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COPY NO. - 19

DATE - August 23, 1962

A TRANSISTORIZED ALPHA COUNTER FOR AN ALPHA GAUGE

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ABSTRACT

A transistorized instrument prototype was designed and constructed to replace a vacuum-tube instrument in an alpha gauge, which measures the thickness density of gases. The instrument amplifies, shapes, discriminates, and counts alpha pulses from a Au-Si surface-barrier detector exposed to an alpha source in a gas-filled chamber.

The circuit consists of a charge-sensitive preamplifier, a main amplifier with pulse clipping, a Schmitt trigger, a diode pump, and a count rate meter.

Preliminary tests gave results comparable to the vacuum-tube instrument. Accuracy of counting was within plus and minus 10% for 0.5- to 10-Mev alpha particles emitted at a maximum rate of 10^6 per sec. The instrument was stable in a temperature range of 25 to 55°C, is small and portable, and costs less than $500. An infinitely thick, alpha source that will give a high count rate is being constructed for final tests.

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INTRODUCTION

After Gardner and Ross[1] had developed an alpha gauge, which measures the thickness density of gases by counting output pulses from a Au-Si surface barrier detector exposed to an alpha source in a gas-filled chamber (and can also be used to measure the thickness of thin films such as Mylar), they requested that a transistorized instrument be designed and developed to replace the vacuum-tube instrument for this gauge. The criteria for the new instrument were that it should amplify and discriminate the pulses from the detector and give an integrated count of the pulses, that it should be small and portable, and that it should be less expensive (less than $500) than the vacuum-tube instrument.

The over-all objective was to study and test several electronic circuits applicable to this type of instrument and to select the best circuit for counting alpha particles being emitted at a maximum rate of 5×10^5 per sec and having an energy range from 0.5 Mev (lowest level of the discriminator) to 10 Mev. The accuracy was to be within plus and minus 10%. The instrument was to be stable in a temperature range of 25 to 55°C. A prototype instrument was to be designed and tested.

GENERAL REQUIREMENTS AND DESCRIPTION OF THE CIRCUIT

The detector was an ORTEC Series A, N-type Au-Si surface barrier detector having a resistivity of 300 ohms and a sensitive area of 1 cm^2. For this detector to stop alpha particles having a maximum energy of 10 Mev, a barrier depth of 62 μ and a bias voltage of 50 v were required. The detector capacitance was 170 pf (see nomograph[2], Fig. 1).

Since an ionizing particle produces a hole-electron pair for each 3.5 ev of energy loss in the sensitive region of the detector, the maximum input charge per pulse for a 10-Mev alpha particle is

$$Q_{max} = \frac{(10^7 \text{ ev})(1.6 \times 10^{-19} \text{ coulomb})}{3.5 \text{ ev}} = 4.6 \times 10^{-13} \text{ coulomb.}$$

The minimum charge per pulse to be discriminated for a 0.5-Mev alpha particle is

$$Q_{min} = \frac{(0.5 \times 10^6)(1.6 \times 10^{-19} \text{ coulomb})}{3.5 \text{ ev}} = 2.3 \times 10^{-14} \text{ coulomb.}$$

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[1]R.P. Gardner and H.H. Ross (Oak Ridge Institute of Nuclear Studies), An Alpha Gauge for the Measurement of Small Density Thicknesses (unpublished).

[2]J.L. Blankenship and C.J. Borkowski (U.S. Atomic Energy Commission, ORNL), IRE Trans. on Nuclear Sci. NS-7 (2-3), 190-95 (1960).

