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TECHNICAL REPORT

"A SOLID-STATE LOW-NOISE PREAMPLIFIER"

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A SOLID-STATE LOW-NOISE PREAMPLIFIER

An all-solid-state, voltage-sensitive preamplifier with low noise and broadband response has been developed for use with an electrostatic hypervelocity accelerator employed in research on hypervelocity impact phenomena.[1,2,3] A large effective resistance and small effective capacitance have been achieved at the input stage by use of a bootstrap loop incorporating a field effect transistor as the input stage. In the hypervelocity test system the preamplifier is used in conjunction with a detector[2,4] to measure particle velocity and charge, but its characteristics make it suitable for other applications as well. The equivalent input resistance of the preamplifier is 10^{10} ohms, and the bootstrapping reduces the effective input capacitance to about 1 pf. For a pass band of 10 kc to 10 Mc the equivalent rms input noise charge is about 250 electrons.

In the hypervelocity test system[3] the detector is a cylindrical tube which is traversed axially by a charged dust particle in a time Δt that is inversely proportional to particle velocity. Electrically, the detector may be considered a capacitance to ground upon which at some instant of time a charge q (the charge of the dust particle) is placed, and from which after a time Δt is removed. The resulting voltage signal is a rectangular pulse of length Δt and of amplitude

$$V_{in} = \frac{q}{C_{eff}}, \qquad (1)$$

where C_{eff} is the effective input capacitance. The pulse has a droop of time constant τ, where τ is given by

$$\tau = R_{eff}C_{eff}, \qquad (2)$$

R_{eff} being the effective input resistance. In actuality, a finite amount of time is required to deposit and to remove the charge q, hence the rise and fall times of the pulse are also finite. The values of these times necessitate a high-frequency response of 10 mc. For analysis of the preamplifier pulse, the low-frequency response is extended to 10 kc.

From Eq. (1) it is clear that the effective input capacitance should be as low as possible, providing, of course, that any technique used to lower the capacitance also improves the signal-to-noise ratio. The desired decrease can be achieved by bootstrapping the input capacitance. It is shown in the appendix that the effective input capacitance with bootstrapping is given by

$$C_{eff} = C_d \ (1 - A) \quad , \tag{3}$$

where A is the open-loop gain of the bootstrap loop and C_d is the input capacitance.

However, an undesirable effect accompanies this reduction of input capacitance. The r se time at the output of the bootstrap loop is given by

