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Ernest O. Lawrence
Radiation Laboratory

GAMMA AMPLIFIER-DISCRIMINATOR

Livermore, California

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Curtis Sewell, Jr.

October 31, 1963

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ABSTRACT

For applications involving the routine laboratory measurement of gamma radiation, an instrument is described which is simple to operate, versatile, compact, and reliable. It consists of a pulse amplifier with charge-sensitive input and delay-line shaping, a differential discriminator capable of covering the range from 2 keV to 4 MeV, a calibration pulser, and a power supply. All controls are calibrated directly in terms of gamma energy. Construction is modular for ease of maintenance, and all circuits use solid-state components.

GENERAL

The entire instrument is comprised of four modules, a pulse amplifier, a differential discriminator, a calibrator, and a power supply. These are interconnected as shown in Fig. 1. Construction is modularized, with modules designed for easy removal. Total height is 3-1/2" and panel width is a standard 19", although only 11.6" is required for this instrument. All printed circuit boards are approximately 2-3/4" x 11". The amplifier, discriminator, and power supply each require two boards, while the calibrator requires one. Each module has its own panel, to which the boards are firmly mounted.

AMPLIFIER

The input of the amplifier is of the charge-sensitive type. This is, in effect, a high-capacitance input, offering the advantage of allowing cables of various lengths to be used to connect the probe to the amplifier without introducing appreciable errors due to cable capacitance. Because of this type of input circuit, the conventional voltage ratio method of specifying gain is meaningless. The sensitivity of the input section is approximately 0.05 volt/pico-coulomb. This is followed by two loops, each with a gain of 10, so that the three outputs from the amplifier represent sensitivities of 0.05, 0.5, and 5.0 volts/pico-coulomb. These three outputs are available simultaneously for use by the discriminators.

The three positions of the two discriminator range switches are labeled 40, 400, and 4000 keV. These are the maximum gamma energies to be measured on these ranges. They correspond to maximum voltages of approximately 3 volts. Since the discriminators cover a range of 20 to 1, the effective discriminator range, overall, is from 2 to 4000 keV. These numbers are for use with an average NaI(Tl) crystal and a photomultiplier gain of approximately 10^5 .

The output pulses from the amplifier are positive, delay-line-shaped, with about 1- μ sec base width. Rise time is determined by the 0.25 μ sec decay time constant of the NaI(Tl) crystal, and the delay-line clipping time is 0.5 μ sec. Thus the pulse is clipped before the full rise is achieved and therefore the gain is somewhat dependent on the rise time. This is not normally a healthy situation; however, it is used here in order to achieve minimum counting losses at high counting rates.

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Fig. 1. Block diagram.

A toggle switch selects either the photomultiplier signal or the calibrator pulse for use by the amplifier. Calibration energy is adjusted on the Calibrator Module.

The input sensitivity of the charge loop is given simply by:

$$V_2 = \frac{Q_{in}}{C_1} \quad (1)$$

If we include the crystal and PM tube, this equation becomes:

$$V_2 = \frac{ekA_{PM}E}{C_1} \quad (2)$$

where

e = electronic charge

k = photoelectrons/keV

A_{PM} = gain of PM tube

E = gamma energy in keV.

Thus with typical numbers, this loop produces 1.5 mV for a 2-keV gamma; a 4-MeV gamma causes an output of 3 volts. These are the extremes within the operating range.

The function of R_2 is to determine the decay time, or recovery, of the output voltage pulse. This time constant is $T_D = R_2 C_1$. The value of 10 μ sec chosen is a compromise between the long time desired for clean clipping and the short time desired for good overload recovery.

In the calibrate position of the input mode switch, a signal from the test pulser is applied through C_c to the amplifier input terminal. This signal is a flat-topped pulse, 2.5 msec long and having a rise time constant of 0.25 μ sec. This produces a current at the amplifier which is a better approximation to a nuclear event than is the more usual "tail pulse." The amplitude of V_2 is determined simply by the ratio C_c/C_1 and the Calibrator dial setting. Since C_c is adjustable, it can be set initially so that the Calibrator dial is direct-reading in keV of gamma energy.

