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PULSE AMPLIFIER MANUAL

E. Fairstein

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INSTRUMENTATION AND CONTROLS DIVISION

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ABSTRACT

This manual was prepared to aid those who use and repair pulse amplifiers. The manual is divided into three sections: section 1 contains a number of generalized testing routines applicable to all pulse amplifiers; section 2 is devoted to test information relating to specific amplifiers; and section 3 is a bibliography of journal articles, reports, and books about pulse amplifiers.

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1. AMPLIFIER TESTING ROUTINES

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1.1 INTRODUCTION

In repairing an amplifier, three precepts should be kept in mind:

1. Most faults can be found with the aid of nothing more complex than a multimeter as the test instrument; however, before being released for service, all amplifiers should be tested for performance by a precision pulser and an oscilloscope.

2. Almost anything done to an amplifier that exhibits an intermittent fault will temporarily clear the fault. Replacement of all tubes will usually clear the fault for a somewhat longer time, but complete tube replacement is usually an expensive and temporary solution.

3. Replacement of all tubes in an amplifier is rarely required and is never justified as a substitute for other test procedures. Experience with systems that use large numbers of tubes indicates that the failure rate of tubes is high for the first 100 hr, drops off to a very low value for the next 3000 to 5000 hr, and then gradually rises as the tubes begin to exhibit faults typical of long use. Routine replacement of old tubes can result in a low and constant failure rate.

A tube tester is helpful, but it should not be considered infallible. Depending on the circuit application, tubes testing "bad" in a tester may be all right in the circuit, and vice versa.

1.1.1 Test Instruments

Three test instruments are required (listed in the order of their importance):

1. Multimeter, 20,000 ohms/v.
2. Oscilloscope, with a rise time as good as or better than the amplifier to be tested.
3. Precision pulser, ORNL Q-1212C or Q-1212D, or equivalent.

Four instruments are a convenience, but not a necessity:

1. Count-rate meter, Q-1237, Q-1511, or equivalent.
2. Vacuum-tube voltmeter, Hewlett-Packard model 400D, 400H, or 400L.
3. Tube tester.
4. Direct-current vacuum-tube voltmeter (VoltOhmyst, Hewlett-Packard 410B, H-P 412A, etc.).

1.1.2 Pulse Generator and Oscilloscope Connections

When possible, the amplifier output signal should be monitored with an oscilloscope while voltage measurements are being made. This may uncover oscillations caused by unwanted feedback or capacitive loading introduced by the voltmeter leads.

From the standpoint of observation, the most desirable setup is that shown in Fig. 1.1 (see Fig. 1.16 for a description of the capacitor fitting).

The important considerations are:

1. All signal-bearing leads are shielded to prevent unwanted feedback between output and input.

2. The "attenuated output terminal" from the generator is terminated in a 100-ohm resistor, preferably at the amplifier end of the connecting lead.

3. The oscilloscope sweep is triggered externally (from the generator) rather than internally from the amplifier signal.

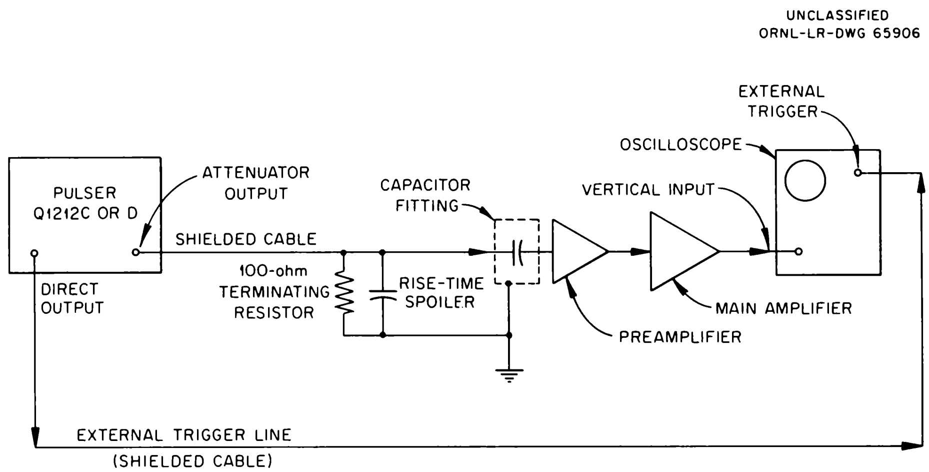

Fig. 1.1. Test Setup for Monitoring an Amplifier.

This method of triggering accomplishes the following: (1) makes it unnecessary to readjust the oscilloscope trigger circuit when changing the amplifier controls or the pulse-generator attenuator-switch settings, (2) keeps the oscilloscope from being triggered by amplifier noise, and (3) holds the starting point of the output pulse constant on the oscilloscope screen for easy timing measurements.

Unfortunately, because of inadequate shielding between the trigger and amplifier circuits in some oscilloscopes, the setup may oscillate with a high-gain amplifier. The possibility of producing such an oscillation is minimized (as is power-line noise pickup) by permitting only one of the line-operated instruments to be grounded. In extreme cases, the external triggering connection should not be used. In all cases, the system should be checked for unwanted feedback by looking for a change in waveform which may result from a shift between the external and internal triggering modes. (The external trigger connection to the pulse generator should be disconnected when using the internal triggering modes.) In the absence of feedback, no change will be observed.

1.2 EXPLANATION AND USE OF THE MULTIMETER METHOD IN DIRECT-CURRENT MEASUREMENTS

The 20,000-ohm/v multimeter is probably the most used instrument in dc measurements. While a high-impedance vacuum-tube voltmeter (vtvm) is required in development work for troubleshooting linear amplifiers, a vtvm rarely will reveal circuit faults hidden to a multimeter.

The test procedures in this report are based on the EXCLUSIVE use of a 20,000-ohm/v multimeter. This section is devoted to its use in ways which minimize its shortcomings.

1.2.1 Corrections for Loading Effects

The internal resistance of the voltmeter loads the circuit being tested and results in a low reading. If the loading is excessive, corrections must be made.

The internal resistances associated with commonly used meter ranges are given in Table 1.1.

Table 1.1. Meter Resistances[a]

Simpson Meters		Triplett Meters	
Scale Range (v)	Internal Resistance (ohms)	Scale Range (v)	Internal Resistance (ohms)
2.5	50 k	3	60 k
10	200 k	12	240 k
50	1 M	60	1.2 M
250	5 M	300	6 M
1000	20 M	1200	24 M
5000	100 M	3000	60 M

[a]Based on 20,000 ohms/v.

When the source resistance of the circuit under test is easily determined, the true voltage is obtained from the formula

$$V_{\text{true}} = V_{\text{meas}} \left(1 + \frac{R_S}{R_M} \right) , \tag{1}$$

where R_S is the source resistance and R_M is the meter resistance, both resistances being expressed in compatible units (ohms, kilohms, or megohms).

When the plate circuit of a tube is involved, part of the source resistance includes the internal plate resistance, which is not always easy to determine. The following test procedure eliminates the need to know the source resistance.

1. A meter reading is obtained on a convenient scale.

2. The meter is then switched to the next higher range and the new reading noted.

3. If the second reading is higher than the first, the meter is loading the circuit, and a correction should be applied. (Recall that for a 2% meter the discrepancy between two readings on different parts of the scale may amount to one minor scale division.)

The true voltage is determined from the meter readings by the expression

$$V_{\text{true}} = V_2 \, \frac{N - 1}{N - (V_2/V_1)} , \tag{2}$$

where

V_1 = reading on the lower scale,

V_2 = reading on the next higher scale,

N = the ratio of the full-scale readings.

From Table 1.1, it can be seen that the standard ratios between successive ranges are 2.5, 4, and 5. In Table 1.2, Eq. (2) is tabulated in terms of these range factors, and in Fig. 1.2, the correction factor is plotted as a function of V_2/V_1 for three scale factors.

Table 1.2. Correction Formulas for Meter Loading[a]

Scale Factor N	True Voltage
2.5	$\dfrac{3V_2}{5 - 2\,(V_2/V_1)}$
4	$\dfrac{3V_2}{4 - (V_2/V_1)}$
5	$\dfrac{4V_2}{5 - (V_2/V_1)}$

[a]See definitions under Eq. (2) for meanings of V_1, V_2, and N.