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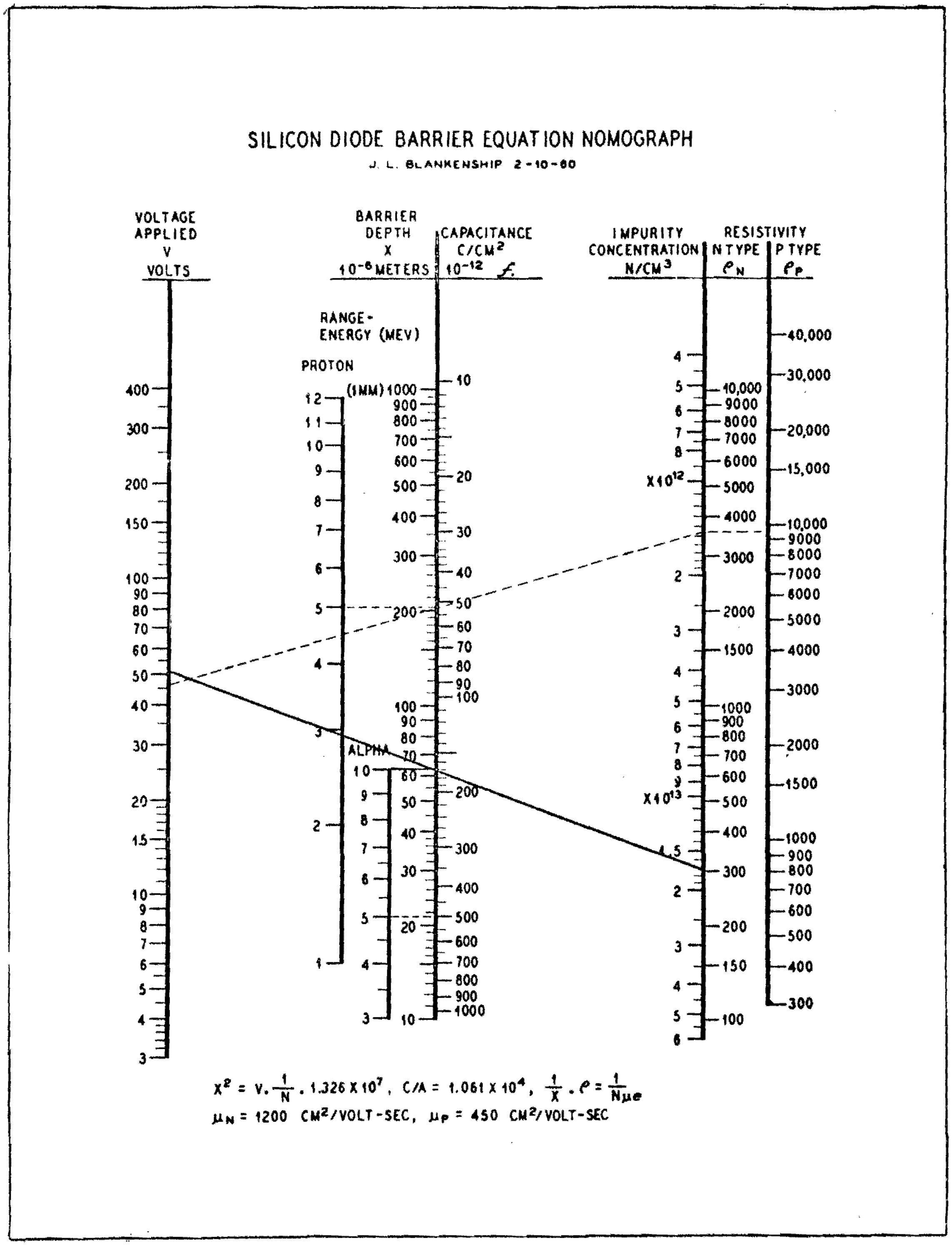

Fig. 1. Nomograph Giving Detector Capacitance, Barrier Depth, and Bias Voltage for a Au-Si Surface-Barrier Detector.

The circuit selected for pulse amplification, shaping, discrimination, and counting consisted of a charge-sensitive preamplifier, a main amplifier with pulse clipping, a Schmitt trigger, a diode pump, and a count-rate meter (Fig. 2).

A charge-sensitive, cascode amplifier, which degenerates the detector capacitance and causes nearly all of the input charge to appear across the feedback capacitor[3], was used as the preamplifier. This type of amplifier has a high input impedance; gives good, high frequency response, and low noise; and gives an output pulse whose amplitude is relatively independent of the detector capacitance.

