A novel wide range electrometer with quasilinear output

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An electrometer using complementary metal–oxide–semiconductor bilateral switches for automatic range control is described. Based on a dual-switching configuration, the electrometer is capable of current measurement over eight orders of magnitude. Together with a voltage shifter, this circuit provides a quasilinear analog output for the measurement of input dc current between $10^{-3}$ and $10^{-12}$ A. The application of integrated circuits makes the whole system compact and extremely low cost. © 1997 American Institute of Physics.

I. INTRODUCTION

Electrometers are the instruments widely used for translating various kinds of physical quantities into electrical signals through suitable sensor devices. They are applied in many research areas such as radiation or particle detection, medical diagnostics, aerosol monitoring, and solid-state characterization. A number of studies on the construction of electrometers have been reported in the literature.1–4,7 It is usually preferable for an electrometer to cover the measurement of electrical current over a multidecade range. Particularly, in the case of long-term recording of a wide range current signal having a fine structured variation in each decade range, a quasilinear electrometer capable of providing a single output for representing the full range of signals with a linear relation would be highly useful.

In order to have the capability of wide range measurement, an electrometer is usually designed with range switching. The inclusion of range switching in an electrometer may be done by changing the feedback resistance of a transimpedance amplifier. This approach requires some high quality contact switches to maintain minimum leakage and noise coupling. There are miniature reed switches utilized for electrical control in various kinds of circuit design. However, the problems of contact bouncing in the reed switch and electromagnetic interference from the driving coil of the switch limit their application in the design of a highly sensitive electrometer. Another scheme for wide range coverage is the use of a logarithmic current-to-voltage converter. Logarithmic conversion devices, usually the diode junction of a bipolar transistor, in the feedback circuit permit the electrometer to measure a wide range of current without switching. However, a complicated circuit is inevitable to minimize thermal drift of the conversion factor.1 Due to the associated temperature-compensating circuit, the logarithmic conversion measurement is limited to a predefined polarity, which is determined by the configuration of the logarithmic device. A decibel-represented signal, as compared to the linear signal, may also be the lack of resolution in a specified range.

In a previous paper,4 we described a simple electrometer circuit associated with complementary metal–oxide–semiconductor (CMOS) bilateral switches. It provides a linear display with automatic range selection over six decade ranges. It has been used satisfactorily in an ionization gauge controller for the measurement of pressure under a high vacuum environment.5 In this report, an improved modification on the electrometer circuit will be described, which gives accurate and stable output over a $10^{-3} – 10^{-11}$ A range under a reasonable humidity and thermal environment. Together with an automatic range switching and voltage shifting circuit, the resultant electrometer provides a quasilinear analog output over eight decade ranges. This electrometer is very compact in size and can be easily built. It was found to be extremely convenient for general purpose current monitoring, especially when a close attachment of the electrometer to the current source is needed.

II. CIRCUIT PRINCIPLE

Figure 1 is the block diagram of the electrometer circuit system. A transimpedance-type electrometer circuit based on a high input–impedance operational amplifier is connected to eight feedback resistors of value $1 \Omega – 10 \Omega$. Each resistor is selected via a pair of bilateral switches. These switches are wired to the automatic range controller, which senses the output of the electrometer to determine the actuation of proper switches. The electrometer output is further processed by the voltage manipulator to give quasilinear voltage output.