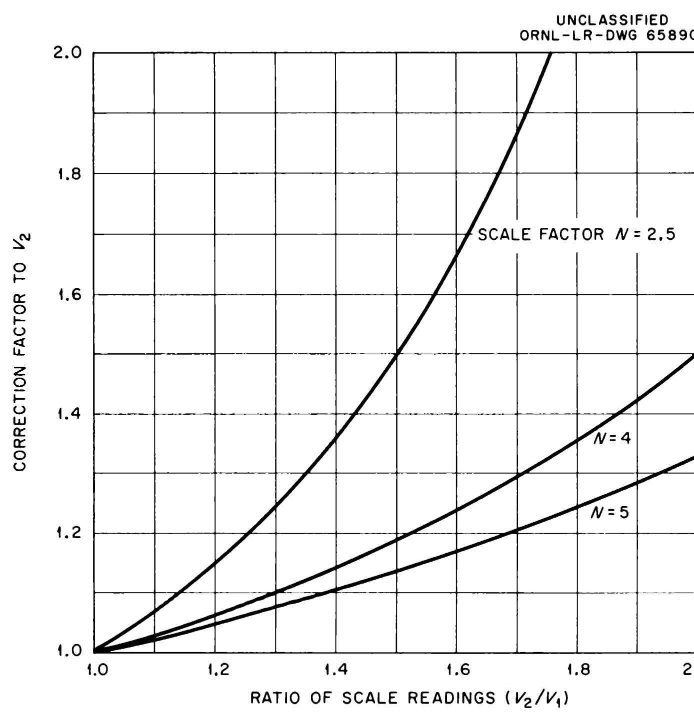

Fig. 1.2. Correction Factor for Meter Loading as a Function of the Ratio of Dial Readings on Different Meter Ranges.

The determination of the true voltage is illustrated by the following example: A meter reads 120 v on the 250-v scale and 130 v on the 1000-v scale. What is the true voltage? Solution: $V_1 = 120$, $V_2 = 130$, and $N = 4$.

$$V_{\text{true}} = \frac{3 \times 130}{4 - (130/120)} = 134 \text{ v} .$$

The method of reading the same voltage on two different ranges may be used for determining the source resistance of the circuit being measured. The general formula is

$$R_S = R_{M_2} \frac{V_2 - V_1}{NV_1 - V_2} , \tag{3}$$

where

R_S = the source resistance,
R_{M_2} = the meter resistance on the higher range,
V_1 = the meter reading on the lower range,
V_2 = the reading on the higher range,
N = the factor between ranges.

From the preceding example, $V_1 = 120$, $V_2 = 130$, $N = 4$, and $R_{M_2} = 20$ megohms.

$$R_S = 20 \text{ megohms } \frac{130 - 120}{4(120) - 130} = 0.57 \text{ megohms.}$$

1.2.2 Feedback Induced by the Multimeter – Inaccuracies and Corrections

When voltages on signal leads are measured, the multimeter may cause the circuit to oscillate and yield a false reading. WHEN MEASURING AMPLIFIER VOLTAGES WITH A MULTIMETER, MONITOR THE CIRCUIT WITH AN OSCILLOSCOPE TO BE SURE THAT THE CIRCUIT IS NOT OSCILLATING.

If an oscillation occurs, the "hot" meter lead may be decoupled from the circuit with a "probe" consisting of a resistor (with its leads clipped short) held in the jaws of the meter-lead test clip. Values between 100 kilohms and 1 megohm are usually necessary.

This added resistor may cause a reading error (see Table 1.2). Conversely, it can be a means of improving accuracy. For example, with a Simpson model 260 multimeter, a voltage of 270 must be read on the 1000-v scale, resulting in less than ⅓-scale deflection. By using a 5-megohm, 1% resistor as a probe, the 250-v scale is converted to a 500-v scale, and twice the former needle deflection is obtained.

1.2.3 Characteristics of the Multimeter Ohms Scale and Their Use

At times it is desirable to introduce a small voltage or current into a circuit under test. If the exact value is not critical, the ohms position of a multimeter may be useful. In Table 1.3, the open-circuit voltages, short-circuit currents, and source resistances of the ohmmeter positions of several multimeters are

Table 1.3. Ohmmeter Characteristics of the Simpson 260, Triplett 630, and Triplett 310 Multimeters

Meter	Ohms Scale	Open Circuit (v)	Short Circuit (ma)	Source Resistance (ohms)	Lead Polarity
Simpson 260	×1	1.55	131	18	Common
	×100	1.55	1.33	1.66 k	–
	×10 k	7.50	0.0615	122 k	
Triplett 630	×1	1.55	335	4.65	Common
	×10	1.55	34.5	45	+
	×1 k	1.55	0.345	4.5 k	
	×100 k	18.2	0.072	255 k	
Triplett 310	×1	1.55	8.0	194	Common
	×10	1.55	0.8	1.94 k	+
	×100	1.55	0.08	19.4 k	
	×1 k	15.5	0.08	194 k	

listed. Under load, the terminal voltage is proportional to 1 minus the fraction of the full-scale deflection, while the current is directly proportional to the fraction of the full-scale deflection.

The use of the information in Table 1.3 is shown in the following example. A Triplett 310 meter is used with readings taken on the ×100 scale. The meter reads 45 on the scale calibrated for 0 to 120. The meter deflection is $45/120 = 0.375 \times$ full scale. The terminal voltage is $(1 - 0.375)(1.55) = 0.97$ v. The current is 0.375×0.08 ma $= 0.03$ ma.

A word of caution: The full-scale current on the low-ohms scale of the Triplett 630 meter is sufficient to damage the end turns on high-resistance wire-wound potentiometers and the junctions of low-power diodes and transistors.

1.3 MEASUREMENTS OF NONLINEARITY OF AN AMPLIFIER

This section consists of a full discussion of integral and differential amplifier nonlinearity. Amplifier nonlinearity may enter into the accuracy with which spectral measurements can be made. If a pulse-height analyzer is used to determine the RELATIVE INTENSITIES of two or more parts of a spectrum, the amplifier nonlinearity is unimportant, but if the ENERGIES of several regions of a spectrum must be compared, then the integral nonlinearity must be known. If the SHAPE of the various parts of the spectrum must be determined, both the differential and integral nonlinearites must be known. For a description of the usual method for measuring integral nonlinearity, refer to Sec 1.3.6, "method I," or to the *Instruction Manual, Mercury Relay Pulse Generator, Model Q-1212C*, ORNL CF-60-9-67.

1.3.1 Definition of Integral Nonlinearity

Integral nonlinearity is the deviation of the amplitude response of an amplifier from that of a perfect amplifier.

Suppose that a calibrated, adjustable pulse generator is connected to the input of an amplifier and that the resulting output pulses are accurately measured for amplitude. If the measurements are graphed, a curve similar to that shown in Fig. 1.3 may result.

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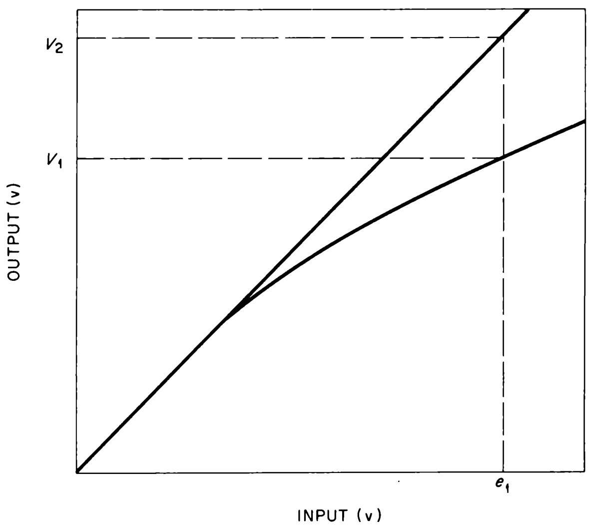


Fig. 1.3. Output Pulse Height vs Input Pulse Height of a Typical Amplifier.

If the amplifier is perfectly linear, the straight line of Fig. 1.3 will be obtained. The deviation from the straight line is the amplifier nonlinearity, and it can be expressed in terms of the percentage of the absolute amplifier output; that is,

$$IN_{abs} = 100 \, \frac{V_2 - V_1}{V_2} \, , \tag{4}$$

which converts to

$$IN_{abs} = 100 \left(1 - \frac{V_1}{V_2} \right) \, , \tag{5}$$

where

IN_{abs} = integral nonlinearity in percentage of the absolute amplifier output,
V_1 = ideal output signal, v,
V_2 = actual output signal, v.

In terms of the percentage of maximum rated amplifier output,

$$IN_{max} = 100 \left(\frac{V_2}{V_3}\right) \frac{V_2 - V_1}{V_2} , \tag{6}$$

which converts to

$$IN_{max} = 100 \left(\frac{V_2}{V_3}\right) \left(1 - \frac{V_1}{V_2}\right) , \tag{7}$$

where

IN_{max} = integral nonlinearity in percentage of the maximum rated amplifier output,

V_3 = rated maximum amplifier output, v.