To degenerate the detector capacitance ($C_D = 170$ pf), the feedback capacitance must be at least $C_F = 10(C_D/A)$, where A is the open-loop gain of the preamplifier. The output pulse amplitude is then very close to $\Delta V = \Delta Q_{in}/C_F$.

Two circuits[4,5] developed for use with semiconductor detectors were tested, and the preamplifier developed by T. Emmer was incorporated in the design because of its better signal-to-noise ratio. The second stage of this preamplifier (Fig. 3) provides an additional gain of $A_e \cong 25$ and a low output impedance.

To count random pulses at a rate of 5×10^5/sec, the maximum width of the output pulse has to be less than 200 nsec, which gives a count rate 10 times greater for equally spaced pulses. This shows the need for a clipping device to transform the step-like shape of the preamplifier output pulses into a short pulse of the required duration. Several coaxial cables were tested. The type RG-174/U with a characteristic impedance of $R_o = 100$ ohms and a delay time of 1.8 nsec/ft gave a pulse rise time of 10 nsec. The pulse amplitude reduction by a factor of 4 for double delay-line clipping and the bulkiness of two coaxial cables of 25 feet each made this clipping device not applicable. Therefore, an RCL shaping network[6] was tested. The underdamping factor was 0.6, and the amplitude reduction was only 0.6. Because it had an adequate pulse shape, low cost, and small size, this network was incorporated in the final design (Fig. 3).

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[3]E. Fairstein (Tennelec Instrument Co., Oak Ridge), IRE Trans. on Nuclear Sci. NS-8 (1), 129-40 (1961).

[4]R.L. Chase, W.A. Higinbotham, and G.L. Miller (Brookhaven National Laboratory, Upton, N.Y.) IRE Trans. on Nuclear Sci. NS-8 (1), p 147-50. (1961).

[5]T. Emmer (U.S. Atomic Energy Commission, ORNL), Instrumentation and Controls Div. Ann. Prog. Rep. for Period Ending July 1, 1961, ORNL-3191, p 6-11.

[6]J. Millman and H. Taub, Pulse and Digital Circuits, p 48-57, McGraw-Hill, New York, 1956.

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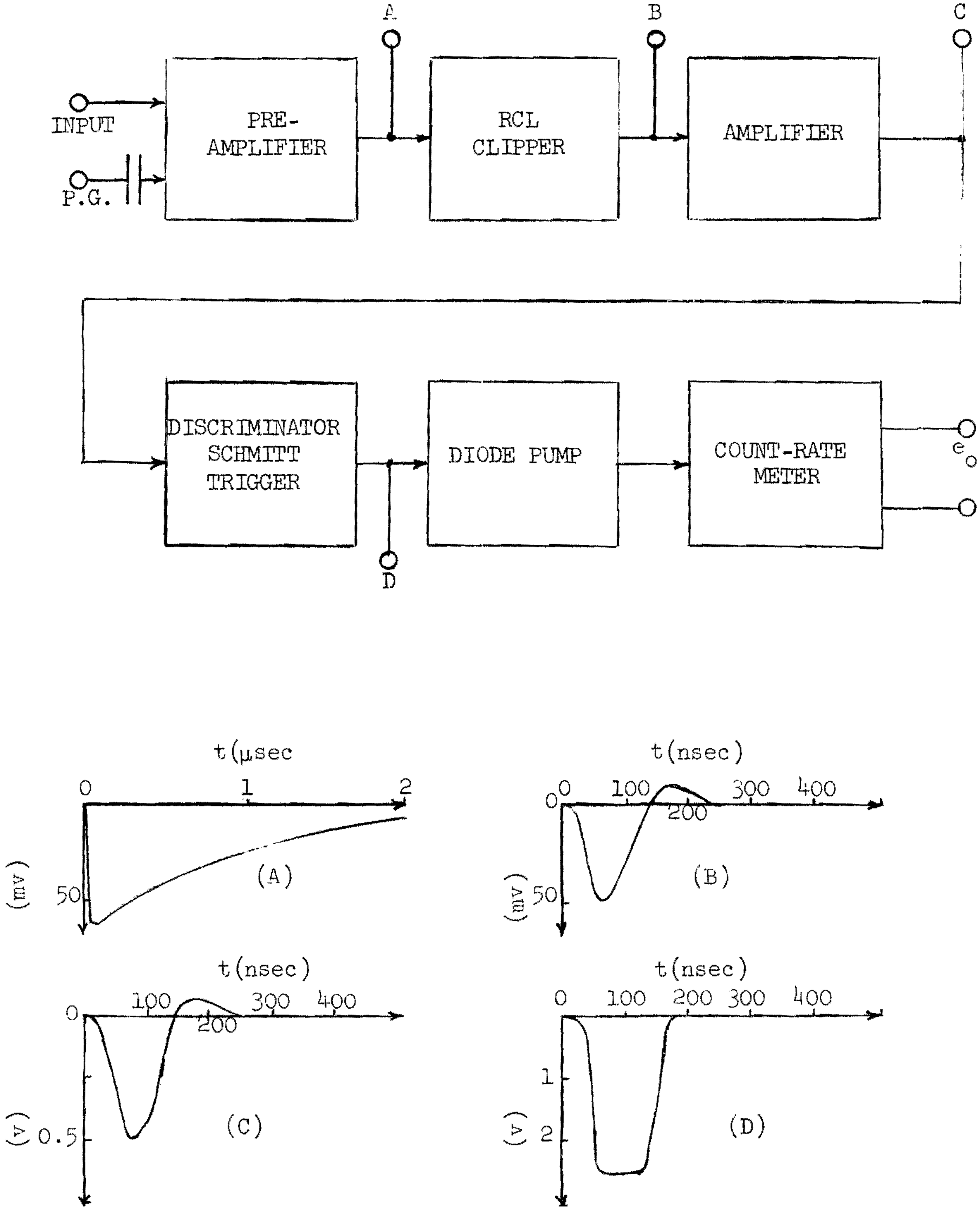