Figure 2 is the circuit diagram of the electrometer and voltage manipulator. The high input–impedance op–amp U1 is configured as the current-to-voltage converter with feedback resistors $R_f$, which are selected via the CMOS switches Sa1-8 and Sb1-8 driven by the control voltage $Q$ and $\bar{Q}$, respectively. Since the leakage current of a typical CMOS switch is around 10 pA, the dual switching configuration must be used for isolation between the different feedback resistors. The leakage from the power supplies and control signals to the sensitive input of the U1 is minimized by the careful design of the guarding circuit around the input terminal. The trimmer resistors VR100K and switches Sc1-8 are provided for offset nulling for each individual range. The output voltage, $V_{o+}$ of the U1 is connected to the summing amplifier U2b to give a voltage shift. The amount of voltage shift is determined by the switches Sd1-8 and the voltage ladder given by the divider resistors $R_d$. The reference voltage $+/-5V$ connected through the switches $S_e$ and $S_f$, and the adjustment of potentiometer Rs give a 1 V shift for every advancement of a decade range. Consequently, a quasilinear
analog voltage, $V_q$, is obtained at the output of U2b. The voltage inverting circuit U2a and switches $S_a, S_b$ are provided for the electrometer to measure either positive or negative input current. The comparator circuits U3a and U3b give the driving signals, $V_a$ and $V_y$, for switches $S_x$ and $S_y$, respectively. Switch SW1 is used to define the assumed polarity of input current $I$. If $I>0$, SW1 should be at position 1 to assure that $S_x$ is open while $S_y$ is closed and $V_a=V_o$. On the contrary, when $I<0$, SW1 should be at position 3 to give $V_a=-V_o$. When SW1 is at position 2, the comparators U3a,b sense the polarity of $V_o$ and automatically make $V_a=-|V_o|$. Voltage $V_a$ is used for driving the automatic range controller.

An automatic range controller described previously is shown with some modification in Fig. 3. Signal $V_a$ is low-pass filtered and sensed by the comparator circuit formed by U4a,c. The upper and lower threshold voltages of the comparator are $-0.095$ and $-0.95$ V, respectively. Its output determines the advancement of the count in the binary up/down counter U5, which is clocked by a pulse generator U4a through switch S1. The binary data of U5 are transformed by the binary-coded-decimal-to-decimal decoder U6 and the digital inverter, U7, U8, to give the control voltage $Q$ and its inverse $Q$. Accordingly, when switch SW2 is at position $Auto$, the electrometer is automatically ranged over eight decade ranges. A voltage of $0–8$ V at $V_q$ represents an input current corresponding to $0–10^{-3}$ A with the quasilinear scale. That is, if input current $I=a \times 10^{-n}$ A, where value $a=0–0.95$ and integer $n=3–10$, then $V_q=\pm (a+(10-n))$ V depending on the setting of SW1 or current polarity. When switch SW2 is at position $Manual$, the range can be selected by pressing the up and down push buttons.

III. PERFORMANCE AND DISCUSSION

This electrometer circuit, based on a general purpose op–amp with a metal–oxide–semiconductor field-effect transistor input stage, CA3130, has been built with well-designed guarding and shielding. The thermal problems affecting the circuit performance are eliminated by careful adjustment of the trimmer resistors. Although many trimmers in Fig. 2 are required to be set, the adjustment of the circuit is straightforward since the trimmers are not interactive with each other. By setting the automatic range controller in $Manual$ mode and the input current at zero, the trimmer $R_s$ at U2b is adjusted to give a precise 1 V shift for one range change. The offset nulling trimmers VR100K at U1 are also adjusted to make $V_o=0$ for each individual range. It was found that the offset nulling trimmers are particularly needed for the ranges with 1 and 10 GΩ feedback resistors. The offset voltage of the unit-gain op–amps U2a and U2b is, typically, 2 mV, which can be properly nulled, if necessary. Once the zero setting for each range has been adjusted at room temperature, the output of U1 showed an offset drift of less than 1 mV when the range was switched over the full decade ranges even for a $+/-5^\circ$ C temperature variation, provided that the ambient humidity is kept low and the circuit components are mechanically rigid. This agrees with the typical 5 μV/°C input offset voltage drift of CA3130. Our experiments showed that the offset drift can be up to 10 mV for the most sensitive range when the humidity was raised from 60% to 80%. This may be due to a possible leakage path through the surface of the glass-fiber based circuit board surrounding the input terminal of U1. It suggests that a housing with a clean and dry atmosphere is essential for our electrometer to make a stable measurement of low direct current. The use of low leakage material such as Teflon and a better guarding would improve the situation. The problem of offset stability can also be eliminated by including an ac coupled circuitry to the whole system, which will be reported elsewhere.