In all equations V_1 and V_2 are, respectively, the ideal and actual output signals (in volts) resulting from an applied input signal, and V_3 is the rated maximum amplifier output (in volts).

The second of the two definitions is preferred, because this is the one used by manufacturers to state the linearity of the 10-turn controls with which amplifier linearity is usually measured. In fact, Eq. (6) can be simplified considerably if the voltage range of the discriminator equals V_3. This is usually the case with built-in discriminators. Thus V_1, V_2, and V_3 are expressed in terms of pulse-height-control dial divisions, and V_3 corresponds to 1000 dial divisions. Equation (7) reduces to

$$IN_{max} = \frac{V_2 - V_1}{10} , \tag{8}$$

where V_2 and V_1 are expressed in pulse-height-selector-control dial divisions.

1.3.2 Graphical Display of Integral Nonlinearity

Linear amplifiers presently available exhibit integral nonlinearities of well under 1%. If the data are plotted as in Fig. 1.3, the deviation from perfect linearity is so slight that it cannot be measured on the graph paper. The preferred display is one in which the DEVIATION from perfect linearity is plotted vs amplifier output, as in Fig. 1.4. These numbers are obtained from Eq. (8). Points below the abscissa indicate an amplifier output less than it should be; those above the abscissa indicate the opposite.

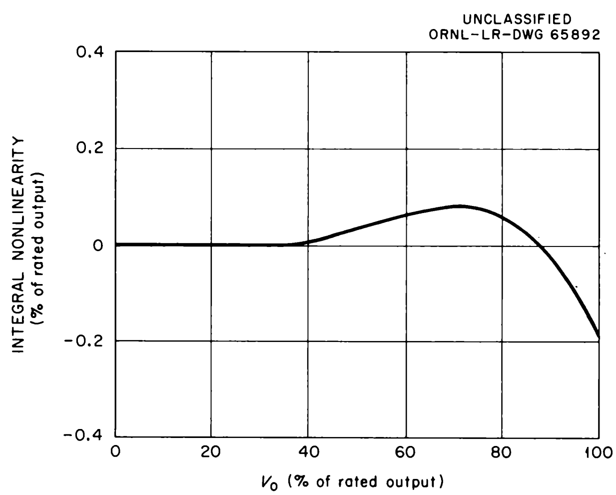


Fig. 1.4. Preferred Plot of Amplifier Nonlinearity (Integral) vs Amplifier Output.

1.3.3 Definition of Differential Nonlinearity

Differential nonlinearity is the variation in amplifier gain resulting from changes in amplitude of the output signal.

At any particular output level, the gain is defined as $\Delta V/\Delta e$, where ΔV is the change in output level resulting from a small change in the input voltage, Δe. This is illustrated in Fig. 1.5 where $\Delta V/\Delta e$, the spot gain of the amplifier, is also the slope or derivative (hence the term, "differential nonlinearity") of the dynamic characteristic.

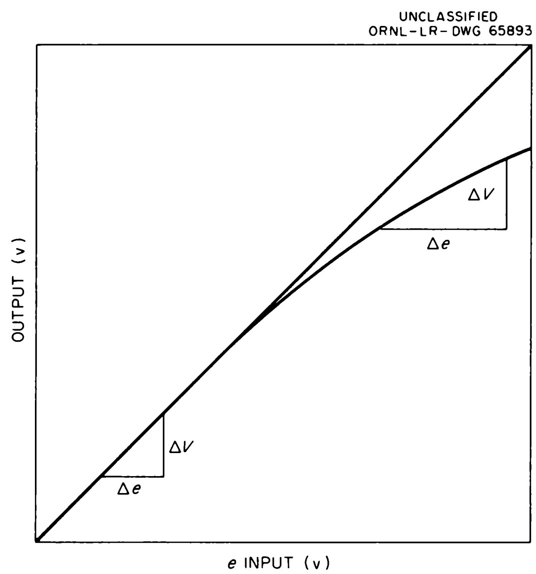
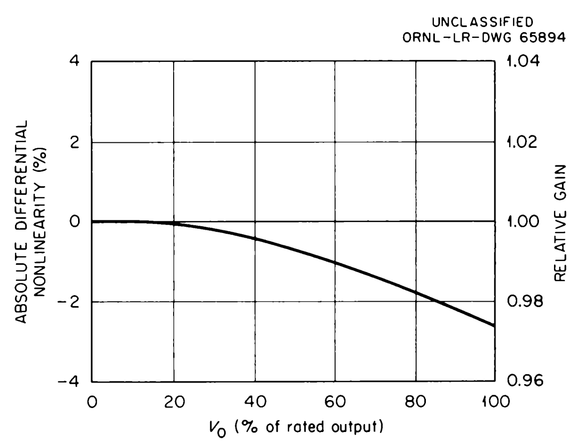


Fig. 1.5. Dynamic Characteristic of a Typical Pulse Amplifier with Axes Chosen to Produce a Slope of 1.00 for the Characteristic of a Perfect Amplifier.

1.3.4 Graphical Display of Differential Nonlinearity

With an appropriate choice of axes, the slope of the ideal characteristic curve can be made unity, and the slope of the actual response curve can be referred to the ideal curve in terms of relative gain. A typical differential nonlinearity curve is shown in Fig. 1.6. This curve was obtained from Fig. 1.5 by plotting $\Delta V/\Delta e$ as a function of the percentage of rated output. Note that the ordinate can be labeled in either of two ways. The "relative gain" title is preferred because it avoids ambiguity with regard to the sign of the nonlinearity. (The gain is relative to an arbitrary point on the dynamic characteristic curve – zero output volts, in this case.)

Fig. 1.6. Differential Nonlinearity of a Typical Pulse Amplifier.



1.3.5 Effect of Differential Nonlinearity on the Shape of the Spectrum

In the region of operation where the relative gain is less than 1.00, the pulse-height analyzer will accumulate more counts than it should, and in regions of high relative gain the pulse-height analyzer will accumulate fewer counts than it should. This is best shown by an extreme case. Consider an amplifier which saturates sharply at 90 v. It follows that all pulses having amplitudes ordinarily exceeding 90 v will accumulate in the pulse-height-analyzer channel centered on 90 v, giving rise to a false peak.

1.3.6 Integral Nonlinearity: Procedures

Two methods of measurement are described. Compared with method II, measurement by use of method I is simpler and faster, and the measurement error is usually less than 0.2%. Measurement by use of method II requires more sophistication on the part of the user, but less instrumentation is required and higher accuracy can be obtained.

Measurement of Nonlinearity, Method I

Instruments Required. – In addition to the amplifier under test, the following instruments are required:

1. Oscilloscope (optional).
2. Mercury relay pulser, Q-1212C or Q-1212D (see ORNL CF-60-9-67).
3. Integral discriminator with a 0.1%, 10-turn pulse-height-selector-control dial. This may be a component part of the amplifier under test, a Schmitt discriminator built for the purpose, or the discriminator of another amplifier patched into the unit under test.
4. Count-rate meter (crm).
5. Line-voltage regulator (or a regulated power line).

Discussion of Method I. – In general terms, the test is performed by applying a pulse-generator signal to the amplifier while measuring the output level with a pulse-height discriminator, using the rate meter as an indicator. Nonlinearity is determined from a series of pulse-height discriminator readings at various output levels. The test is particularly easy to perform if the dial readings of the generator and the discriminator are made to track each other in the linear portion of the dynamic range, because it is then easy to plot deviations vs dial readings.

Procedure. – The step-by-step test procedure follows below.

1. Connect the instruments according to the block diagram shown in Fig. 1.7. Let all instruments come to temperature equilibrium before proceeding with the test. The rise-time spoiler is used to simulate the pulse shape from slow detectors; for example, NaI(Tl) scintillation detectors have a time constant of 0.25 μsec, requiring a capacitor of 0.005 μf in the location shown.

2. Check the mechanical zeros of the pulse generator and discriminator dials. Note: In most cases the electrical and mechanical zeros of the pulse-generator dial do not coincide. With the control turned to the counterclockwise stop, the dial reading should be 0.5 of a small division. This can be checked with the aid of a millivoltmeter connected to the appropriate test terminals. (The pulse-generator-dial calibration is 10 mv per small dial division.)

3. If the discriminator has a zero adjustment, set it as follows: (1) set the pulse-generator and discriminator dials to 500.0. (2) Adjust the amplifier gain controls and the pulse-generator normalize control to the half-triggering-point (htp) of the discriminator, as indicated by a 30-cps reading – half the generator frequency. (Such a point can always be found, because amplifier noise spreads the counting threshold over one or more dial divisions.) (3) Set the pulser dial to 250.0 and the discriminator dial to the htp. If these two dial readings do not agree, note the difference in readings.

