# An Ultra-Low Drift dc Chopper Amplifier Using MOSFET for Large Value of Signal-Source Resistance

MINORU ABE, JUN NAKAZOE, MEMBER, IEEE, AND ZENMON ABE

*Abstract*—This paper describes an ultra-low drift, MOSFET, and dc chopper amplifier which can be used for signal-source resistance of 0.1 ~ 10 MΩ. The sensitivity of this chopper amplifier decreases with the increase in signal-source resistance. We found that the decrease in the sensitivity is caused mainly by the drain-substrate and source-substrate capacitances in the MOSFET, and corrected the problem by the bootstrapping method. When using bootstrapping, these capacitances are reduced, and the temperature drift due to high signal-source resistance is improved. For an input signal source of 10 MΩ, the drift is reduced to one-tenth of its original value. Thus we were successful in building a dc amplifier with a temperature coefficient of within ±10 nV/°C for a signal-source resistance of 0.1 ~ 10 MΩ.

## I. Introduction

A CHOPPER amplifier is usually used to amplify a microvolts order dc signal. Since the MOSFET has a lower leakage current between the gate and drain, and between the gate and source, a MOSFET is better suited for a chopper element than a JFET.

In the conventional MOSFET chopper amplifier, the amplification limit is set by the spike output voltages from gating signals which are produced by the capacitance of the chopper element. These unwanted output spikes are amplified and demodulated by an ac amplifier and a phase-sensitive detector (PSD), respectively. The unwanted output becomes an offset voltage and produces "zero drifting" owing to ambient temperature variations. Because of these reasons, it was difficult to build a stable dc amplifier using the MOSFET.

It is considered that zero drift is caused by small changes in the interelectrode capacitance of the MOSFET. In order to reduce the spike voltage, we examined the possibility of building an ultra-low drift chopper amplifier which has a shunt switch circuit operating at twice a frequency of gating signal [1]. The new chopper amplifier has an input offset voltage temperature coefficient of 0.1 nV/°C for a signal-source resistance of less than 1 kΩ, and a long-term drift of 2 nV/18 h for a signal-source resistance of 6.1 Ω.

However, the chopper amplifier for voltage amplification should have a stable performance for high signal-source resistance (≃ 10 MΩ). The above-mentioned amplifier did not provide satisfactory results for such purposes. In this paper we analyzed the operation of the MOSFET dc chopper amplifier for high signal-source resistance. In particular we focused on the decrease in sensitivity and zero drift due to the interelectrode capacitances in the MOSFET. As a corrective measure, the capacitances were reduced by bootstrapping and we designed an ultra-low drift chopper amplifier for high signal-source resistance.

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The authors are with the Department of Electronics and Communication Engineering, Faculty of Engineering, Musashi Institute of Technology, Tokyo, Japan.

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## II. Circuit and its Basic Operation

Fig. 1 is the block diagram of the series-shunt chopper amplifier with a switching circuit for spike compensation. The one-pole two-position switch is connected to line "Sub." An "$A$" is the normally closed switch position and a "$B$" is the open position. The "$A$" position is referred to as "Bootstrapped" and "$B$" as "Non-bootstrapped." The small dc signal applied from $V_S$ is converted into a square wave $V_0$ (chopped signal) by a series-shunt chopper $Q_1$ and $Q_2$. The square wave $V_0$ is fed to the drain of an additional shunt switch $Q_3$, which is driven by the gating signal $V_{g3}$ at twice the frequency of the gating signals of $Q_1$ and $Q_2$. The waveform during the operation is shown in Fig. 2. During the period when spikes are produced by gating signals $V_{g1}$ and $V_{g2}$ due to the interelectrode capacitance, the circuit is shunted by $Q_3$ and the magnitude of these spikes decreases. However, new spikes due to the gating signal $V_{g3}$ are generated by the gate-source capacitance of $Q_3$. These spikes are then reduced, mainly by the compensation capacitor $C_C$, and the compensation voltage $V_{g4}$. The fundamental frequency component of the remaining spikes are twice the frequency of the gating signal. In addition, it contains waves of integer multiples of the fundamental frequency. By using the filtering action of selective amplifier $A_2$, and by the frequency/phase selective action of the PSD, the pulses developed by $Q_3$ do not appear at the output. Thus we were able to construct the low-drift amplifier described in the report [1]. As shown in Fig. 2, the output signal of the chopper is produced when $V_{g3}$ is low and $Q_3$ is off during the $(T/2 - t_{on})$ period. In Fig. 3 an equivalent circuit is shown for the period when $Q_1$ is closed and $Q_3$ changes from closed to open, and then changes again from open to closed after $(T/2 - t_{on})$. In this equivalent

