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Put FETs to Work in Electrometers

Field-effect transistors, with their high impedance and low noise, can effectively measure very low currents. To bring the FET up to electrometer standards (so that it can measure currents of, say, 10^{13} amp) the designer must know where the noise is and what to do about it.

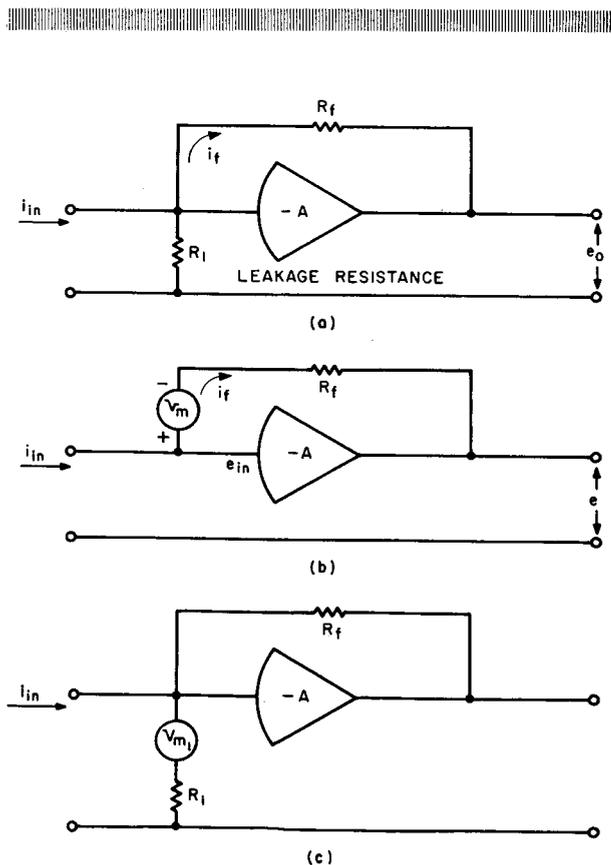
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MANY MEASUREMENTS in physics, chemistry and industrial process control involve very small currents (less than 10^{-10} amp). Until the advent of the field-effect transistor (FET), the only way to achieve the high input impedance and low noise level for a pico-ampere current measurement was to use a high-grade electrometer type of vacuum tube in a carefully constructed laboratory circuit.

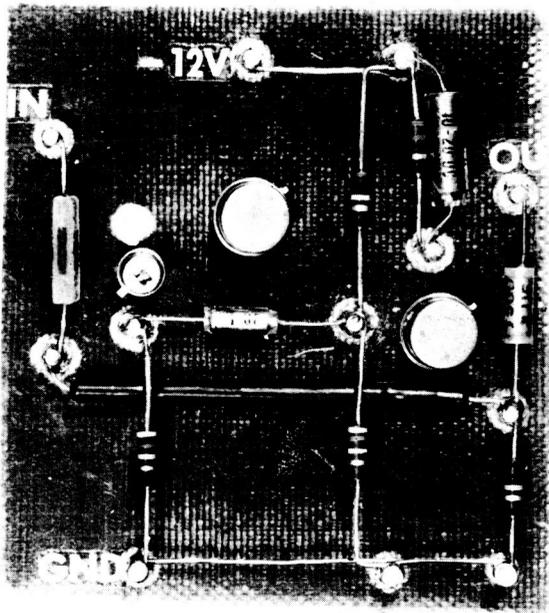
FETs can perform as well as, and even better than, vacuum tubes in electrometer-type instrumentation. They do, however, need careful circuit design.

Fig. 1 shows one way to measure very low currents, by capitalizing on the high input impedance of an FET. Here the current being measured at the input is transformed by an operational amplifier and feedback resistor setup into a sizable voltage level at the output. The FET is used as the input stage for the operational amplifier. With this arrangement, and using commercially available FETs, currents as low as 10^{-13} amp can be measured over a 1-cps bandwidth.