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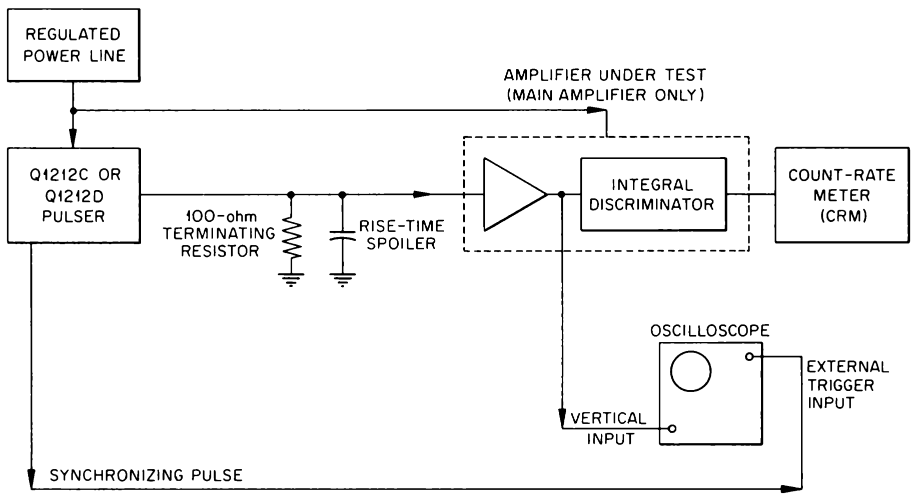

Fig. 1.7. Equipment for Testing Linearity.

When a discriminator other than one in a DD2 or a Q-1151-1 amplifier is tested and the two dial readings do not agree at the 250.0 mark, set the discriminator dial to a point on the other side of the 250.0 mark such that the difference between the new reading and 250.0 is half the previously noted error. (Example: With the pulser set at 250, a typical discriminator reading at the htp might be 240. The error is 10 divisions. Set the discriminator dial to 255.) Next, change the zero adjustment until the htp is established. Repeat the whole procedure until the discriminator reading corresponds to the pulser reading at both the 500 and 250 dial marks.

When a discriminator in a DD2 or a Q-1151-1 amplifier is tested and the two readings do not agree at the 250 dial mark, set the discriminator dial to a point on the opposite side of the 250 mark by an amount equal to the full error. (In the example above, the DD2 or Q-1151-1 dial would be set to 260.)

The foregoing procedure results in a rapid setting of the zero adjustment. The 500 and 250 dial marks were chosen because most amplifiers are linear in this region, and the extension of the dynamic characteristic curve through these points will pass very nearly through the origin.

4. If the discriminator has no zero adjustment, such as that of the A1D amplifier, do the following: perform step 3 of the procedure, note the error at the 250 mark, and double the error. The doubled value is the zero offset of the discriminator dial. If the error is on the *low* side of the 250 mark, for all future tests

set the discriminator to a dial reading *lower* than the desired reading by an amount equal to the doubled error. [Example: With the pulser and discriminator dials set at 500, adjust the gain and normalize controls for the htp. Set the pulser at 250 and the discriminator for the htp. Note the dial reading. Let us assume that it is 240. Twice the discrepancy is 2(250 – 240) = 20. For all future tests, set the discriminator dial 20 divisions lower than the desired setting. To check this, set the discriminator at 480, corresponding to a desired value of 500, and the pulser at 500. Adjust the normalize control for the htp. Set the generator at 250 and the discriminator for the htp. Assuming that no errors were made in the earlier readings, the discriminator dial should read exactly 230.]

5. After the offset error has been reduced to zero or its value noted, the remainder of the linearity test consists of taking a set of readings of differences between pulser and discriminator dial settings at the htp. For rough checks, four readings at the 250, 500, 750, and 1000 dial marks are sufficient. (The readings at 250 and 500 will already have been set to zero.) Where greater accuracy is required, readings should be taken at every 100 divisions, including the 50 and 950 dial marks.

Below the 50 dial mark the discriminator can be expected to be nonlinear.

Accuracy of Method I. – The 10-turn potentiometers used in discriminators and pulsers at ORNL have linearity tolerances of ±0.1% of the full-scale reading (±1 dial division on the standard 10-turn dial). Since two dials are involved in nonlinearity measurements, the error in any single reading can be as much ±2 dial divisions (±1 each for the pulser and discriminator dials). If it should happen that these maximum errors occur at the critical points of 250 and 500 on the dials (critical because these points are used to establish the zero setting of the discriminator dial), a pessimistic nonlinearity measurement will result. This is illustrated by the following situation.

Consider the readings in Table 1.4. These are plotted in Fig. 1.8, in which a best line was drawn through the points. At first glance, it would appear that the amplifier nonlinearity is 0.5%, or 5 dial divisions (at maximum output). However, a study of Secs 1.3.1 and 1.3.2 shows that amplifier nonlinearity must result in a curved, rather than straight, line when plotted as in Fig. 1.8. The straight line of nonzero slope in Fig. 1.8 is a result, not of amplifier nonlinearity, but of a false discriminator zero setting resulting from potentiometer errors at the critical 250 and 500 dial settings.

If the line obtained in Fig. 1.8 were curved, amplifier nonlinearity would be indicated – of magnitude determined by the maximum deviation of the curved line from a straight line drawn through the curve and intersecting the curve at the 250 and 500 points.

The foregoing example, though unlikely to occur in practice, emphasizes the importance of plotting the data in cases wherein the nonlinearity appears to exceed that of the control potentiometers.

Measurement of Integral Nonlinearity, Method II

Instruments Required. – The test instruments required are:

1. Oscilloscope.
2. Mercury relay pulser, Q-1212C or Q-1212D.
3. Integral discriminator (optional).

Table 1.4. Example of Amplifier Test Data

Discriminator Dial Setting	Pulser Dial Setting	Deviation (dial divisions)
50	52.5	−2.5
100	102.0	−2.0
200	202.0	−2.0
250	250.0	0
300	300.5	−0.5
400	399.5	+0.5
500	500.0	0
600	598.0	+2.0
700	698.0	+2.0
800	797.0	+3.0
900	896.5	+3.5
1000	995.0	+5.0

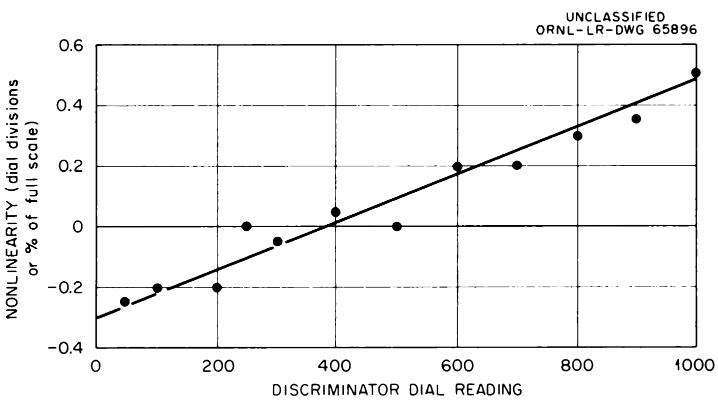

Fig. 1.8. Graph of Sample Nonlinearity Measurement Data Shown in Table 1.4.

Discussion of Method II. – This is an indirect method capable of high accuracy if carefully performed. It is applicable only to feedback amplifiers. The method is based on a measurement of the loop gain, $\mu\beta$, where β is the fraction of the output signal fed back to the input and μ is the forward gain of the amplifier. From this measurement the nonlinearity can be determined.

The loop gain is measured at a succession of output voltages. One output voltage is selected as a reference point in a region where the loop gain is constant (indicating linear operation), and the nonlinearity is computed from the following formulas (the derivation is given in Sec 1.3.7):

$$IN_{abs} = 100\,\frac{(\mu\beta)_2 - (\mu\beta)_1}{(\mu\beta)_2[1 + (\mu\beta)_1]}\,, \tag{9}$$

or

$$IN_{max} = 100 \frac{V_2}{V_3} \frac{(\mu\beta)_2 - (\mu\beta)_1}{(\mu\beta)_2 [1 + (\mu\beta)_1]} , \tag{10}$$

where

IN_{abs} = integral nonlinearity in percentage of absolute amplifier output,

IN_{max} = integral nonlinearity in percentage of maximum rated amplifier output,

$(\mu\beta)_1$ = loop gain at a reference point on the linear portion of the dynamic characteristic curve,

$(\mu\beta)_2$ = loop gain at some other output voltage, V_2,

V_3 = rated maximum amplifier output, v.

