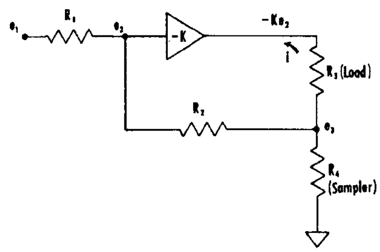

Fig. 1—One form of constant current generator.

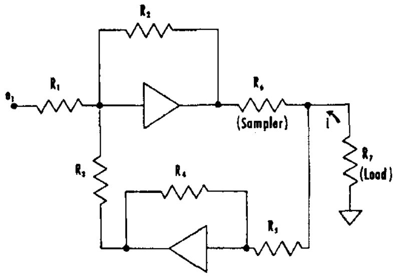

Fig. 2—Alternate form of constant current generator for grounded load.

The current actually flowing in the load, given as a function of the control voltage, is

$$\frac{i}{e_i} = \frac{KR_2+(K+1)R_4}{R_1R_3+(K+1)R_1R_4+R_2R_3+R_2R_4+R_3R_4}. \quad (1)$$

For $K=\infty$, this reduces to

$$\frac{i}{e_1} = \frac{R_2+R_4}{R_1R_4}, \quad (2)$$

which is not a function of R_3.

This shows that for an infinite-gain amplifier, the load current is independent of load resistance, and the source exhibits an effective output impedance of infinity.

This is a very useful scheme, but it has the disadvantage that the load must be lifted above ground by the sampling resistor, R_4. In addition, it should be noted that for $K<\infty$, there is no combination of resistance values which will yield an infinite output impedance.

If the load and sampling resistors are interchanged, as shown in Fig. 2, the load may be operated with one side grounded, or connected to any arbitrary point, as long as amplifier output capabilities are not exceeded.

Analysis of Fig. 2, again on the basis of infinite-gain amplifiers, shows the actual load current to be related to control voltage by

$$\frac{i}{e_1} = \frac{R_2R_3R_5}{R_1R_3R_5R_6+R_1R_7(R_3R_5+R_3R_6-R_2R_4)}. \quad (3)$$

The second term in the denominator may be made equal to zero by setting

$$R_4 = \frac{R_3(R_5+R_6)}{R_2}, \quad (4)$$

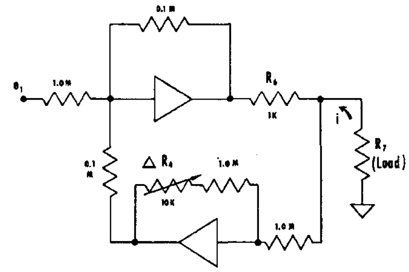

Fig. 3—Constant current source as patched on the analog computer.

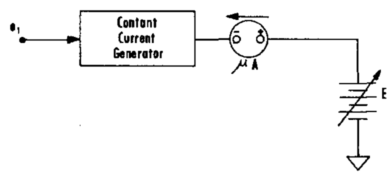

Fig. 4—Measurement of output impedance.

in which case

$$\frac{i}{e_1} = \frac{R_2}{R_1R_6}. \quad (5)$$

This last expression is not a function of R_7, the load resistance, so that the source exhibits an effective output impedance of infinity. It is of interest to note that even for finite-gain amplifiers, there is a value of R_4 which yields infinite output impedance.

If R_4 is made larger than the value given by (4), the output impedance becomes negative. At the same time, the loop gain will exceed unity unless the load resistance, which is part of a voltage divider, is kept below a critical value. For

$$R_4 > \frac{R_3(R_5+R_6)}{R_2}, \quad (6)$$

stability requires that

$$R_7 \leq \frac{R_3R_5R_6}{R_2R_4 - R_3(R_5+R_6)}. \quad (7)$$

The circuit is patched on the computer as shown in Fig. 3. For this circuit,

$$\frac{i}{e_1} = \frac{1}{10R_6} = 10^{-4}, \quad (8)$$

yielding full scale variables of ± 100-v input and ± 10-ma output. The value of R_4 was made variable, to compensate for inaccuracies in the other resistors and the noninfinite amplifier gains.

The output impedance was measured as shown in Fig. 4. The voltage, E, was varied, and the change in current measured on the microammeter. The output impedance at dc is then given by

$$R_0 = \frac{\Delta E}{\Delta I}. \quad (9)$$

With a stable, high-resolution rheostat at ΔR_4, a stable output impedance in excess of 100 megohms is easily obtainable.

The realization of negative resistance is easily demonstrated by a further increase in ΔR_4. A positive-going change in E will then result in an increase of current *out* of the source. Stability is ensured by the low value of load resistance, made up of the resistances of the microammeter and battery in series.

Eq. (5) may be rewritten as follows to show operation as a current amplifier:

$$i = \frac{e_1}{R_1}\cdot\frac{R_2}{R_6} = i_{in}\cdot\frac{R_2}{R_6}.$$

An input impedance of zero is realized for infinite-gain amplifiers, and the output impedance will be infinite as before.

The compliance of a current source is defined as the voltage which may be developed across the load without loss of current control, and is roughly analogous to the current rating of a voltage-regulated supply. In this circuit, the compliance is just the output voltage capability of the amplifier minus the drop across R_6, the sensing resistor.

The bandwidth over which the high output impedance will hold is determined by the amplifiers, the reactive components of the precision resistors used, and the care taken in layout and patching. Conventional equipment and techniques yield more than adequate bandwidth for ordinary simulations.

The input impedance for the control voltage is just the value of R_1, or, in this case, one megohm.

Although the circuit requires an extra amplifier, it permits returning the load directly to ground or to any desired point. In addition, the provision for adjustment of output impedance, including negative values, may be very useful.

R. W. Thorpe
General Dynamics/Pomona
Pomona, Calif

Circuits Using Tunnel Diode Flip-Flops and *PNPN* Diodes*

Two well-known and fundamental solid-state negative resistance devices are the tunnel diode (Esaki diode), and the *pnpn* diode (Shockley diode).

Proper utilization of these devices in switching-type circuits reduces the complexity of the system, increases the efficiency, improves the speed and reduces the size. The bilateral behavior of such two-terminal devices is partially compensated by the driving stage.

A basic bistable flip-flop circuit using a tunnel diode and actuated by a positive pulse has been explained previously by the author.[1] With a modification of the driving network this circuit can be triggered with negative pulses. Such a circuit is shown in Fig. 1.

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[1] V. Uzunoglu, "A bistable flip-flop circuit using tunnel diode," Proc. IRE (*Correspondence*), p. 1440; September, 1961.