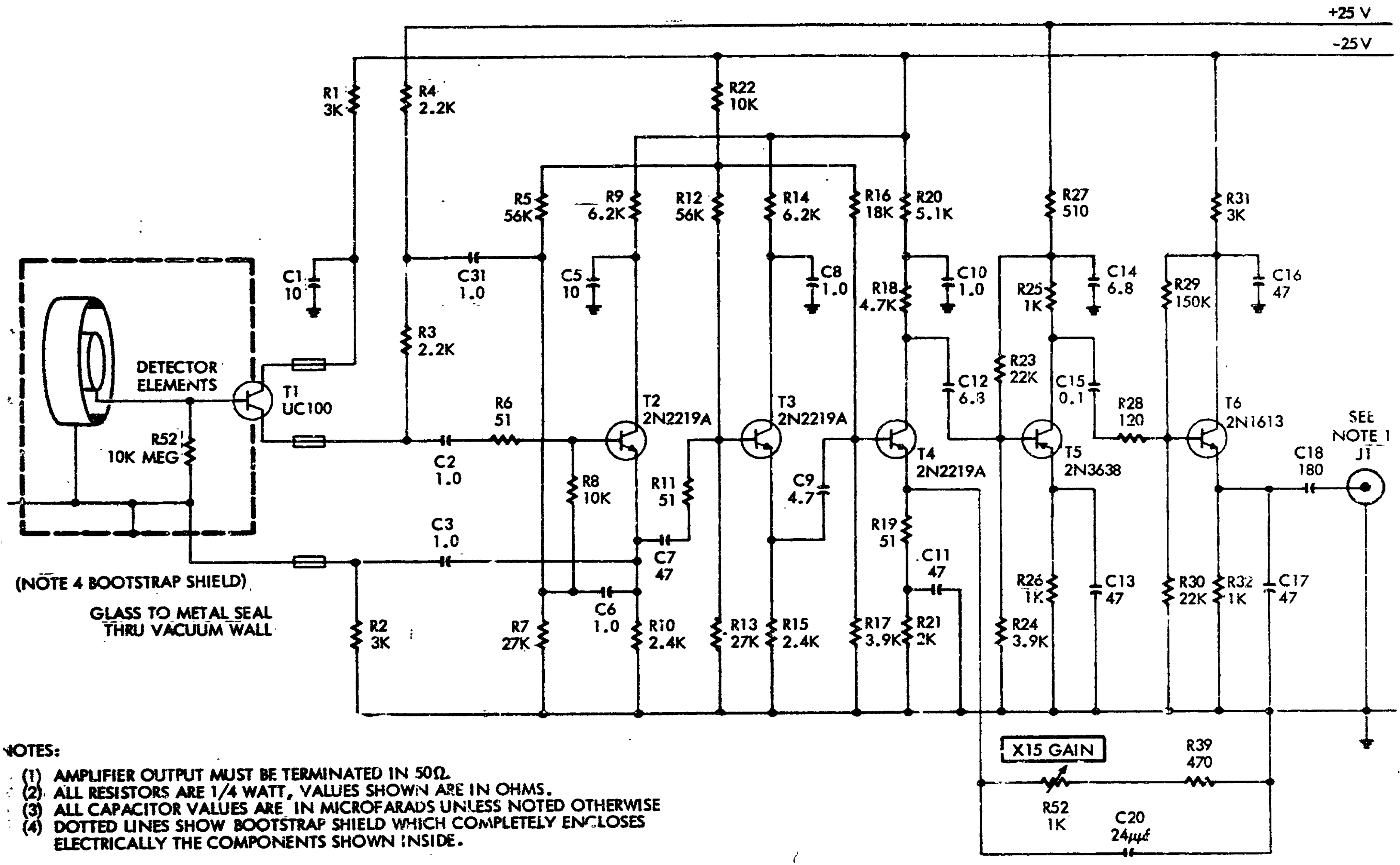
$$t_r = \frac{2.2 \ R_o C_b}{(1 - A)} \quad , \tag{4}$$

where R_o is the output impedance of the bootstrap and C_b is the load capacitance that must be driven by the bootstrap loop. This equation points out that the effective load capacitance driven by the bootstrap loop is increased by the same amount that the effective input capacitance is reduced. C_b has a value of 20 pf, and a value in excess of 0.95 was obtained for A; therefore, for a high-frequency cutoff of 10 mc, R_o must be no greater than 50 ohms. An output impedance R_o of approximately 20 ohms is easily obtained with an emitter follower, resulting in the desired high-frequency cutoff.

Vacuum tubes were originally used in the input stage of the preamplifier because tubes are quite superior to conventional bipolar transistors when high input resistance, low input capacitance, and low noise are required. However, with the development of the field effect transistor (FET) the semiconductor family became very competitive in this area, and in some instances, superior to vacuum tubes for such applications. In the source-follower configuration the FET has an equivalent input resistance of 10^{11} ohms or greater, allowing the use of a gate resistance of 10^{10} ohms, whereas the maximum value of grid resistance for tubes is much lower. It will be shown that a high value for this characteristic is very important with respect to the noise properties of the preamplifier. The interelectrode capacitances of FET's are lower, and the noise properties of this type of transistor are better than those of tubes. In addition, when vacuum tubes are used precautions must be taken to avoid the introduction of noise into the preamplifier through the filament supply; with FET's, of course, no filament supply is needed.