The loop gain, $\mu\beta$, is determined in two steps:

1. A signal of known amplitude is applied to the input of the amplifier, and the output voltage is measured with a discriminator or oscilloscope. The two voltages are related by

$$\frac{V_o}{V_i} = \frac{\mu}{1 + \mu\beta} , \tag{11}$$

where V_o is the output voltage and V_i is the input voltage.

2. The feedback is disabled by the method described in Sec 1.4.3, and the input voltage (V_i) is reduced to V_i', the value that results in the same output voltage as before. With the feedback loop broken,

$$\frac{V_o}{V_i'} = \mu . \tag{12}$$

Dividing Eq. (12) by Eq. (11) results in

$$\frac{V_i}{V_i'} = 1 + \mu\beta , \tag{13}$$

from which $\mu\beta$ can be determined.

The test is performed on the last feedback group in the amplifier, since this is the principal source of nonlinearity. (In some cases it may be necessary to test also the feedback section immediately preceding the gain control, since at minimum gain setting the signal swing in this part of the amplifier may be sufficient to cause nonlinearity.)

The output cathode follower, if separate from the main feedback group, can also be a source of nonlinearity. Cathode followers, however, are difficult to test by the method described here, and it is recommended that their nonlinearities be computed from measured tube parameters. A discussion of this technique is beyond the scope of this report.

Procedure. – The test setup is the same as that shown in Fig. 1.7 except that the count-rate meter is unnecessary – the oscilloscope is a sufficiently accurate indicator of the htp in this case.

When high accuracy is required, the discriminator is used for measuring the pulse height after the discriminator has been zeroed according to the directions in method I, "Procedure." When lower accuracy is sufficient or when the discriminator is not available, the oscilloscope can be used for measuring the amplifier pulse height.

The step-by-step procedure follows below.

1. The generator is set for the desired pulse height, and the dial reading is noted.

2. The feedback is disabled by the methods described in Sec 1.4.3, and the generator signal is reduced to a setting that restores the original output pulse height. For high accuracy, rated maximum output should occur at a dial reading in excess of 500 dial divisions. In reducing the generator output, the attenuator switches should be used so as to maintain this high dial reading. The new generator setting is recorded. When an oscilloscope is used as the pulse-height indicator, a change in pulse shape will be observed when the feedback is disabled. The peak of the pulse should be used in determining pulse height.

3. Repeat the measurement at a succession of pulse heights.

The results are tabulated (see Table 1.5 for typical data) for substitution in Eq. (10), which is repeated here for convenience:

$$\text{Integral nonlinearity in percentage of rated maximum output voltage} = 100\,\frac{V_2}{V_3}\,\frac{(\mu\beta)_2 - (\mu\beta)_1}{(\mu\beta)_2\,[1 + (\mu\beta)_1]}\ .$$

Table 1.5. Typical Data from Measurement of Integral Nonlinearity by Method II

V_0 (v)	V_i (dial divisions)	V_i' (in units of V_i)	$\mu\beta$ $\left(= \frac{V_i}{V_i'} - 1\right)$	$(\mu\beta)_2 - (\mu\beta)_1$	Absolute Nonlinearity (%)	Nonlinearity (% of 100 v)
(1)	(2)	(3)	(4)	(5)	(6)	(7)
99	1000	91.6	9.9	−1.6	−1.29	−1.28
75	750	66.1	10.8	−0.7	−0.519	−0.389
50	500	42.0	11.2	−0.3	−0.214	−0.107
25	250	20.0	11.5	0	0	0

The data were tabulated in the following manner. The measured output voltage was entered in column (1), and the dial reading of the pulse generator was entered in column (2). After the feedback loop was disabled, the new generator readings were entered in column (3). All the dial readings shown in column (3) are below 100. Actually, dial readings 10 times higher were obtained by using one of the 10× attenuator switches on the pulse-generator panel. It is necessary that the step-attenuator settings be used for all data points in a particular nonlinearity determination. If different steps are used during the measurements, discrepancies between steps will give misleading results. The computed values of

$\mu\beta\{[\text{column (2)} \div \text{column (3)}] - 1\}$ were entered in column (4). The reading (11.5) at 25-v output was selected as $(\mu\beta)_1$, and the differences between 11.5 and the remaining $\mu\beta$ values in column (4) were entered in column (5). The nonlinearities in column (6) were computed from Eq. (10), disregarding the term V_2/V_3. The values in column (7) were calculated by multiplying the values in column (6) by V_2/V_3, where V_3 was assumed to be 100 v and V_2 was the actual output voltage.

Accuracy of Method II. – Greater accuracy is the principal advantage of method II over method I. The linearity of the oscilloscope or the discriminator does not affect the accuracy of the linearity determination, because the observation point on the oscilloscope face or discriminator dial is not changed during a measurement of a particular value of $\mu\beta$. Also, small errors introduced by the nonlinearity of the pulse-generator dial or by resetting the output voltage are reduced in importance by the factor $[1 + (\mu\beta)_1]$. In the test illustrated in Table 1.5, the accuracy is improved 12.5 times, which is sufficient to permit a determination of the differential linearity from the slope of the absolute integral linearity curve obtained by this method of measurement.

1.3.7 Derivation of Equation (9)

Let V_1 be the output voltage of the amplifier that would result if the amplifier were perfectly linear, and let V_2 be the actual output voltage. From Eq. (5),

$$\frac{V_1}{V_i} = \frac{1}{\beta} \frac{(\mu\beta)_1}{1 + (\mu\beta)_1} \tag{14}$$

and

$$\frac{V_2}{V_i} = \frac{1}{\beta} \frac{(\mu\beta)_2}{1 + (\mu\beta)_2} \ , \tag{15}$$

where V_i is the input voltage, $(\mu\beta)_1$ is the loop gain at a linear region of the dynamic characteristic, and $(\mu\beta)_2$ is the loop gain at V_2. Substituting for V_1 and V_2 in Eq. (4) gives:

$$IN_{\mathrm{abs}} = 100 \frac{\frac{1}{\beta}\left[\frac{(\mu\beta)_2}{1 + (\mu\beta)_2}\right] V_i - \frac{1}{\beta}\left[\frac{(\mu\beta)_1}{1 + (\mu\beta)_1}\right] V_i}{\frac{1}{\beta}\left[\frac{(\mu\beta)_2}{1 + (\mu\beta)_2}\right] V_i} \ . \tag{16}$$

This reduces to Eq. (9):

$$IN_{\mathrm{abs}} = 100 \frac{(\mu\beta)_2 - (\mu\beta)_1}{(\mu\beta)_2 [1 + (\mu\beta)_1]} \ .$$

Equation (9) differs from Eq. (10) only in not being normalized with respect to V_3, the rated maximum output.

Equation (11) may be written as

$$IN_{abs} = 100 \frac{1 - [(\mu\beta)_1/(\mu\beta)_2]}{1 + (\mu\beta)_1} . \tag{17}$$

Written in this way, it can be seen that the numerator is the nonlinearity in the absence of feedback, and the denominator is the factor by which the feedback reduces the nonlinearity.

1.3.8 Measurement of Differential Nonlinearity

The two principal methods of determining differential nonlinearity are by taking the derivative of the absolute integral nonlinearity curve and by using a sliding pulser. The first method has been discussed, and the remainder of this discussion will be concerned with the accuracy of this method.

Accuracy with Which an Absolute Integral Nonlinearity Measurement Can Be Converted to a Differential Nonlinearity Figure

In principle, the simplest means of determining the slope of an integral linearity curve is to take successive differences of pulse-generator and discriminator dial readings by use of method I, Sec 1.3.6, but this method is not sufficiently accurate. The rated linearity tolerance of a 10-turn potentiometer such as used at ORNL is ±0.1% of full scale (i.e., ±1 minor dial division). If successive differences should be 50 dial divisions apart, the possible error in the determination becomes 4 dial divisions in 50 (1 dial division at each of 4 boundaries), or 8%, which is obviously intolerable.

A better way of using the same data is to take many closely spaced points and to plot the best average integral linearity curve on graph paper, bearing in mind that conventional amplifiers do not exhibit abrupt changes in linearity. After the curve has been drawn, its slope can be determined on a point-by-point basis and a new curve derived.