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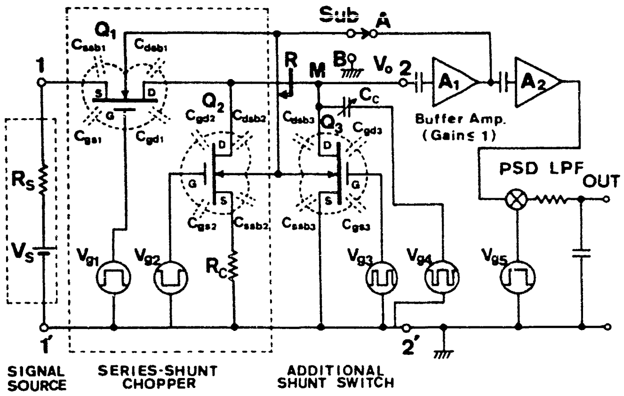


Fig. 1. Schematic diagram of the new type of series-shunt chopper amplifier. ($R_{on1}$, $R_{on2}$, and $R_{on3}$ are the channel resistance when $Q_1$, $Q_2$, and $Q_3$ are in the ON state, respectively. $C_C$ and $V_{g4}$ compensate the spike introduced by $V_{g3}$ at the point $M$. $A_2$ is the selective amplifier for the fundamental frequency of gating signals $V_{g1}$ and $V_{g2}$.)

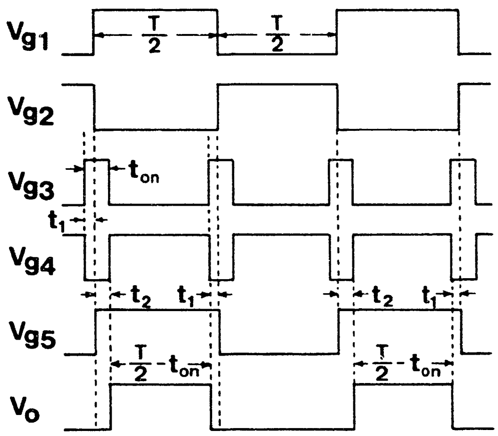


Fig. 2. Timing diagram of the gating signals in Fig. 1 and the chopped signal $V_0$. (Each MOSFET is in the ON state when corresponding gating signal is high level.)

circuit, because ac components in $V_0$ are considered, the coupling capacitance to the ac amplifier can be ignored. $C_T$ represents parallel capacitances created by each drain-substrate and source-substrate capacitances of $Q_1$, $Q_2$, and $Q_3$, and the input capacitance of the ac amplifier.

On the other hand, $R_T$ represents parallel resistance of a series resistance of the off-state resistance of $Q_2$, compensation resistance $R_C$, and the input resistance $R_a$ of the ac amplifier $A_1$. All of these appear at point $M$ and the ground, which are between terminals 2-2'. When $Q_1$ is in the on state, $R_{on1}$ is very small compared with $R_S$ and is negligible. The time constant of the circuit, when $Q_3$ is open and the voltage between the terminals 2-2' increases, is $C_T(R_S//R_T)$. On the other hand, when $Q_3$ is closed and the voltage drops, and assuming the resistance of $Q_3$ when it is ON to be $R_{on3}$, the time constant becomes $C_TR_{on3}$. The voltage output $V_0$ becomes distorted by these time constants and results in the distortion of square waves, $V_0$ in the fundamental frequency component of their amplitudes and phases. Accordingly, the output from the PSD decreases and the sensitivity of the chopper decreases.

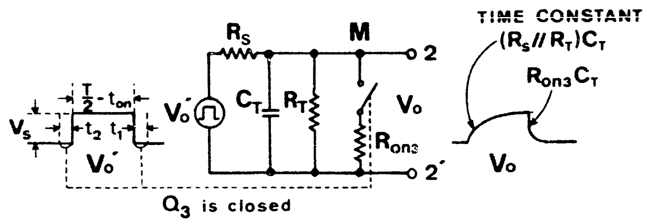


Fig. 3. Equivalent circuit of the chopper in Fig. 1.