Referring now to the Amplifier schematic, Fig. 2, the charge-sensitive loop is composed of Q_1 , Q_2 , and Q_3 . Its emitter followers give a high input impedance (except for the effect of feedback) and a low output impedance. Bootstrapping by means of C_5 increases the effective value of Q_2 collector load impedance and hence the loop gain. Q_1 is chosen to be a high-beta type

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Fig. 2. Amplifier schematic (LE 10303-11).

and operates at a low collector current so as to minimize noise. Bias voltages are chosen so that pulse pileup due to high counting rates causes negligible changes in gain. Although the largest single pulse produced at low counting rates is 3 volts, the amplifier will deliver more than 15 volts before gain reduction is noticeable. Since this loop is directly followed by the delay-line clipper, the later loops do not need this overload capability. Diodes CR_1 and CR_2 are included to protect the base-emitter junctions from voltage surges which may result from turnon or turnoff.

The delay-line clipper is shown in elementary form in Fig. 3. Waveshape A is the input pulse, B (solid line) is the i_{01} output current. Note the undershoot due to the rapid decay time. Waveshape B (dotted line) is the bridge-balance current i_{02} which compensates for this undershoot. The degree of compensation can be almost exactly adjusted by means of potentiometer R_9 . Because of the rather slow rise time of a pulse from NaI(Tl), the actual pulse resembles the waveshape at D.

The current-fed mode of clipping was chosen instead of using a high input impedance loop because of stray capacitance considerations at the end of the delay line. Even 1 or 2 pf at this point causes a noticeable mismatch. With care in construction, the present system allows this capacitance to be minimized.

The following loop uses a common emitter-common collector pair, with feedback by means of R_{15} . Two separate emitter followers are used at the output of this loop; one to drive remaining stages of the amplifier, and one to furnish the lowest-gain output to the discriminators. This minimizes interaction of switches. This loop contributes just enough gain to compensate for losses incurred in the delay-line clipping stage. The voltage gain adjustment is set for unity gain from the output of the first loop to the output of the second loop.

The next two loops are similar, each incorporating low input impedance current feedback. In each loop, a control allows the gain to be set to exactly 10. All loops are dc coupled internally to minimize bias shifts and prevent overload blocking. Each of the three gain loops overloads sharply at 5 to 6 volts and recovers as soon as the overload is completed. Thus if the delay-line controls are properly set, an overload of a factor of 100 or more causes saturation for no more than twice the width of a normal pulse.

Fig. 3. Delay line clipper.

An Auxiliary Output is provided in addition to the three main outputs in order that a pulse-height analyzer or other equipment can be used. Driver for this stage is a complementary White emitter follower to provide low output impedance for both polarities of signal.

Note that three outputs are simultaneously available, corresponding to three separate gains, or sensitivities. Thus the two discriminators could be switched so that the Lower Level discriminator triggers above 20 keV, while the Upper Level discriminator triggers above 2 MeV. This arrangement contributes to the versatility of the instrument.

DISCRIMINATOR

Two discriminators, one for upper level and one for lower level, are provided in the discriminator module. An anticoincidence gate and shaper is also in this module. Each discriminator receives its input signal through its own three-position Range switch from the three outputs of the amplifier. Each also has its bias level adjustable over a 3-volt range by means of ten-turn potentiometers. Bias of the two discriminators can be completely independent of each other, or the upper level can be referenced to the lower to enable a scan, or window, operation. These two modes of operation are selectable by a front-panel toggle switch labeled " E_U or ΔE ." By these two modes of operation it is possible, for example, to count all pulses from gammas in the range of a few keV up to a few MeV (but not cosmic rays) when the E_U mode is selected, or, in the ΔE mode, a narrow window of a few keV can be moved over a wide range of energy for a differential spectral analysis or selective counting of one photopeak.