The slope of the curve can be accurately determined with the aid of a round clear-plastic rod. At the point where the slope is to be determined, a pencil mark is placed on the curve approximately perpendicular to the curve. The rod is then placed over the curve with the long axis of the rod tangent to the curve (Fig. 1.9). The magnifying properties of the cylinder will emphasize the curvature and will also spread the

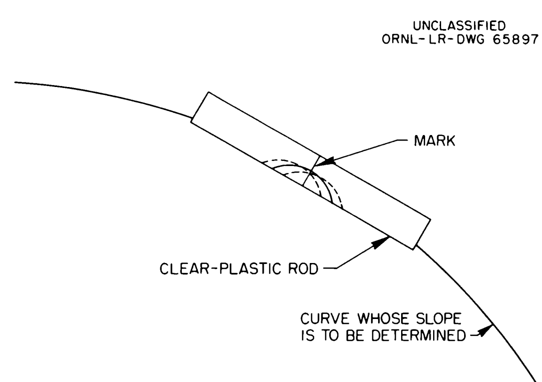

Fig. 1.9. Method for Establishing the Slope of a Curve.

mark over the diameter of the rod. If the rod is not tangent to the curve, the curve will occupy one of the dotted positions shown in Fig. 1.9. When the rod is tangent to the curve, the curve will be symmetric with the perpendicular tick mark, as shown by the solid line in Fig. 1.9. The rod can be rolled to convenient calibrating points on the axes to determine the slope.

A differential nonlinearity curve determined by the above method should be accurate to about 1%. If an integral discriminator is used to determine individual points and (1) if the discriminator is not preceded by a dc-coupled amplifier or (2) if the discriminator is not dc coupled between the input and output tube with a frequency-compensated divider, points below 5-v input (to the discriminator) should not be used because they will be nonlinear.

A differential linearity curve derived from the data obtained by method II, Sec 1.3.6, can be obtained in the same way as just described. Reading errors will be reduced by $(1 + \mu\beta)$.

Differential Nonlinearity Measured with a Sliding Pulser

The second method for measuring differential nonlinearity involves the use of a pulser in which the pulse height is changed slowly and at a very linear rate over the amplifier output range of interest. The pulser is used with a differential discriminator or multichannel pulse-height analyzer. When the pulse amplitude exceeds the lower edge of a channel, counts are recorded in the channel. The counting continues until the pulse amplitude exceeds the upper boundary. The number of counts recorded in any channel depends on the width of the channel, the differential linearity of both the amplifier and pulser in that region, and the pulse repetition frequency.

Since the success of the method depends on the differential linearity of the sweep circuit in the pulser, circuits which use a motor-driven potentiometer to generate the sweep should not be used. An electronically generated sweep, such as that used in the Q-1450A pulser, is much to be preferred.

1.4 FEEDBACK AMPLIFIERS

The block diagram of a typical feedback amplifier is shown in Fig. 1.10. By feeding back to the input a portion of the output in a way that causes the two signals to oppose each other at the summing

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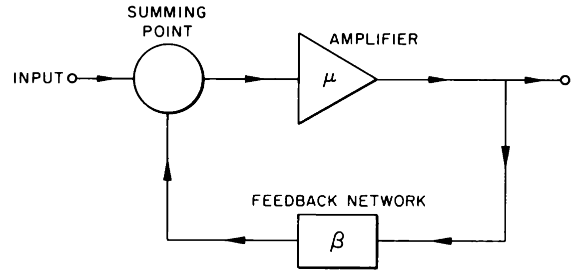

Fig. 1.10. Typical Feedback Amplifier.

point, some desirable characteristics are obtained for the system – principally, good gain stability, linearity, and speed of response. These characteristics are paid for with a loss of gain, however. Up to a point, the greater the gain loss the better the overall performance of the system. (In feedback-amplifier terminology, gain loss is referred to as excess gain, or feedback factor.)

1.4.1 Terminology

In the following discussion, the terms used to describe feedback systems are enclosed with quotation marks.

The "feedback loop" consists of the path around the "μ" and "β" boxes in Fig. 1.10. The "forward gain" of the system is denoted by "μ" and is the gain with the feedback loop broken. The "feedback ratio" is denoted by "β" and is the fraction of the output signal fed back to the input. The "loop gain", as the name implies, is the gain around the loop, or $\mu\beta$. The "summing point" may be the grid or cathode (base or emitter) of the input tube (transistor), depending on the type of feedback network.

With the loop closed the gain of the system is

$$A = \frac{\mu}{1 + \mu\beta} , \tag{18}$$

or

$$A = \frac{1}{\beta} \frac{\mu\beta}{1 + \mu\beta} . \tag{19}$$

(This assumes negative feedback.) The denominator $1 + \mu\beta$ is the feedback factor and is the factor by which gain changes in the forward path are reduced by the feedback action. As an example, let $\mu = 1000$ and $\beta = 0.1$; then, $\mu\beta = 100$ and $1 + \mu\beta = 101$. Suppose that because of tube aging, μ drops to 500, representing a 50% change in the unfed-back gain. When $\mu = 1000$, from Eq. (19)

$$A_1 = \frac{1}{0.1}\left(\frac{100}{101}\right) = 9.9 .$$

When $\mu = 500$,

$$A_2 = \frac{1}{0.1}\left(\frac{50}{51}\right) = 9.8 .$$

Evidently, feedback has reduced the gain variation from 50% to 1%, and the overall gain is very nearly equal to $1/\beta$. If the β network is composed of low-temperature-coefficient resistors and capacitors, very stable gain systems can be built.

1.4.2 Measurement of Feedback Factors

Since the feedback factor is directly related to the forward gain of the loop, it can be used as a sensitive indicator of tube (or transistor) condition. In most cases the measurement is easily made and yields a more reliable indication of tube condition than a tube tester.

A typical test setup is shown in Fig. 1.11. (In high-gain systems, when measurements are being made on the main amplifier, it may be desirable to feed the generator signal directly to the main amplifier.)

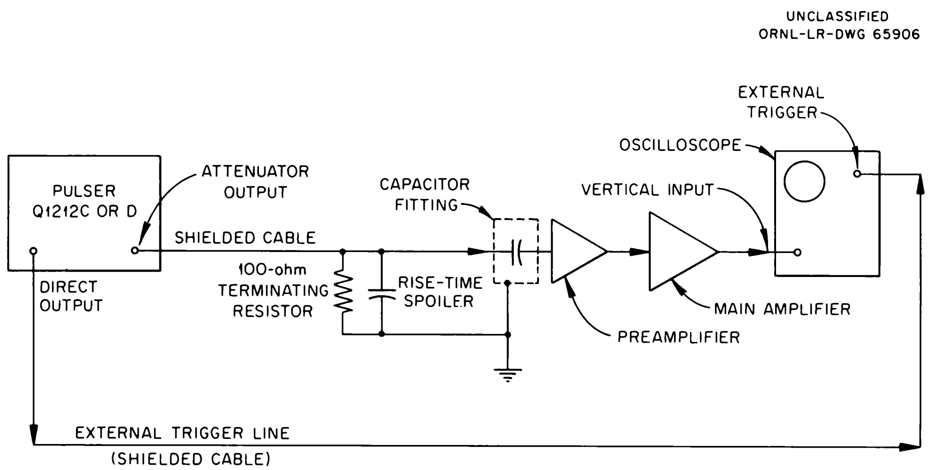

Fig. 1.11. Test Setup for Measuring Feedback Factors. See Fig. 1.16 for makeup of the capacitor fitting.

The output signal is adjusted to a level between 50 and 100 v (the amplifier must not be saturated), and the generator dial reading is noted. The feedback is then disabled in a way which does not affect the circuit loading (this point will be discussed further), and the generator output signal is reduced to the point where the original output signal is obtained. The ratio of the two generator signals is then the feedback factor, $1 + \mu\beta$.

Proof: From Eq. (18),

$$V_o = \frac{\mu}{1 + \mu\beta} V_1 \, , \tag{20}$$

where V_o is the output voltage and V_1 is the initial generator reading. When the feedback is disabled, $\beta = 0$; therefore

$$V_o = \mu V_2 \, , \tag{21}$$

where V_2 is the new generator reading required to re-establish the same peak output voltage. By equating the preceding expressions, it follows that

$$1 + \mu\beta = V_1/V_2 \, . \tag{22}$$

1.4.3 Disabling the Feedback Loop: Summing Point at Cathode

A typical feedback group is shown in Fig. 1.12. The β network consists of R_k, R_f, and the trimming capacitor, C_f. The nominal value for β is $R_k/(R_k + R_f)$, where R_f is usually much greater than R_k. In simple cases, the feedback can be disabled by shunting R_k with a large capacitor (25 μf at 6 v).

Fig. 1.12. Typical Feedback Group with the Summing Point at the Cathode.