Let us define the ratio of terms of Fourier expansion of square wave $V_0$ at the fundamental frequency component $f = 1/T$ to the dc signal $V_S$ as the chopper efficiency $\eta_f$. Assuming the time constant $C_TR_{on3}$ to be zero because of the condition $C_T(R_S//R_T) >> C_TR_{on3}$, then $\eta_f$ can be given as

$$\eta_f = \frac{R_T}{R_S + R_T}\left|\frac{j}{2\pi}(1 + e^{-j\omega t_{on}}) + \frac{1}{\frac{T}{\tau} + j2\pi} \cdot \{e^{-(T/2 - t_{on})/\tau} + e^{-j\omega t_{on}}\}\right| \tag{1}$$

where $\tau = C_T(R_S//R_T)$ and $\omega = 2\pi/T$.

The first condition to keep $\eta_f$ high is to set $R_T >> R_S$, to make $R_T/(R_S + R_T) \simeq 1$. When the switch is in "$B$" position as in Fig. 1 this circuit causes distortion of $V_0$ by $C_T$. We can reduce distortion by making $C_T$ small. As can be seen from Fig. 1, when $Q_1$ is closed and $Q_2$ and $Q_3$ are open, the exact $C_T$ is given by the following equation excluding stray capacitances:

$$C_T = (C_{SSb1} + C_{dSb1} + C_{dSb2} + C_{dSb3}) + (C_{gd1} + C_{gd2} + C_{gd3}) + C_C + C_a \tag{2}$$

where the first term on the right-hand side represents the capacitance $C_{Sb}$ related to the substrate, the second term is related to gate-drain electrodes, and $C_a$ is the input capacitance of the ac amplifier. $C_{Sb}$ is the largest capacitance in $C_T$ and is approximately 10 times larger than the second term. In order to reduce the capacitance $C_{Sb}$, the bootstrapping was applied to the substrate to achieve the same ac signal which appears at point $M$ (the switch is in "$A$" position in Fig. 1). The feedback, where a part of the output voltage from the amplifier $A_1$ in Fig. 1 is fed to the line "Sub," reduces $C_{Sb}$ and results in $C_{Sbe} = (1 - A_1)\, C_{Sb}$. By setting $A_1 \leq 1$, $C_{Sbe}$ can be reduced.

On the other hand, as $C_{Sb}$ gets reduced by bootstrapping, the distortion of spikes generated at point $M$ by gating signal $V_{g3}$ due to the variation of $C_{Sb}$ becomes small. $V_{g3}$ produces spike (spike $A$) at point $M$ through $C_{gd3}$ in the circuit of Fig. 3 with $V_0'$ and $R_s$ being replaced by $V_{g3}$ and $C_{gd3}$, respectively, and attachment of $R_S$ in parallel to $R_T$. In this case, since $C_T >> C_{gd3}$ and $R_T >> R_S$ (discussed later), the time constant of the spike generating circuit is $C_TR_S$. Now when $Q_1$ is OFF and $Q_2$ is ON, by replacing $R_S$ and $R_C$, the time constant for the spike generating circuit (spike $B$) is $C_T'R_C$ where $C_T'$ is the capacitance of $C_T$ in (2) and where $C_{SSb1}$ is replaced by $C_{SSb2}$. The wave shapes of these spikes generated by $V_{g3}$ are mostly determined by the time constants $C_T R_S$ and $C'_T R_C$.

Let us note that $\Delta C_T = C_T - C'_T$ and $\Delta C_T = C_{SSb1} - C_{SSb2}$. Since $R_S = R_C$ in the present chopper, which render $R_{on1}$ and $R_{on2}$ negligible in comparison [1], by the condition $\Delta C_T \neq 0$ the shapes of spike $A$ and spike $B$ are different from each other. Then the frequency component with period $T$ develops in the wave consisting of spike $A$ and spike $B$ and causes offset voltage. Although temperature variation may produce a change in $\Delta C_T$, the change becomes reduced by $(1 - A_1)$ times. By setting $A_1 \lesssim 1$ the variation of the difference in the wave shape between spike $A$ and spike $B$ becomes small and the variation of offset voltage also becomes small. Thus bootstrapping reduces temperature drift.

In general, the input impedance of an amplifier should be far larger than signal-source resistance. Unfortunately, the conditions of the shunt-chopper amplifier do not satisfy this requirement. During a half period, the input terminals are shunted by the ON-state resistance $R_{on}$ of the chopper. Actually the input resistance becomes $R_{on}/2$. Nevertheless, it can be used for the high signal-source resistance. This present chopper has the same property. Namely, when $Q_1$ closes, as shown in Fig. 2, the input terminals of the chopper are shunted by the ON-resistance of $Q_3$ for $t_{on}$. If this circuit configuration is not desirable for the signal source under consideration, it is possible to install an additional series-shunt switch instead of an additional shunt switch $Q_3$.