Fig. 2. Block Diagram of the Alpha-Gauge Pulse Amplification and Counting Circuit. The input and output pulse shapes are shown for each component.

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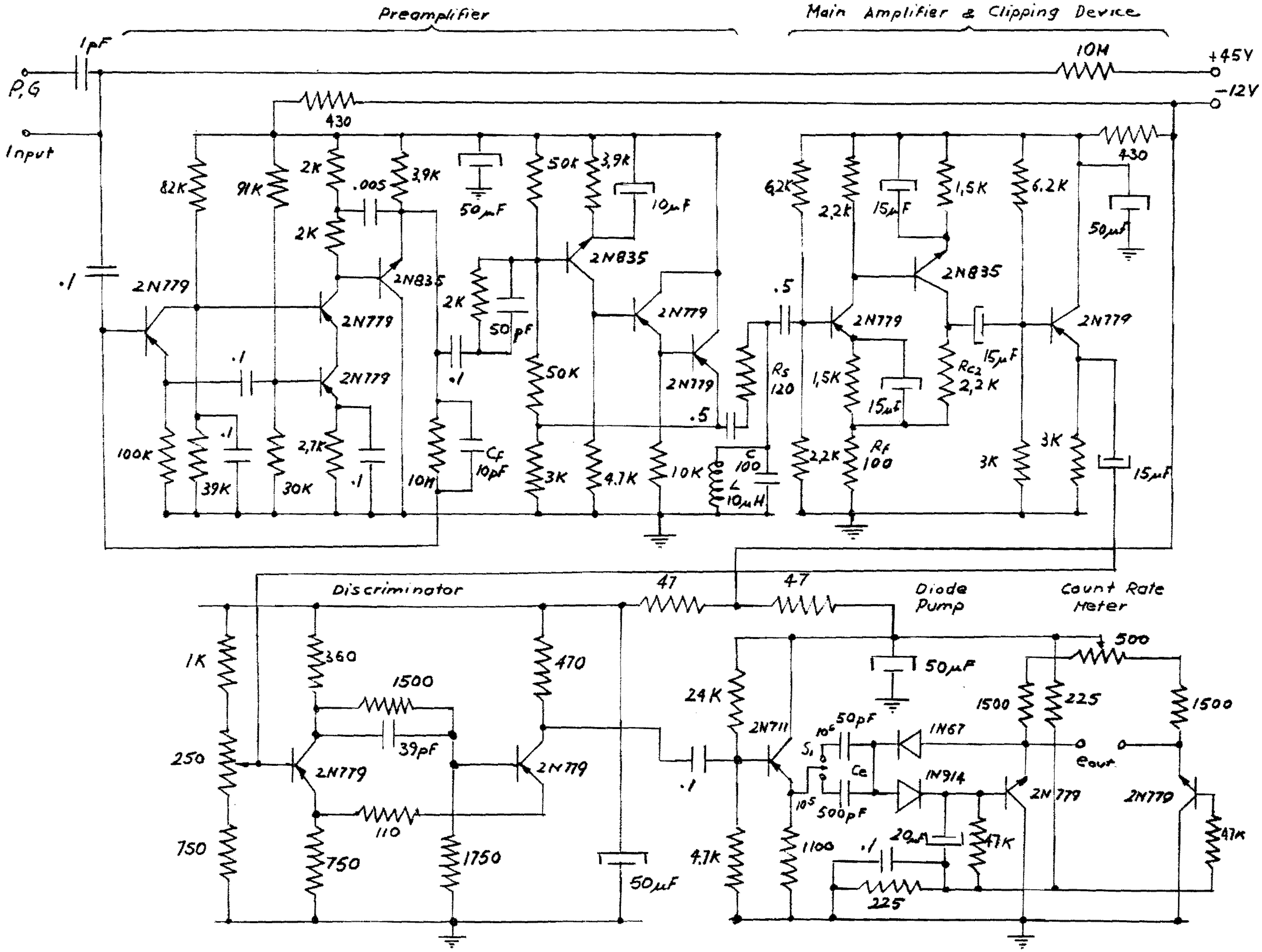

