experiment was performed to test the temperature transition. The thermostat controlling the amplifier environment was moved from room temperature (25 °C) to 50 °C immediately. The resulting time traces of internal temperature and gain drift are displayed simultaneously in Fig. 5. The appearance of continued positive deviation after temperature transients is caused by the inevitable minor thermal gradient between the sensor and the high-value resistor during the heating up.

The amplifier module exhibited high-temperature stability, i.e., the average temperature coefficient was 4.3 ppm/K at 0–50 °C and a low-noise performance 0.4 fA/$\sqrt{\text{Hz}}$ with a current gain of 10^{11} V/A, even when an ordinary high-value resistor with high TCR was employed. This was attributed to two factors: (1) the proposed FT-PGA circuit provides a fine trimming resolution for nullifying gain drifts in real time; (2) the overall system is small in size, and because they are mounted on identical thermally conductive material, all components' gain drifts were compensated simultaneously. This method provides a flexible approach to compensate for any nonlinear or nonmonotonic thermal drift in a current amplifier. The elimination of thermal stabilizers and the ability to use high-value resistors with large TCR values remarkably reduce the cost and the size. The system is especially suitable for high-precision, low-current measurements in outdoor environments.

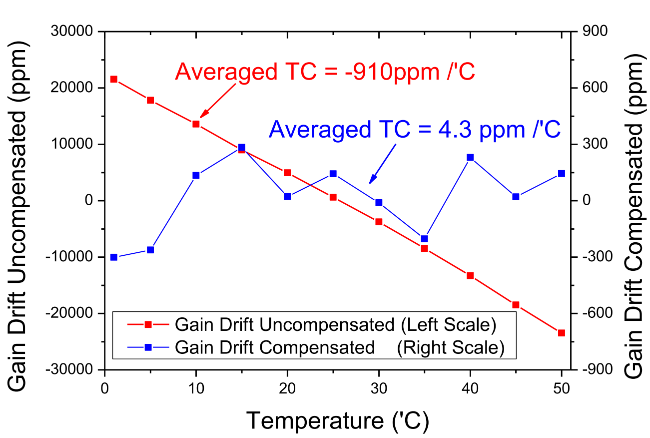

FIG. 4. Gain drift vs. ambient temperature.

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