## III. Experimental Results

In the circuit of Fig. 1, when the series chopper of $Q_1$ is closed, the input resistance of the chopper should be far larger than signal-source resistance $R_S$. The above-mentioned $R_T$ is considered to be the input resistance. $R_T$ is the parallel connection of the OFF-state resistance ($\simeq 10^{11}$ $\Omega$) for $Q_2$ and $Q_3$, and the input resistance ($R_a \simeq$ several 100 M$\Omega$) of the ac amplifier $A_1$, so for $R_S =$ several 10 M$\Omega$, $R_S << R_T$.

### A. Sensitivity

In the circuit of Fig. 1, when the switch connected to "Sub" is in "$B$" position (referred to as "Nonbootstrapped"), $C_T = 50$ pF. As shown in Fig. 4, the decrease in sensitivity is seen when $R_S$ is substantially smaller than $R_T$. To prevent this decrease in sensitivity, it is necessary to decrease $C_T$.

When the switch is in "$A$" position (referred to as "Bootstrapped"), the capacitance generated by the substrate $C_{Sb}$ will be multiplied by $(1 - A_1)$. Since $A_1 = 0.98$, it will be $\frac{1}{50}$ of the original value. In MOSFET 3SK38A, the drain-substrate and source-substrate capacitances are 9 pF each, the gate-source and gate-drain capacitances are 1 pF each, $C_C$ is several pF, and the input capacitance of the ac amplifier $A_1$ is approximately 10 pF. Accordingly, from (2) $C_T \simeq 50$ pF, $C_{Sb} \simeq 36$ pF, and

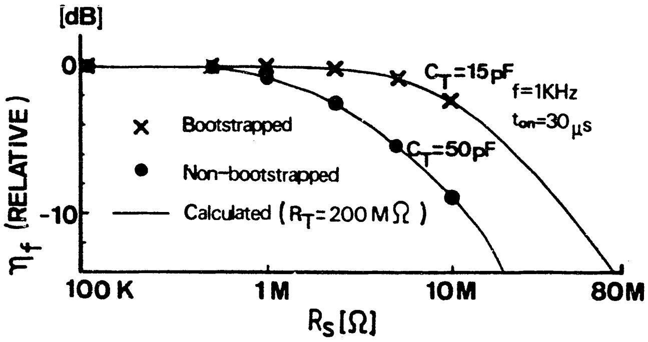


Fig. 4. Chopper efficiency $\eta_f$ versus signal-source resistance $R_S$.

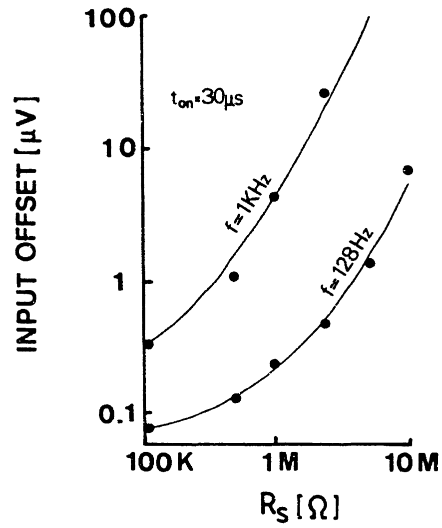


Fig. 5. Input offset voltage versus signal-source resistance $R_S$.

$C_{Sb}$ can be found to be the largest capacitance in $C_T$. By using bootstrapping $C_{Sb}$ decreases to 0.7 pF which is $\frac{1}{50}$ of the original value. The chopper efficiency characteristics under these conditions are shown in Fig. 4.

When $R_S$ is compared with "Bootstrapped" and "Non-bootstrapped" circuits at $\eta_f = -6$ dB, the $R_S$ was four times larger for "Bootstrapped" than for "Non-bootstrapped." The upper value of $R_S$ is mostly limited by the input capacitance of $C_a$ of the ac amplifier. $R_S$ can be increased further by reducing $C_a$.