How far can this scheme be pushed and what are the design considerations for the operational amplifier? In the ideal case, in which the operational amplifier is assumed to have an infinite input leakage resistance, the only limitation to the system's sensitivity would be due to the thermal noise from the feedback resistor R_f . It will be shown that



1. **Feedback amplifier scheme** (a) is ideal for converting small currents to measurable voltages. Its sensitivity is determined by two noise sources, the feedback resistor's thermal noise (b) and the amplifier's input leakage noise (c).



when only this source is considered, the signal-to-noise ratio is proportional to the square root of the feedback resistance. Fig. 1b models the ideal situation by including a thermal-noise voltage generator, v_n , in series with the feedback resistor R_f . A nodal equation for currents at the amplifier's input yields:

$$i_{in} = i_f \quad (1)$$

Substituting R_f 's branch voltage-amperage relationship for i_f , we get:

$$i_{in} = \frac{e_{in} - v_n - e_o}{R_f} \quad (2)$$

Writing e_{in} in terms of e_o and solving for e_o we get:

$$e_o = \frac{-i_{in} R_f - v_n}{(1 + 1/A)} \quad (3)$$

This basic relationship demonstrates that, if the noise is negligible, the output voltage is a direct measure of the input current and is related to the transresistance R_f . Therefore, as we have already indicated, if R_f can be made very large, the circuit of Fig. 1 provides a feasible technique for measuring small currents.

How Noise Affects The Basic Scheme

But what happens to the noise when we make R_f large? The noise voltage v_n increases but, fortunately, the noise current

that interests us most decreases. This can be seen when the signal-to-noise voltage ratio is developed from Eq. 3.

$$S/N = \frac{i_n R_f}{v_n} \quad (4)$$

Here the signal-to-noise voltage ratio is defined as the output rms voltage due to current input compared with the output voltage due to noise input. This ratio is more meaningful than the signal-to-noise power ratio when the objective of the circuit is to measure current.

The value for the rms thermal noise voltage, v_n , is given by:

$$v_n = (4KTBR_f)^{1/2} \quad (5)$$

where:

K = the Boltzman Constant, 1.38×10^{-23} Joules/deg K

T = absolute temperature in K

B = bandwidth in cps

and the S/N ratio becomes:

$$S/N = i_{in} \left[\frac{R_f}{4KT B} \right]^{1/2} \quad (6)$$

Surprisingly, the noise performance of this feedback scheme improves with increasing feedback resistance, as shown by Eq. 6. If temperature and bandwidth are fixed, the signal-to-noise ratio is directly proportional to the square root of the feedback resistance. However, in practical circuits, leakage resistance limits the degree to which the signal-to-noise ratio can be obtained by increasing the feedback resistance.

In actual amplifiers—even those using FETs in their input stages—there will be some leakage.* The noise associated with the leakage into the operational amplifier must also be considered. Fig. 1c shows the equivalent circuit of the operational amplifier when the thermal noise voltage associated with this

*The newer MOS (Metal-Oxide-Semiconductor) FETs may have even lower leakage currents, because their gates are almost completely insulated from the main channel being controlled. Both devices build up their control voltage fields with capacitor-like gate structures. But whereas the back-biased diode used by the ordinary ET has a certain amount of leakage, the true capacitor structure of the MOS's gate is separated from the main channel by a near-perfect insulator. However, since the MOS field-effect transistors are surface-operated devices, some experts feel they will always have higher noise figures than the earlier diode-gate FETs.

leakage resistance is considered but the noise associated with the feedback amplifier is ignored. If the input current is zero, the noise voltage appearing at the output can, by a process similar to that used for the feedback resistor noise, be represented by:

$$e_{o1} = v_{n1} \frac{R_f}{R_1} \left(\frac{1}{1 + 1/A} \right) \quad (7)$$

This shows that the output noise due to leakage varies inversely as the leakage resistance, R_1 .