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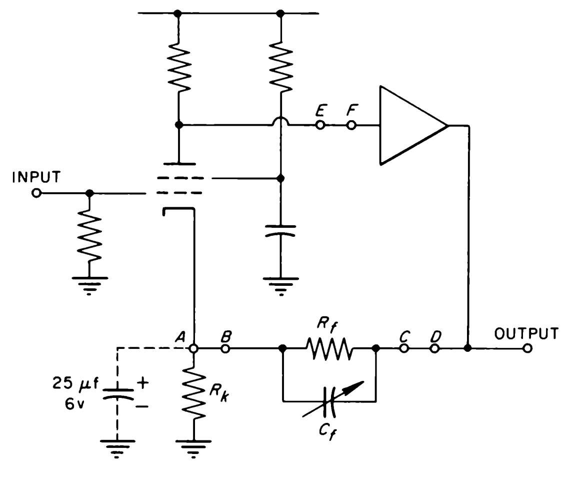

At this point it is necessary to digress. It was stated in Sec 1.4.2 that in measuring the feedback factor, the feedback must be disabled in a way that does not affect the circuit loading. The success of the method hinges on this point, and the following detailed qualitative analysis of the circuit shown in Fig. 1.12 should enable the reader to comply with the necessary test conditions in a variety of cases.

In the circuit of Fig. 1.10, the μ and β paths are shown separately with summing points at input and output (the summing point at the output is not explicitly shown). In practice, the μ network may load the β network, and vice versa, so that the distinction between μ and β paths may not be ideal.

The following statements are based on a detailed quantitative analysis (which will not be given here) of the circuit of Fig. 1.12:

1. The β network (consisting of R_k, R_f, and C_f) loads the μ network, but the reverse is not true. If the output stage in the μ path is a cathode follower (output impedance 500 ohms or less), the loading effect of the β network will almost always be negligible. If the output stage is a plate amplifier, the loading effect of the β network will almost always be appreciable. If the plate load resistance is R_1, the effective load resistance becomes $R_1 \,||\, (R_k + R_f)$, where $||$ means in parallel with.

When the feedback is disabled, it should be done in such a way as to keep the loading constant. This may require the insertion of additional resistors (between points C and D in Fig. 1.12, for example) when parts of the feedback network are short-circuited.

2. To the input stage, the feedback network appears as an unbypassed cathode resistor of magnitude $R_k \,||\, (R_f + Z_o)$, where Z_o is the output impedance of the μ path. (The capacitor C_f is nearly always so small as to look like an open circuit to the input stage.) This represents local feedback to the input tube, and its gain will be reduced in the ratio

$$1 : \{1 + [g_m + g_s][R_k \,||\, (R_f + Z_o)]\}\,,$$

where g_m is the plate transconductance and g_s is the screen transconductance.

Let us consider an example. A 6AK5 tube is used for an input stage. Plate current = 3 ma; the corresponding g_m is approximately 3 ma/v. Screen current (therefore g_s) is one-third as large, so that $g_m + g_s = 4$ ma/v. $R_k = 470$ and $R_f = 15$ kilohms. This combination results in an overall feedback group gain of (15 + 0.47)/0.47 = 33. Assuming a cathode-follower output stage having negligible Z_o, the effective cathode resistor is 470 || 15 kilohms = 455 ohms. Therefore $1 + [g_m + g_s]\,[R_k \,||\, (R_f + Z_o)] = 1 + (4 \times 0.455) = 2.82$.

In the foregoing example, it is seen that the presence of the feedback resistors reduces the input-tube gain by the factor 2.82. If the feedback is disabled by short-circuiting R_k, both overall and local feedback are removed. This may be important when making a measurement of nonlinearity by the method described in Sec 1.3.6, "Measurement of Integral Nonlinearity, Method II." LOCAL FEEDBACK IN THE INPUT STAGE DOES NOT IMPROVE THE LINEARITY OF THE OUTPUT STAGE. (The linearity of the feedback group is almost always controlled by the linearity of the output stage.)

In Fig. 1.12, the overall feedback may be disabled without affecting the local feedback by breaking the connection between A and B and connecting B to ground. It is necessary to do this when making nonlinearity measurements by method II of Sec 1.3.6, in order to keep from getting an optimistic (and erroneous) nonlinearity figure.

Note that in the example cited, the parallel combination of R_k and R_f gives a value for the cathode resistance only 3% different from the value of R_k above. In disabling the feedback by the method described in the preceding paragraph, it is rarely necessary to go to the added trouble of shunting R_k with a resistor of magnitude R_f.

3. When the feedback between input and output is dc coupled, care must be taken in disabling the feedback to ensure that the dc operating point of neither the input nor output stage is disturbed. An auxiliary resistor network may be necessary to accomplish this.

4. The output terminal receives signals from the β path as well as from the μ path. The signal from the β path is known as the "direct feed-through" signal and is usually (but not always) negligible. The signal arises in the following way: In Fig. 1.12, assume the connection between E and F to be broken; this opens the feedback loop in the μ path. By cathode-follower action, a signal will appear at A having a gain (referred to the input) of $gZ_k/(1 + gZ_k)$, where g is the sum of plate and screen transconductances and Z_k is the cathode load. A fraction of this signal will appear at the output terminal, the fraction being determined by the attenuator formed by R_f and Z_o of the output stage. It is this signal which is the direct feed-through signal.

Since the direct feed-through signal is always less than the input signal, it need be considered only in those instances when the gain of the feedback group is very low (say 3 or 4).

1.4.4 Disabling the Feedback Loop: Summing Point at Input Grid

Feedback amplifiers in which the summing point is at the input grid are known as operational amplifiers, charge-sensitive amplifiers, or shunt-fedback amplifiers. A typical circuit is shown in Fig. 1.13.

The β path consists of the driving impedance of the generator, C_1, C_3, R_3, and the parasitic capacitance, C_2. The μ path consists of the generator impedance, C_1, and the tubes between A and V_o. Note that the generator impedance and C_1 are common to both paths. This fact complicates the measurement of the feedback factor. Another possible complication is the relatively large direct feed-through term.

The problem of measuring feedback factor resolves itself into separate measurements of μ and β.

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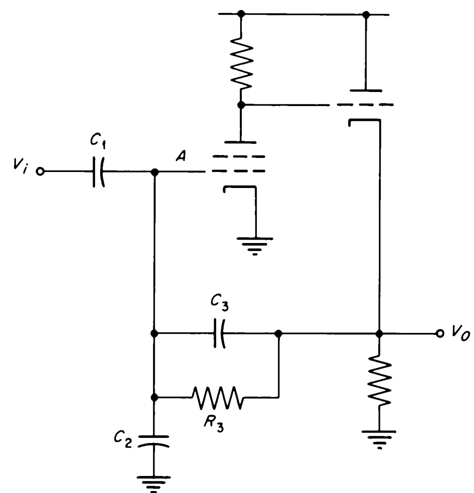

Fig. 1.13. Typical Circuit for a Feedback Amplifier with the Summing Point at the Input Grid.

Measurement of μ

It is assumed that the feedback amplifier is a charge-sensitive preamplifier (or an intermediate feedback group in a pulse amplifier); μ is not likely to be more than 20,000. (Operational amplifiers of the types used in computers have such very high gains – 10^7 to 10^8 – that the technique described here cannot be used; the amplifier noise would swamp the signals with the feedback disabled.)

If practical, C_1 should be short-circuited. If this changes the dc operating point of the input tube, C_1 should be shunted instead with a capacitor large compared to μC_3. This procedure assumes prior knowledge of μ; if only one stage is used, it is unlikely that μ exceeds 2000, even with positive feedback applied. If C_3 is 33 pf, μC_3 becomes 0.066 mf, and a 1-mf shunting capacitor is required. Alternatively, C_3 may be disconnected from the input grid and grounded, permitting a much smaller value of the shunting capacitor. In either case, the quantity μ is measured as the ratio between the output signal and a signal applied to the input grid. It is necessary also that the signal generator have an output impedance certainly no larger than 50 ohms and preferably no larger than 5 ohms.

Measurement of β

It is assumed that C_1 in Fig. 1.13 corresponds to the detector capacitance with which the preamplifier is used.