Fig. 3. Diagram of the Pulse Amplification and Counting Circuit.

Since the voltage gain of the amplifier was required to be only 20, a voltage feedback amplifier was used, which had good stability, high input impedance, low output impedance, and good, high frequency response.

Because the discriminator circuit was required to have a stable trigger level, fast rise time, and constant amplitude of the output pulse, a Schmitt trigger circuit was studied for incorporation into the total circuit. Two circuits were tested.

The output pulse of the discriminator was fed through an impedance-matching emitter-follower into a diode pump, giving an output current of

$$I_o = nq = n\Delta VC_e,$$

where

n = the average number of pulses per unit time,

ΔV = pulse amplitude,

C_e = coupling capacity between emitter-follower and diodes.

To make the transferred charge (q) independent of n, a bootstrapped emitter-follower was used as a count-rate meter. In order to stabilize the circuit for changes in ambient temperatures, a differential emitter-follower stage was tested for linearity and temperature effect.

RESULTS AND DISCUSSION

Preamplifier Circuit

The cascode circuit was connected to the input via an impedance-matching emitter-follower. The bias current was taken from the output through a 10-megohm resistor which, together with the feedback capacitor (C_F = 10 pf), determined the integrating time constant of the preamplifier. The open-loop gain of the cascode circuit was $A_e \cong 1500$.

The second stage added a voltage gain of 25 and gave a low output impedance to drive the clipping device.

The preamplifier equivalent input noise at full width, half-maximum was 20 kev at zero input capacitance and 100 kev at 200 pf input capacitance (see Fig. 4). The signal output was 67 mv/Mev input, the pulse rise time was 30 nsec, and the fall time was 2 μsec.