Since the system is linear, the principle of superposition holds, and the total noise contributed by each of the processes can be found by addition of the noise powers. However, we are really interested in output voltage, not output power, so we must first convert voltages to power, take the sum, and then convert back to voltage. This may be done by taking the square root of the sum of the squares of the voltages in question. The total noise from the two thermal-noise resistance sources R_f and R_1 then is:

$$e_{o \text{ total noise}} = \frac{1}{(1 + 1/A)} \left(v_n^2 + v_{n1}^2 \frac{R_f^2}{R_1^2} \right)^{1/2} \quad (8)$$

Introducing the parameters for the thermal noises, v_n and v_{n1} ,

$$e_{o \text{ total noise}} = \frac{1}{(1 + 1/A)} \left[4KTBR_f \left(1 + \frac{R_f}{R_1} \right) \right]^{1/2} \quad (9)$$

Then the total signal-to-noise ratio becomes:

$$S/N = i_{in} \left[\frac{R_f}{(4KT B) \left(1 + \frac{R_f}{R_1} \right)} \right]^{1/2} \quad (10)$$

This is a most important result, since it describes how leakage resistance R_1 deteriorates the performance of the current-measuring system. Since the FET has an inherently high leakage resistance, good signal-to-noise ratios can be achieved at very low current levels.

Thermal noise originating in the drain circuit of the FET has been neglected in this analysis. The current gain of the FET is so large that the thermal noise of the input predominates. Recombination noise is negligible in the drain circuit because the FET is a majority-carrier device.¹

FET's Qualities Accented By Amplifier Circuit

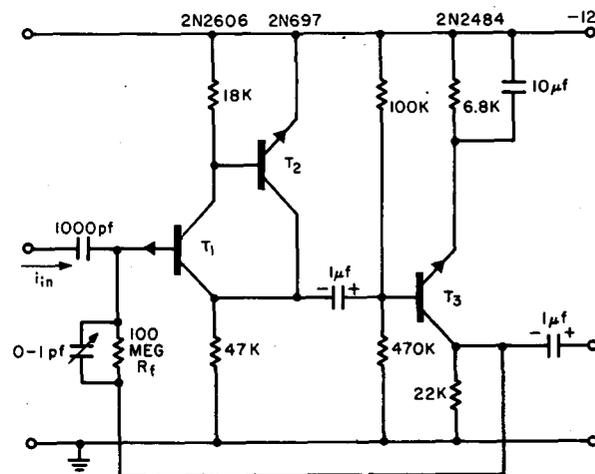
Fig. 2 shows a circuit for the operational-amplifier that optimizes the noise performance of the FET in several ways. First, the

circuit is relatively wideband so that the user has reasonable latitude in employing additional filters to tailor his instrumentation bandwidth for minimum noise. Second, there are no unnecessary noise-producing resistors in the input circuit. The FET's gate is biased by a feedback circuit.

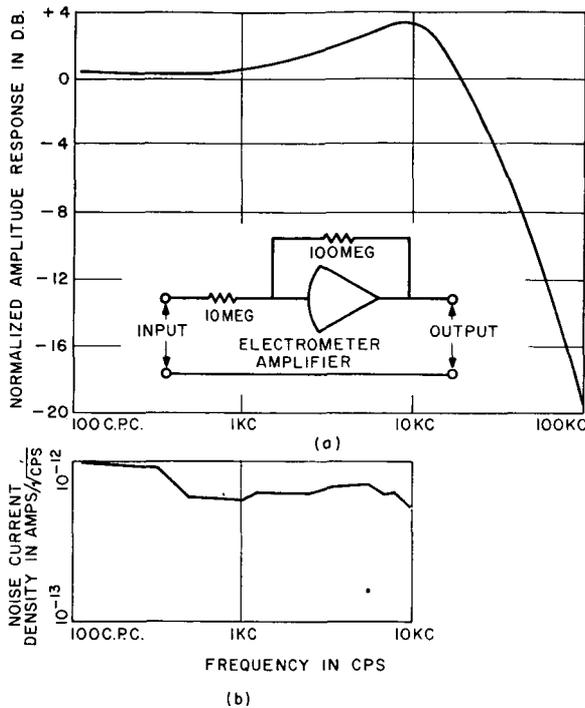
All transistors suffer from 1/f noise and the FET is no exception. However, above 1 Kc this type of noise is rarely a problem.^{2, 3} For this reason the circuit in Fig. 2 is ac-coupled. Since the external filtering would probably cut the response down to a narrow bandpass anyway (which because of 1/f noise would remain above the very low frequencies), this is not a drawback.