Two output signals are provided from the discriminator through BNC connectors on the rear. One of these, Channel Output, is the signal resulting from a pulse lying between the upper and lower discriminator levels. The other, LL Output, is the signal caused by a gamma of energy above the lower level. This signal could be fed into a scaler to count in a purely integral mode; however, it is normally used by the scaler-timer to provide a live-time measurement. In this use, the timer is gated off during all times in which the lower level discriminator is energized. This provides correction for the primary cause of dead-time loss; however, it does not correct for base-line shift caused by intense amounts of low-energy radiation.

Referring to the schematic, Fig. 4, and considering only the Lower Level section first, Q_1 and Q_2 , together with CR_8 and CR_7 , form the discriminating element. The base of Q_1 is held at approximately +0.5 V by means of the Zero Adj. control. The base of Q_2 is positive, variable from approximately +0.6 V to +4.5 V by means of the E_L level helipot. Both transistors are normally conducting approximately the same amount of current, but the emitter of Q_2 is more positive than the emitter of Q_1 . This causes CR_8 to conduct and CR_7 to be cut off. If a positive pulse of sufficient amplitude is applied to Q_1 , then CR_7 will begin to conduct, reducing the current in CR_8 and in Q_2 , and causing a positive pulse to appear at Q_2 collector. This pulse is coupled through the emitter follower to a Schmitt trigger.

Temperature stability is achieved because of the balanced nature of the circuit. Every semiconductor is compensated by a similar semiconductor in a balanced configuration. Bias voltage is derived from a 1N823A temperature-compensated zener reference diode.¹

The Schmitt trigger circuit Q_4 , Q_5 , and Q_6 , is unique in the arrangement of diodes which control sensitivity and hysteresis. In the Schmitt circuit, good snap action is accomplished by a high ratio of the cross-coupling resistors, thus giving a high closed-loop gain. This is usually accompanied by a large degree of hysteresis. In other words, once triggered, the input base has to be driven considerably toward reverse bias or the circuit will latch. This sets a limit on usable sensitivity, which can never be less than the hysteresis. Thus the usual design involves a compromise between sensitivity and "snappiness."

In this unit, the cross-coupling resistors, R_{20} and R_{21} are chosen for good snap action, which would make the base of Q_5 either +2.2 V or -1.3 V, if it were not for the diodes. This would give a hysteresis of 3.0 volts, unusably high. However, the diodes limit the voltage swing at Q_5 base to +0.6 or +0.35 V, while Q_4 base is normally about +0.1 V. Thus all voltage levels are set by diode voltage drops, rather than resistor ratios. This allows high sensitivity, good snap action, and low hysteresis, while permitting the use of nonprecision components. Sensitivity is about 0.4 volt.

¹The above section is based on a design by Goulding of Lawrence Radiation Laboratory at Berkeley, and on suggestions by C. Van den Heuvel of Lawrence Radiation Laboratory at Livermore.

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If a narrow input pulse is applied, the output pulse has a minimum width of about 0.5 μ sec. This could be decreased somewhat with some loss in "snappiness" by decreasing the cross-coupling speedup capacitor C_7 .

The anticoincidence circuit, Fig. 5, is composed of diode gates, CR_1 , CR_2 , CR_3 , and CR_4 , and a saturating amplifier, Q_1 and Q_2 , followed by gating transistor Q_3 and an output stage. The amplifier is normally cut off because of the action of CR_2 and CR_4 . If the Lower Level discriminator is triggered, CR_4 opens but CR_2 still holds the amplifier off and no inhibiting pulse occurs. At the end of the LL disc period, the LL pulse is differentiated and the resulting negative pulse is passed through Q_3 to the Channel Output stage.