Clipping Device and Amplifier

The undamped period of the RCL circuit is

$$T_o = 2\pi\sqrt{LC},$$

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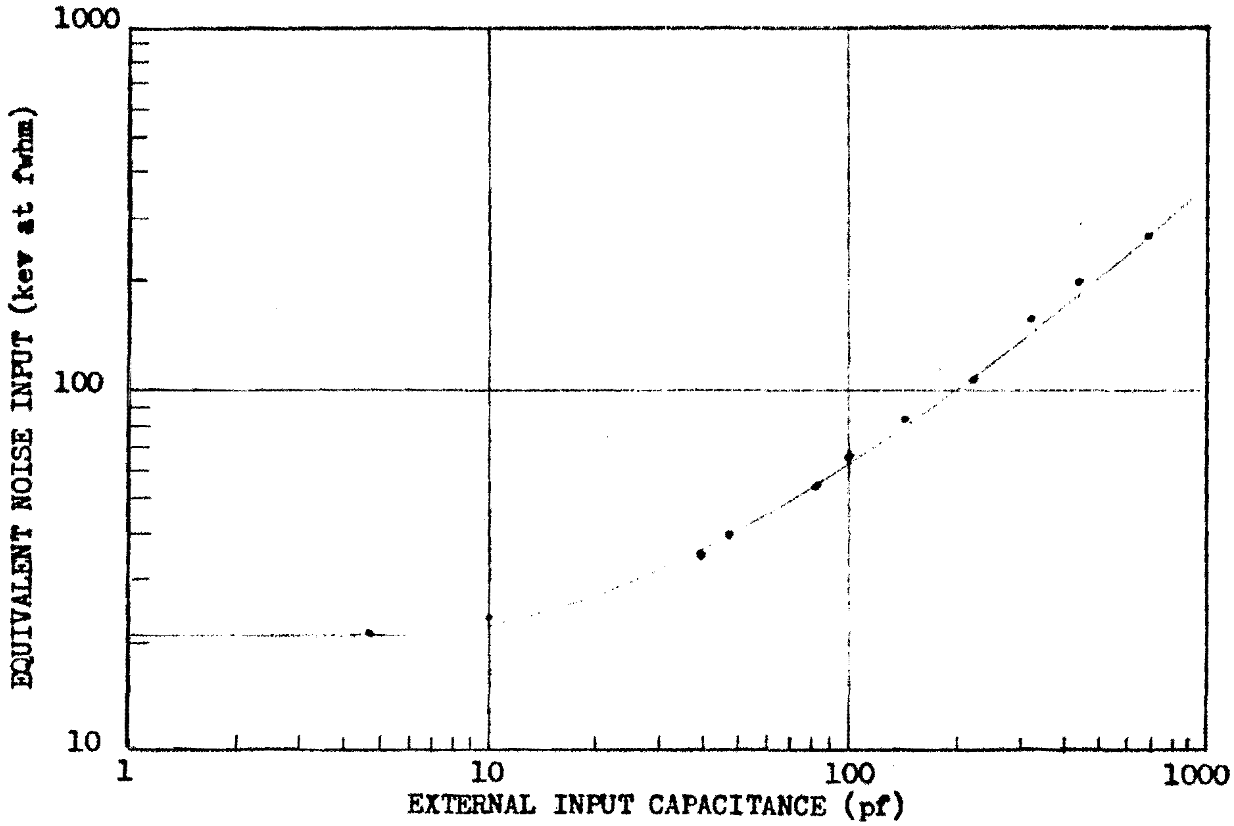

Fig. 4. Preamplifier Noise as a Function of Input Capacitance.

where

L = inductance in henrys,

C = total capacitance in parallel with the inductance in farads.

To obtain T_o = 200 nsec with an inductance of L = 10 μh,

$$C = (\frac{T_o}{2\pi})^2(\frac{1}{L}) = (\frac{2 \times 10^{-7}}{2 \times 3.14})^2 \times 10^5 = 10^{-10} = 100 \text{ pf.}$$

The selected damping factor was

$$K = \frac{1}{2R}\sqrt{\frac{L}{C}} = 0.6,$$

where

R = the total resistance in series with the LC circuit. It is the sum of the Thevenin generator resistance and the series resistor.

In this circuit

$$R = \frac{1}{2K}\sqrt{\frac{L}{C}} = \frac{1}{1.2}\sqrt{\frac{10^{-5}}{10^{-10}}} = 260 \text{ ohms.}$$

The output impedance of the preamplifier (R_{o_1}) was ≅ 150 ohms; the series resistor (R_s) was adjusted to 120 ohms.