Simplicity was another design objective. The first active element, FET T_1 , is paired with a conventional bipolar transistor T_2 in a close-coupled feedback arrangement. This has several advantages. It permits in-phase operation so that with only one additional transistor stage, T_3 , the over-all amplifier will have the proper phase inversion for operational performance.

Transistor T_2 cancels the FET's source-to-gate capacitance and at the same time lessens the gain loss due to the FET's source-follower configuration. This configuration is, of course, desirable to maximize its input impedance. T_2 raises the pair's gain by ap-



2. **Electrometer amplifier** for the Fig. 1 feedback scheme uses a field-effect transistor at its input. The FET's ability to maintain a low noise while providing a high input impedance is further optimized by the source-follower arrangement. The conventional bipolar transistor, T_2 , helps to minimize the FET's source-to-gate capacitance. The feedback circuit from output to the FET's gate also supplies bias, thus eliminating input resistor bias networks—another source of noise.



3. **Frequency response of the amplifier** when used in a gain-of-10 circuit (a) and noise characteristics (b). In practice, as much of the lower frequencies under 1 Kc as possible would be cut off to eliminate the $1/f$ noise. The lower graph indicates that the noise levels of the amplifier are low enough over 100 cps to permit it to see 10^{-13} amp.

proximately 0.6, producing a very nearly unity gain. This reduces the equivalent capacitance between the gate and the source. The circuit's input time constant is now limited only by the gate-to-drain capacitance. Maximum bandwidth is achieved, since the input time constant is limited only by the gate-to-drain capacitance.

Fig. 3a shows the frequency response of the circuit when it is used as an operational gain-of-10 amplifier (a 10-Meg input resistance and a 100-Meg feedback resistance). Fig. 3b shows the amplifier's noise-current density, plotted as a function of frequency. It demonstrates that for presently available FETs, it is possible to measure currents as low as 10^{-13} at room temperatures, over 1-cps bandwidths. When higher-purity materials and FETs with higher leakage resistances become available, even greater sensitivities should be attainable. ■ ■

References

1. A. Van der Ziel, "Thermal Noise in Field-Effect Transistors," *Proc. IRE*, Vol 50, pp 1808-1812, August 1962.
2. Staff, Texas Instruments Inc., *Transistor Circuit Design*, McGraw-Hill 1963. (A curve of spot noise is shown on p. 513).
3. Dewitt and Rossoff, *Transistor Electronics*, McGraw-Hill 1957 (see p. 342).

ERRATA
 FOR ARTICLE,
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 ELECTRONICS DESIGN, MAY 11, 1964

<u>PAGE</u>	<u>COLUMN</u>	<u>ERROR</u>	<u>CORRECTION</u>
1	Abstract, Line 3	10^{13}	10^{-13}
1	1 Figure 1, B	V_m	V_n
1	1 Figure 1, B	e	e_o
1	1 Figure 1, C	V_{m_1}	V_{n_1}
1	1 Figure 1, C	-----	e_{o_1}
2	2 Equation 5	Boltsman	Boltzmann
2	2 Footnote, Lines 4 & 5	devides	devices
2	2 Footnote, Line 7	ET	FET
3	1 Line 22, following equation (8)	Introducing the parameters for thermal noises V_n and V_{n_1}	Substituting the thermal noise values computed in equation (5) we have:
3	2 Last line	t_2 raises the pairs gain by approximately 0.6 producing a very nearly unity gain.	The bipolar transistor t_2 improves the gain of the input source follower circuit by feedback from approximately +0.6 to very nearly +1.
4	1 Figure 3B	10^{-13} 10^{-12}	10^{-14} 10^{-13}