Although the transconductance of FET's with the above characteristics is considerably lower than that of tubes, in many cases this is not a serious disadvantage, since a conventional transistor may be used to effectively increase the g_m of the FET. (This technique is used in the preamplifier under discussion.) In using FET's care must be taken, as with all semiconductors, to insure that voltage and current ratings are never exceeded.

Figure 1 is a complete circuit diagram of the preamplifier. The bootstrap shield, which is a cylindrical tube, completely surrounds the detector and T1, the FET. The bootstrap loop is taken from the emitter of T2 back to the detector. This arrangement provides a low bootstrap-loop output impedance R_o allowing the desired high-frequency response. T3 provides a high load impedance for T2, so that the bootstrap loop gain A may be kept as close to unity as possible. (A value slightly in excess of



NOTES:
(1) AMPLIFIER OUTPUT MUST BE TERMINATED IN 50Ω.
(2) ALL RESISTORS ARE 1/4 WATT, VALUES SHOWN ARE IN OHMS.
(3) ALL CAPACITOR VALUES ARE IN MICROFARADS UNLESS NOTED OTHERWISE
(4) DOTTED LINES SHOW BOOTSTRAP SHIELD WHICH COMPLETELY ENCLOSES ELECTRICALLY THE COMPONENTS SHOWN INSIDE.

Figure 1. Complete Circuit Diagram of Preamplifier

0.95 has been obtained for A.) The detector capacitance, nominally 10 pf, is reduced to less than 0.5 pf by bootstrapping, and since the gate-to-drain capacitance of the FET averages about 0.5 pf, the total effective input capacitance is approximately 1 pf.

The two emitter-followers T2 and T3 contribute only a small amount of noise because their source resistances are quite low (500 and 20 ohms, respectively). T4 through T6 comprise a feedback amplifier with an open-loop gain of 300, which has been cut back to 15 with a resulting linearity in gain of ± 1 percent for output pulse heights to 10 volts. The emitter-follower output stage T6 allows a terminated 50 ohms transmission line to be used as the output load.

With C_{eff} equal to 1 pf, the FET input resistance of at least 10^{11} ohms allows the low-frequency cutoff to be extended to 10 kc as desired. However, this large input resistance is even more important in its effect on the noise properties of the preamplifier. The FET, the detector, and the bias network are the major sources of noise in the preamplifier. The noise equivalent of the input circuit is shown in Fig. 2, where R is the FET bias resistor, E_{tn} is the equivalent rms noise generator of the resistor, C_{eff} is the equivalent input capacitance, and V_{in} is the equivalent rms noise voltage appearing at the input of the preamplifier. The rms noise voltage amplitude of the generator is determined by

$$E_{tn}^2 = 4kTR\Delta f \quad . \qquad (5)$$

It can be shown[5] that

$$\Delta f = 1.57 \, f_{-3db} \quad , \qquad (6)$$

where f_{-3db} is the upper half-power frequency. By use of this relationship a value for E_{tn} that is independent of R is obtained:

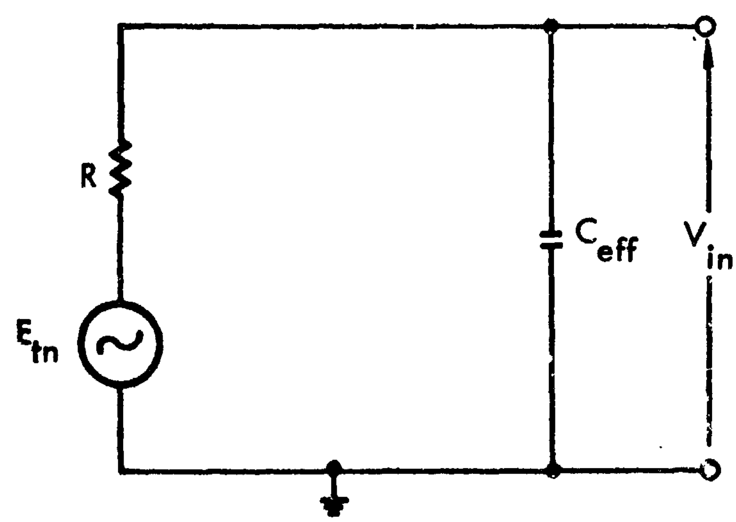


Figure 2. Noise Equivalent Input Circuit.