The clipped pulse was fed into a two-transistor, series-feedback amplifier. The signal was amplified in the first stage and direct-coupled to the complementary second stage. The collector current from the second stage flows through an unbypassed part of the emitter-resistor (R_f) of the first stage and provides the negative-voltage feedback. The voltage gain of this amplifier was

$$A_f \cong \frac{R_{c_2} + R_f}{R_f} = \frac{2200 + 100}{100} = 23,$$

where R_{c_2} = collector resistor of the second stage, the input impedance (R_i) was ≅ 200 kilohms, and the output impedance (R_{o_2}) was ≅ 500 ohms. The pulse rise time (t_r) was ≅ 30 nsec, and the pulse width (t_d) was ≅ 120 nsec.

Discriminator

This circuit[7] consists of a conventional Schmitt trigger, with special

[7]G.T. Webb (Master's Thesis, University of Tennessee), A Transistor Crossover Pickoff Circuit for Use in Fast-Slow Coincidence Systems, University of Tennessee, Department of Electrical Engineering, Scientific Report No. 1, ORNL Subcontract No. 2075, Sept. 11, 1961.

care taken to minimize temperature effects on the trigger level and pulse height. Therefore, the bias circuits had very low impedances, the circuit was driven from a low-input emitter-follower, and the value of the common emitter-resistor was relatively high.

Temperature tests (between 25°C and 65°C) showed a change of 5 mv/10°C for the upper trigger point, 0.3 mv/10°C for the lower trigger point, and 5.3 mv/10°C for the hysteresis. The resistor between the two emitters affects the hysteresis, which was adjusted to be 50 mv. The output pulse amplitude was 2.5 v,with a rise time and fall time of 25 nsec each.

Diode Pump and Count-Rate Meter

The output impedance (R_{O_3}) of the emitter-follower connected between the discriminator and diode pump was approximately 20 ohms.

The coupling capacitor (C_e) between these two circuits has to be charged up in less time than the pulse duration. Therefore, the time constant ($R_{O_3}C_e$) has to be

$$R_{O_3}C_e < \frac{t_d}{5},$$

where td is the pulse duration.

In this circuit

$$C_e < \frac{t_d}{5R_o} = \frac{120 \times 10^{-9}}{5 \times 20} = 1.2 \times 10^{-9} = 1200 \text{ pf}.$$

Either of two ranges of 10^6 counts/sec and 10^5 counts/sec can be selected by connecting the proper capacitor to the circuit with the switch S_1 (Fig. 3). For both ranges the full-scale current output of the diode pump was

$$I_{fsd} = (n\Delta V)(C_e) = (10^6)(2.5)(50)(10^{-12}) = 125\ \mu a,$$

where

n = average number of counts per second,

ΔV = pulse amplitude, v,

C_e = value of the connected coupling capacitor in farads.

This current flows into a bootstrapped emitter-follower, which makes the voltage in the cathode of the 1N67 diode equal to the anode voltage of the 1N914 diode, and therefore, the transferred charge per pulse is almost independent of the count rate.

The emitter-follower was connected in a differential mode, and the output voltage (e_o), which appeared between the two emitters, was

$$e_o \cong 600 \text{ mv for a count rate of } 10^6 \text{ counts/sec with } C_e = 50 \text{ pf},$$

$$e_o = 600 \text{ mv for } 10^5 \text{ counts/sec with } C_e = 500 \text{ pf (see Fig. 5)}.$$

The voltage drift with temperature was $\Delta e_{oT} = 0.3$ mv/°C between 25°C and 70°C. The nonlinearity was less than 3% of full-scale deflection (Fig. 5). The voltage drift over a long time for a relatively constant ambient temperature was 2.5 mv during the first 20 min after the circuit was connected and then within 5 mv per 24 hr over a period of 8 days. These results were obtained with unmatched transistors ($h_{fe_1} = 26$ and $h_{fe_2} = 35$).

Since the count rate as a function of gas pressure, density, or film thickness is an experimental curve, a final calibration has to be made when an alpha source of high activity is available. Preliminary tests with a weak alpha-particle emitter gave results that were in accordance with test results obtained with a vacuum-tube amplifier and count-rate meter.