### B. Input Offset Voltage

In the shunt switch of Fig. 1, if $R_S$ and $R_C$ are increased, the spike shapes generated by $Q_3$ tend to receive the influence of the capacitance which are connected in parallel to $R_S$ and $R_C$. When these two capacitances are not equal to each other, the spikes generated by $Q_3$ contain the frequency components of period $T$. These spikes are demodulated by PSD and become the source of offset voltage. As $R_S$ gets larger, the offset will increase, and higher gating frequency produces larger offset voltages (Fig. 5). The reason for this is that as the area of spikes is constant, the percentage of the gating signals sent to the area in one

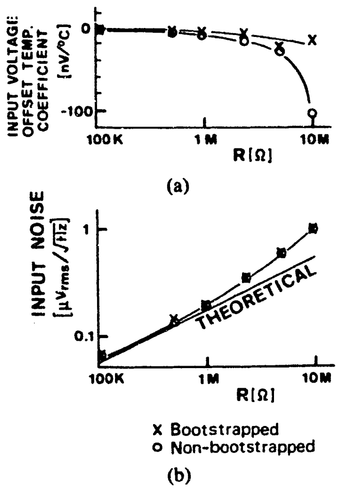


Fig. 6. Noise and drift in the chopper with additional shunt switch. (a) Input noise versus signal-source resistance $R_S$. (b) Input voltage offset temperature coefficient versus signal-source resistance $R_S$.

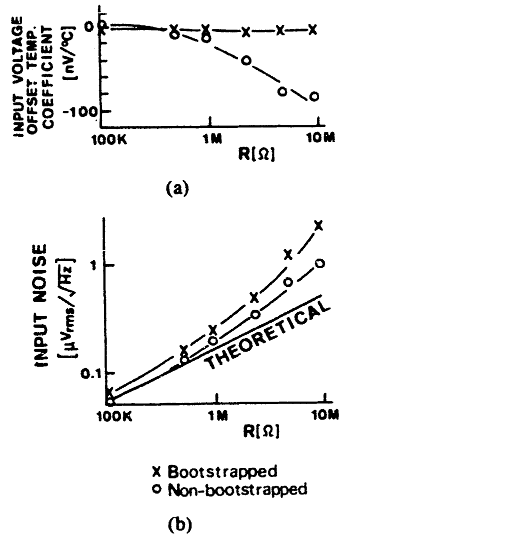


Fig. 7. Noise and drift in the chopper with additional series-shunt switch. (a) Input noise versus signal-source resistance $R_S$. (b) Input voltage offset temperature coefficient versus signal-source resistance $R_S$.

period increases. So the following experiments are conducted at low gating frequency (= 128 Hz).

### *C. Noise and Drift*

In Figs. 6 and 7, noise and drift characteristics of a shunt-switched chopper and a high-input resistance series-shunt switch chopper are shown. As shown in Fig. 7(a), when bootstrapping was applied the effect of current type noise was observed for the $R_S$ as being larger than 5 MΩ. The value of the current is $i_n \simeq 2.5 \times 10^{-13}$ $A/\sqrt{\text{Hz}}$. After a proper warming up, we measured the input voltage offset temperature coefficient for $R_S$ (Figs. 6 and 7(b)).

According to our measurements, as the value of $R_S$ increases the effect of bootstrapping on temperature coefficient becomes eminent. At $R_S$ = 10 MΩ, it becomes one-tenth of the original drift. For the series-shunt switched chopper amplifier, the input voltage offset temperature coefficient was reduced to within ±10 nV/°C for $R_S$ of 0.1 ~ 10 MΩ.

## IV. Conclusion

To apply the ultra-low drift chopper amplifier using MOSFET [1] for the use of high signal-source resistance, we used the bootstrapping method and obtained the following results.

1) The major cause in the decrease of sensitivity of a chopper amplifier, the drain-substrate and source-substrate capacitances, were reduced to a negligible level compared with the input capacitance of the following ac amplifier.

2) The cause of temperature drift of the chopper amplifier and source-substrate capacitance was reduced by bootstrapping. The temperature drift was improved and was reduced to one-tenth of the original value at $R_S$ = 10 MΩ. We developed a dc amplifier with an input voltage offset temperature coefficient of within ±10 nV/°C for a signal-source resistance of 0.1 ~ 10 MΩ.

The authors are studying the application of this chopper for differential amplifiers.

## Reference

[1] M. Abe, I. Sugisaki, J. Nakazoe, and Z. Abe, "An ultra-low drift amplifier using a new type of series-shunt MOSFET chopper," *IEEE Trans. Instrum. Meas.*, vol. IM-34, pp. 54-58, Mar. 1985.