However, if a large signal causes both discriminators to trigger, both CR_4 and CR_2 are opened, the amplifier saturates because of current in R_1 , and a portion of the amplifier output is fed back through CR_3 to hold it in saturation after the UL disc has recovered. An instant later, when the LL disc recovers, it acts through CR_4 to reset the amplifier to a cutoff condition. During the time that the amplifier is saturated, a negative inhibit signal is produced at Q_3 emitter, preventing the differentiated LL pulse from passing through to the Channel Output. Because of storage time in the amplifier, this inhibit signal lasts long enough to be effective. There is almost no dead time following this cycle, however, thus this circuit offers advantages over previous circuits which used stretching or one-shot elements. Since it is dc coupled, there are no time restrictions on widths of discriminator pulses. Because of the unusual amplifier-discriminator interconnections, it is possible for the UL discriminator to fire before the LL disc. With this anti-coincidence circuit, this does not impair the operation.

The entire discriminator can be used at repetition rates as high as 1.5 Mc with less than 3% shift in bias level. In a double-pulse test, with a 10-volt pulse followed 1 μ sec later by a 1-volt pulse, and the bias set for 1 volt, the second pulse will trigger at 0.97 volt, a 3% shift in bias level caused by the 10X overload. Pulse widths were 0.1 μ sec.

Two modes of operation are provided, selected by a DPDT toggle switch on the panel. In the E_U position, bias of the Upper Level and Lower Level discriminators are both referred to ground. Thus the energies read from the helipot dials are both expressed as a fraction of the full-scale range selected by the appropriate Range switch.

Fig. 5. Anticoincidence circuit.

In the ΔE position, the bias of the Upper Level discriminator is the sum of the two helipot and both discriminators have their range selected by the E_L Range switch. Some examples as shown in the table (Table I) will clarify this action.

Table I.

E_L Range	E_U Range	E_L Disc	E_U Disc	Mode	Scan width
400	4000	200	250	E_U	80 to 1000 keV
"	"	"	"	ΔE	80 to 180 keV
"	400	"	100	E_U	>80 to <40 inoperable
"	"	"	"	ΔE	80 to 120 keV

CALIBRATOR

The calibrator module generates a negative step of voltage with a rise time equal to the NaI(Tl) signal rise time. This signal is coupled through a small, adjustable capacitor into the amplifier input. Amplitude is controlled by a ten-turn potentiometer and a three-position switch. Thus the three ranges correspond to 40, 400, and 4000 keV full-scale, the same as the three amplifier-discriminator ranges. Repetition rate is approximately 6000 cpm. A trigger output BNC is furnished on the rear to aid in servicing with an oscilloscope.

Referring to the schematic, Fig. 6, Q_1 and Q_2 form a conventional free-running multivibrator, with a frequency of approximately 100 cps. The negative portion, approximately 2.5 msec wide, is clipped and amplified by Q_3 and applied to the base of switch Q_4 . The load for Q_4 consists of the zener reference diode, CR_2 , in parallel with the three-position attenuator. Thus the pulse delivered to the cascade emitter follower Q_5 and Q_6 has its amplitude accurately determined by CR_2 and its rise time accurately determined by R_{18} and C_6 . The emitter follower employs bootstrapping by means of C_7 to further increase its input impedance. It applies the previously standardized pulse to the helipot for precise amplitude control.

Although the circuit is simple and straightforward, its stability is almost as good as the best mercury-relay pulsers available. Temperature coefficient is less than 0.01%/°C. Output impedance is approximately 120

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Fig. 6. Calibrator schematic (LE 10303-31).

ohms. For proper operation, this pulser should operate into a high impedance, such as the Cal Input of an amplifier. It is not designed to drive a long cable. Linearity is determined primarily by the helipot to be $\pm 0.1\%$ of full scale. Correlation between decade ranges is determined primarily by the metal film resistors and is within $\pm 0.25\%$ of the dial reading. When used in this instrument, line voltage fluctuations between 100 and 130 V ac have no measurable effect.