FIG. 1. Block diagram of the electrometer circuit system.

FIG. 2. Circuit of the transimpedance electrometer and voltage manipulator with U1, CA3130; U2, U3, LM358; Sa, Sb, Sc, Sd, Sx, Sy, CD4066. The circuit is powered by $+/-5$ V except U2, which is powered by $+/-12$ V.

FIG. 3. Circuit of the automatic range controller with U4, LM324; U5, CD4029; U6, CD4028; U7, U8, CD4069; S1, S2, S3, CD4066. The circuit is powered by $+/-5$ V.
The measured mean-square noise over 1 kHz bandwidth was below 2 mV for the most sensitive range. This noise level is closely dependent on the input equivalent noise of op-amp U1 and its peripheral circuit components. The two diodes at the input for circuit protection are normally unbiased and introduce very little noise. The noise injection through off-switch capacitance of Sa1–Sa8 and Sb1–Sb8 is negligible since they are located as close to U1 as possible and the terminals of these switches are virtually at ground potential. The Johnson noise generated by the 10 GΩ feedback resistor at room temperature is

\[
\overline{v^2} = (4kT RΔf)^{1/2} = 407 \mu \text{V}
\]

Assuming that the excessive 1.96 mV noise is due to the white Gaussian noise of U1 alone, the calculated input current noise is 6.2×10^{-15} A/√Hz, which is about the typical value of a low noise field-effect transistor-input op amp such as CA3130. Actually, a simple low-pass buffer amplifier following the output \( V_q \) to narrow down the signal bandwidth will greatly reduce the noise voltage. Since the electrometer is basically linear, a suitable source modulation together with correlation detection can be adopted to further improve the noise characteristics of the whole system.

Figure 4 shows an oscillogram of the output \( V_q \) for zero input current when the electrometer was scanned manually over the full decade ranges. There were transient spikes at the range boundaries due to the charging and discharging the circuit capacitance in the feedback loop, which is more pronounced for higher feedback resistance of the transimpedance amplifier.

This electrometer has been tested using a current source based on a precision programmable voltage source with 1 mV resolution and bridge-calibrated resistors up to 10 GΩ. Because of the dual-switching configuration and the small on-state resistance of the CMOS analog switch, the accuracy of the electrometer is mainly determined by the tolerance of feedback resistors. It was found that the measurement accuracy can be within 10% for the most sensitive range. An accuracy of better than 2% was achieved by using resistors of <1% tolerance at ranges with 10 MΩ, 1 MΩ, 100 kΩ, and 10 kΩ resistors. The potentiometer for the 1 kΩ feedback resistor is needed for compensating the on-state resistance, typically, 150 Ω, of a CMOS switch. When the input current is varied slowly from 0 to +1 or −1 mA, the automatic ranging feature functions satisfactorily for both polarities. Figure 5 shows a typical measurement of reverse saturation current of a p–i–n diode under a thermal cycling test. The diode at room temperature was heated over the period \( t_1 \) to \( t_2 \) and then cooled to the ambient temperature by itself. A quasilinear scale of reverse current was recorded over three decade ranges between 10^{-11} and 10^{-8} A. The resultant small variation in the reverse saturation current during the heating cycle was monitored directly. The transient spikes at the boundary of range switching having a width of several milliseconds, which depends on the time constant in the feedback loop of electrometer, can be easily identified. It was noted that a discontinuity appeared on the range boundary near 150 s. This is a consequence of the large width transient spike produced when the circuit switches into the most sensitive range.

In conclusion, a transimpedance electrometer using CMOS bilateral switches with a dual-switching configuration for range scaling has been shown to have high performance over eight decade ranges of input current. The combination switching, together with automatic range selection and a voltage shifter, provides a quasilinear analog output. The compactness, solid-state circuitry, and extremely low cost make this electrometer design attractive for general purpose experiments.

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6 COSMOS Integrated Circuit, RCA Corp. 1978.