$$E_{tn}^{2} = \frac{kT}{C_{eff}}, \text{ or}$$
$$E_{tn} = \sqrt{\frac{kT}{C_{eff}}} . \qquad (7)$$

The voltage transfer function for Fig. 2 is

$$\frac{V_{in}}{E_{tn}} = \frac{1}{1 + i\omega RC_{eff}} \qquad (8)$$

and the power transfer function $G(\omega)$ is the square of the voltage transfer function. For purposes of calculation it will be assumed that the preamplifier has a sharp lower cutoff at 10 kc. (If such an approximation introduces appreciable error another transfer function may be introduced to describe the details of the low-frequency cutoff.) E_n, which represents the total rms noise voltage appearing across V_{in}, the preamplifier input, can be obtained from

$$E_n^{2} = 4kTR \int_{2\pi x10^{4}}^{2\pi x10^{7}} \left| G(\omega) \right| \, d\omega \quad . \qquad (1)$$

When this integral is evaluated the value of E_n is found to be 5 µv, which is a factor of 25 less than the noise from E_{tn}. Since the maximum value of R that can be obtained with vacuum tubes is much lower (well below 10^{10} ohms), using vacuum tubes in this circuit would result in a considerably larger value of E_n.

Equivalent noise circuits have been derived for FET's;[6] however, as a first-order approximation an FET may be regarded as a noisy resistor of value $1/g_m$. The g_m of the FET used in the preamplifier is 2000 mhos, resulting in an equivalent rms noise voltage at the input of 10 µv. The measured noise voltage referred to the input was 40 µv.

In summary, on the basis of the design considerations discussed in this paper, a preamplifier with the following characteristics has been constructed:

1. An equivalent input resistance of 10^{10} ohms
2. An equivalent input capacitance of 1 pı
3. An equivalent rms input noise charge of 250 electrons
4. Broadband response, from 10 kc to 10 mc
5. Use of all solid-state components.

APPENDIX

The expression for the effective input capacitance with bootstrapping may be found with the aid of Fig. A-1. In this analysis the input impedance is assumed to be infinite and the reactance of C is ignored. If it assumed that at some instant a charge q is placed at the input, the following equations may be written:

$$V_i = V_{C_d} + V_o = \frac{q}{C_d} + V_o \qquad \text{(A-1)}$$

$$V_o = A\ V_i = A\ \frac{q}{C_{eff}} \quad . \qquad \text{(A-2)}$$

From these, it is found that

$$C_{eff} = C_d(1 - A) \quad . \qquad \text{(A-3)}$$

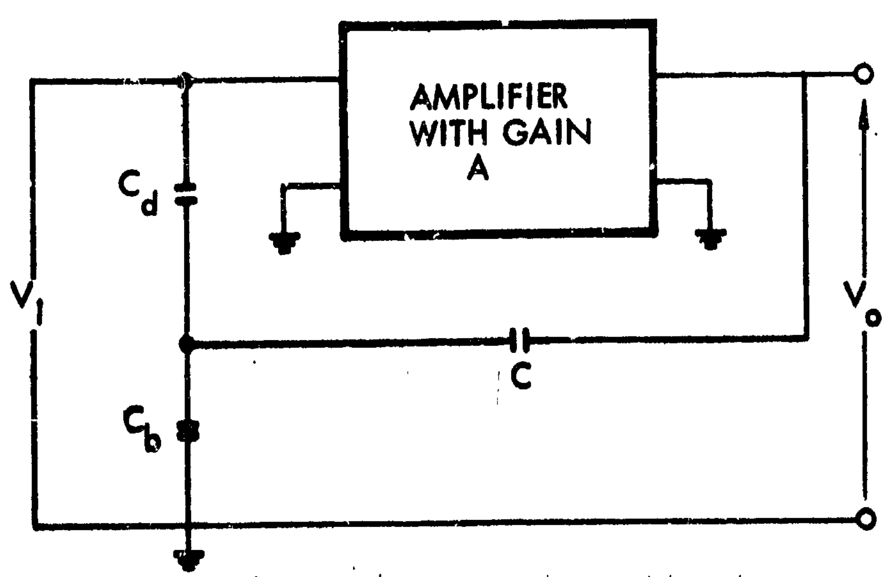


Figure A-1. Equivalent Circuit for Determination of Effective Input Capacitance with Bootstrapping.

In the preceding derivation it was assumed that the output impedance of the preamplifier was zero. The circuit in Fig. A-2 may be used to find the rise time. Again, it is assumed that at some instant a charge q is placed at the input. Then,

$$V_o = AV_i - iR_o \quad , \tag{A-4}$$

$$V_i = \frac{q}{C_d} + V_o \quad ; \text{ and} \tag{A-5}$$

$$V_o = \frac{1}{C_b} \int i \, dt \quad . \tag{A-6}$$

Differentiating Eq. (A-6) and eliminating i gives

$$\frac{dV_o}{dt} + \frac{V_o(1 - A)}{C_b R_o} = \frac{Aq}{C_d C_b R_o}$$

which upon solution yields

$$V_o = \frac{Aq}{(1-A)C_d} \left[1 - e^{-\frac{(1-A)t}{R_o C_b}} \right] .$$

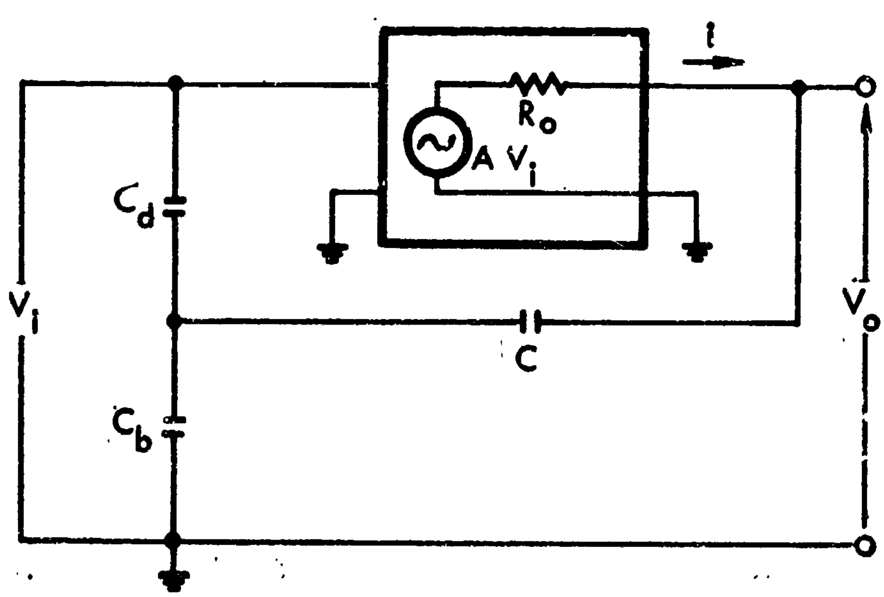


Figure A-2. Equivalent Circuit for Determination of Effective Output RC Time Constant.

The effective RC time constant is

$$RC_{eff} = \frac{R_o C_b}{(1 - A)} \quad . \tag{A-7}$$

It should be noted that if the input capacitance of the next stage is appreciable, its value should be added to C_b.

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