POWER SUPPLY

All power supply components except the magnetically stabilized power transformer and its capacitor are mounted on a plug-in module. Electronic regulators furnish two voltages: +24 volts at 125 mA and -12 volts at 80 mA for operation of all circuits, and in addition a floating 15-volt magnetically regulated voltage for the upper level discriminator bias. Stability is sufficient so that supply voltages have no effect on circuit operation. Input voltage should be within the range 100 to 130 volts, 60 cps.

APPENDIX

The current pulse from a nuclear detector such as a scintillation crystal and photomultiplier tube has the following form:

$$i = \frac{KE}{T_0} e^{-t/T_0} \quad (1)$$

where

E = gamma energy

T_0 = decay time constant of crystal

K = proportionality constant.

If this expression is integrated, we find:

$$\int_0^\infty i \, dt = Q_{in} = KE. \quad (2)$$

Thus the total charge delivered to the amplifier is proportional to the energy of the incident gamma ray, as expected. If this pulse is applied to the input of a loop such as the charge amplifier used here, the output from the charge amplifier loop has the form:

$$e_0 = \frac{KE}{C_F} \times \frac{T_1}{T_1 - T_0} \times \left(e^{-t/T_1} - e^{-t/T_0} \right) \quad (3)$$

where

$$\begin{aligned} C_F &= \text{feedback capacity} \\ T_1 &= \text{feedback time constant} \\ &= R_F C_F. \end{aligned}$$

This is the representation for a pulse having an exponential rise time constant of T_0 and a decay time constant of T_1 .

The pulse most often used for calibration is a "tail" pulse having a fast rise and an exponential decay. When such a pulse,

$$e_1 = E e^{-t/T_D} \quad (4)$$

is applied to a charge loop, the output voltage has the form:

$$e_0 = E \frac{C_1}{C_F} \times \frac{T_D}{T_D - T_1} \times \left(e^{-t/T_1} - \frac{T_1}{T_D} e^{-t/T_D} \right) \quad (5)$$

where

$$\begin{aligned} C_1 &= \text{input coupling capacitor} \\ C_F &= \text{feedback capacitor} \\ T_D &= \text{Cal pulse decay constant} \\ T_1 &= \text{feedback time constant.} \end{aligned}$$

Examination of this equation shows an inherent undershoot following the pulse. Thus the tail pulse does not perfectly simulate a nuclear event. [Compare the exponential terms of Eq. (3) with Eq. (5).]

If, however, a flat-topped calibrating pulse having an exponential rise time equal to the crystal decay time is used, the output from the charge loop is of the form:

$$e_0 = E \frac{C_1}{C_F} \times \frac{T_1}{T_1 - T_R} \times \left(e^{-t/T_1} - e^{-t/T_R} \right) \quad (6)$$

where

$$T_R = \text{Cal pulse rise time constant.}$$

Comparison with Eq. (3) shows that this is exactly the proper shape to simulate the desired nuclear pulse. It does have one disadvantage, in that since it is a "square wave," there are alternate positive and negative portions. The above equation considers only one polarity. However, the second pulse can usually be ignored, since it has a slower rise time and therefore suffers in differentiation; also, the design of the amplifier is such that opposite polarity signals are not passed very well. Therefore, the alternate pulse will

normally be much smaller than the correct one, and can be ignored. In any event, there are several milliseconds between these pulses so that, when servicing, if the Trigger Signal is used to trigger the scope, there is no possibility of confusion.

In order to properly perform the "Tail Adj" and "DL Match" adjustments in setting up the amplifier, it is necessary to examine in detail the portion immediately following the main pulse, both under normal and overloaded conditions. If the calibrating pulse does not perfectly simulate a nuclear pulse, this is the portion that will give sizable errors. This is the reason that a flat-topped calibrating pulse is superior to a tail pulse and is chosen for this application.

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