ACKNOWLEDGEMENT

The author wishes to thank J. B. Davidson and J. T. DeLorenzo for their many helpful suggestions.

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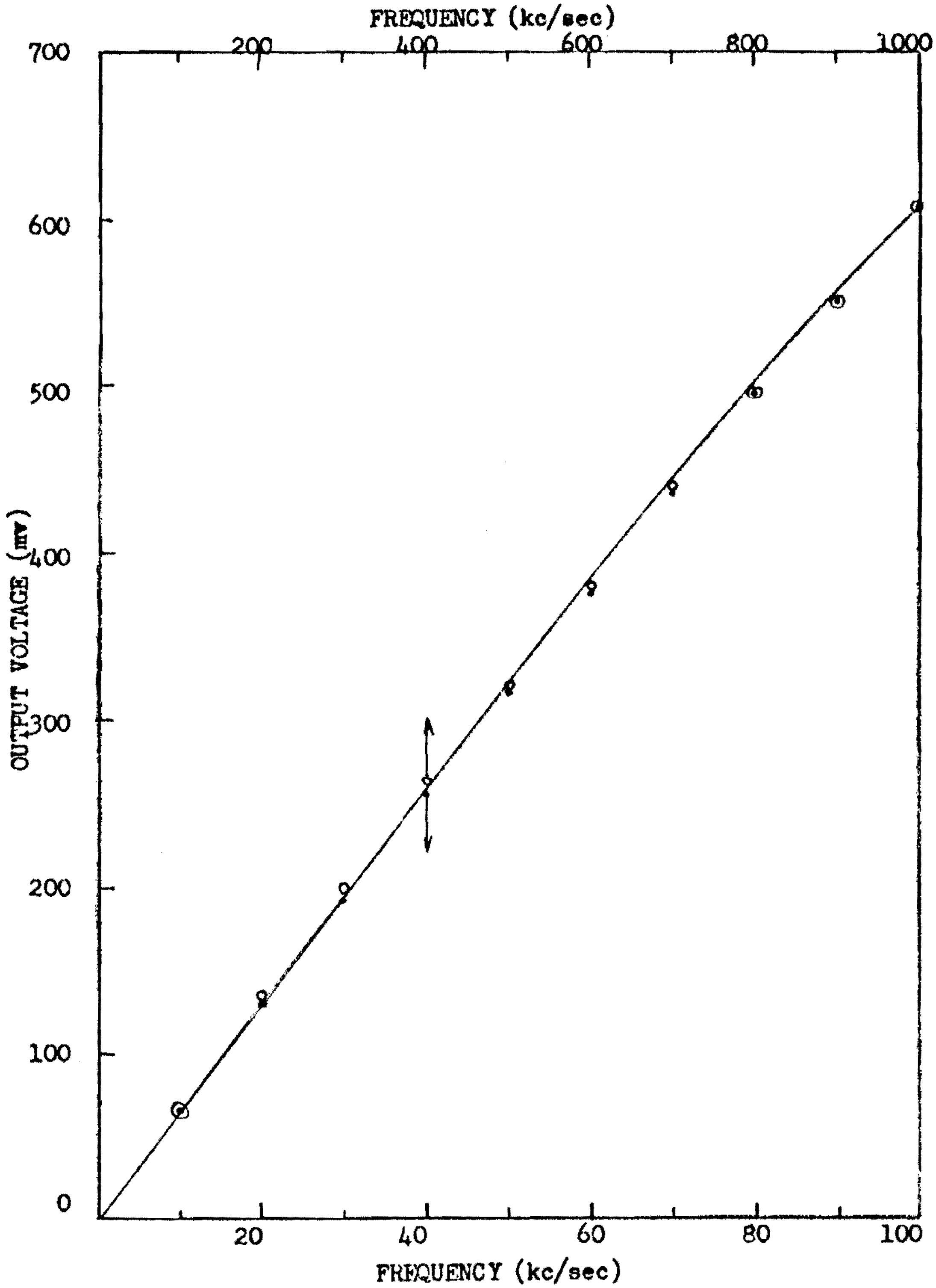

Fig. 5. Output Voltage as a Function of Frequency.

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