The quantity β is $C_3/(C_1 + C_2 + C_3)$. C_1 and C_3 are usually known, but C_2 may be unknown. In this case, turn the amplifier power off, ground V_o, and measure the pulse voltage at point A. The measured voltage will be $V_i C_1/(C_1 + C_2 + C_3 + C_p)$, where C_p is the oscilloscope probe capacitance, which is usually known. From this, C_2 and therefore β can be computed. (The value of C_2 should be increased by 2 to 3 pf to account for the hot capacitance of the input tube.) As an example:

Given

$$C_1 = 33 \text{ pf}, \; C_3 = 33 \text{ pf}, \; C_p = 10 \text{ pf}, \; V_i = 1 \text{ v}, \text{ and } V \text{ (at } A) = 0.4 \text{ v},$$

therefore,

$$0.4 = 33/(33 + 10 + 33 + C_2),$$

from which

$$C_2 = 6.5 \text{ pf}.$$

To determine β :

$$\beta = C_3/(C_1 + C_2 + C_3) = 33/(33 + 6.5 + 33) = 0.46 .$$

The determination of β can be made without measurement in an amplifier in which resistors are used in place of the capacitors C_1, C_2, and C_3: The value of β can be computed from the known resistance values.

In Fig. 1.14 it is assumed that the generator impedance is much smaller than R_1. By inspection,

$$\beta = \frac{(R_1 \,||\, R_2)}{(R_1 \,||\, R_2) + R_3} . \tag{23}$$

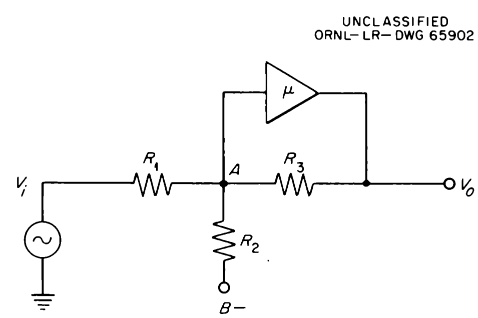

Fig. 1.14. Circuit for Computation of β from Known Resistance Values.

A point of interest is that the β network attenuates the generator signal seen by the μ portion of the circuit. In Fig. 1.14, for example, the voltage appearing at the summing point with the μ path broken is

$$V \text{ (at } A) = V_i \frac{(R_2 \,||\, R_3)}{R_1 + (R_2 \,||\, R_3)} . \tag{24}$$

This attenuation has no direct effect on the feedback factor.

1.5 MEASUREMENTS OF AMPLIFIER NOISE

In detectors that cannot in themselves amplify, such as ion chambers and the present generation of semiconductor detectors, the lowest-energy particle that can be detected is limited by the noise generated in the input stage of the amplifier. In certain spectral measurements, such as alpha-particle spectra in pulse ion chambers, the width of the spectral lines has a lower limit imposed by amplifier noise. For the foregoing reasons it is necessary to specify and measure amplifier noise in an unambiguous way that the user of the equipment can relate to his experiments.

1.5.1 Definitions and Terminology

If the output of an amplifier is observed with an oscilloscope under conditions of high sensitivity, "grass" will be seen on the baseline. This grass is noise generated in the input stages of the amplifier. At sweep speeds comparable to the high-frequency response of the amplifier, individual traces will be resolved in which the displacement from the baseline is random in time. At slow sweep speeds, individual traces will not be resolved on the oscilloscope screen; instead, a swath of light will be seen which diminishes in intensity with displacement (in either direction) from the baseline. If the intensity of light or the frequency of occurrence of specific pulse amplitudes is graphed vs displacement from the baseline (in volts), the familiar bell-shaped, normal (or Gaussian) statistical distribution will be obtained, as shown in Fig. 1.15.

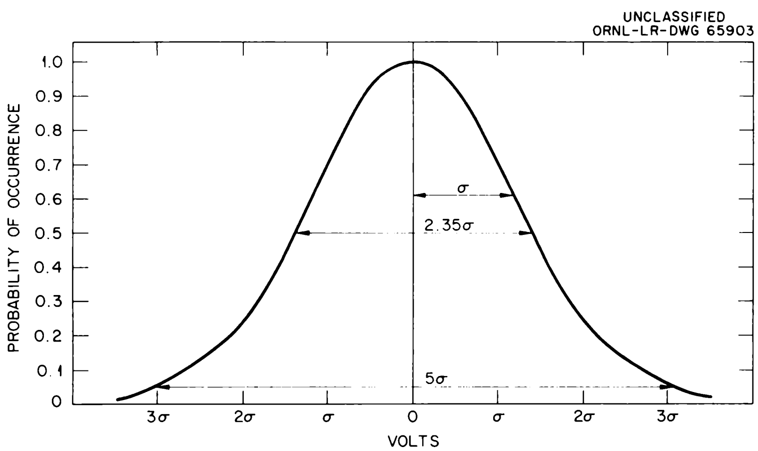

Fig. 1.15. Gaussian Distribution Curve.

1.5.2 Some Properties of a Gaussian Distribution

The relative probability of occurrence (normalized to the most probable value at zero volts) is

$$P = e^{-1/2(x/\sigma)^2} . \tag{25}$$

A voltmeter sensitive to the root-mean-square (rms) value of the applied voltage when connected to the output of a device producing a normal voltage distribution will be deflected by an amount proportional to the standard deviation, σ, of the distribution. The standard deviation and the rms value are synonomous, and both occur at a relative probability of 0.607.

Spectral lines with widths determined by amplifier noise will have full-widths-at-half-maximum (fwhm) of 2.35σ.

In estimating the noise voltage from an oscillograph observation, the eye tends to pick out the 0.05 probability point, which occurs at $\pm 2.5\sigma$; that is, the apparent peak-to-peak noise voltage as observed on an oscilloscope set for a fairly slow sweep rate will be approximately five times the rms noise voltage.

Noise voltages do not add linearly but as the square root of the sum of the squares. (Other ways of saying the same thing are "the voltages add in quadrature" or "the voltages add quadratically.")

The square of the rms value, $\bar{\sigma}^2$, is known as the mean-squared value. Mean-squared voltages add linearly.

A meter sensitive to the average value of the applied waveform, such as the Hewlett-Packard 400D, 400H, or 400L, will yield the rms value of a random noise voltage if its reading is multiplied by 1.13. This factor arises because the meter face was calibrated by the manufacturer to read the rms value of a pure sine wave, which has an rms-to-average ratio of 1.11. The rms-to-average ratio of a normal distribution is 1.26. The correction factor is 1.26/1.11 = 1.13.

The bandwidth of the Hewlett-Packard 400D meter is such that no more than a 10% error will occur if the amplifier cutoff frequency is determined by a time constant of 0.15 μsec or more.

Meters with response proportional to the peak value of the signal, such as the RCA VoltOhmyst and the H-P 410B, will give grossly inaccurate readings and should not be used.

1.5.3 Amplifier Noise: Definition

The experimenter usually knows in advance the quantity of charge or the number of electrons released in a detector from a radioactive interaction. Principally for this reason, amplifier noise should be defined in terms of an equivalent rms charge or an equivalent rms number of electrons ($Q_e = 1.6 \times 10^{-19} N_e$, where Q_e and N_e are, respectively, equivalent charge in coulombs and equivalent number of electrons). The use of equivalent rms volts or current is discouraged, partly because of the foregoing and partly because information about the internal structure of the amplifier must be given before the noise voltage has real meaning to the experimenter.

> Definition: Equivalent noise charge is that quantity of charge which, if deposited on the input terminal of the amplifier in a time short compared with the amplifier time constants, would produce a voltage pulse at the output just equal in peak amplitude to the rms noise voltage measured at that point.

Note that while the noise is measured at the output, it is defined in terms of an input signal. This avoids the necessity of stating the amplifier gain.

When the equivalent noise charge is given, it should be accompanied by a statement of the total external capacity connected to the input of the amplifier, since this capacity affects the signal-to-noise ratio (s/n) and is a figure which must be known in comparing amplifiers or in designing experiments.

1.5.4 Measuring Amplifier Noise

General Principles

The two principal methods of measuring amplifier noise involve the use of an ac voltmeter and a pulse-height analyzer. The first method is simpler and requires less equipment. In either case, extreme care should be taken to avoid extraneous noise pickup. All signal leads should be shielded, shield covers should be in place, only one of the group of test instruments should be grounded to the power line, and all test instruments should be plugged into the same group of power-line outlets.

Measuring Amplifier Noise with a VTVM

This method is based on the measurement of the peak-signal-to-rms-noise ratio at the output of the amplifier. Then, from the value of the applied input signal and with the assumption that the s/n ratio is the same at input and output, the input noise can be computed. The assumption is valid in the linear operating range of the amplifier.

Equipment. – The required equipment is:

1. A model Q-1212C or Q-1212D pulser.
2. A Hewlett-Packard 400D, 400H, or 400L oscilloscope.
3. Fittings having capacitors of 10, 33, 100, 330, and 1000 pf.
4. Connecting cables.

The capacitor fittings are necessary to feed a known charge into the amplifier and/or to simulate the detector capacitance.

A simple way of making the fittings is shown in Fig. 1.16.

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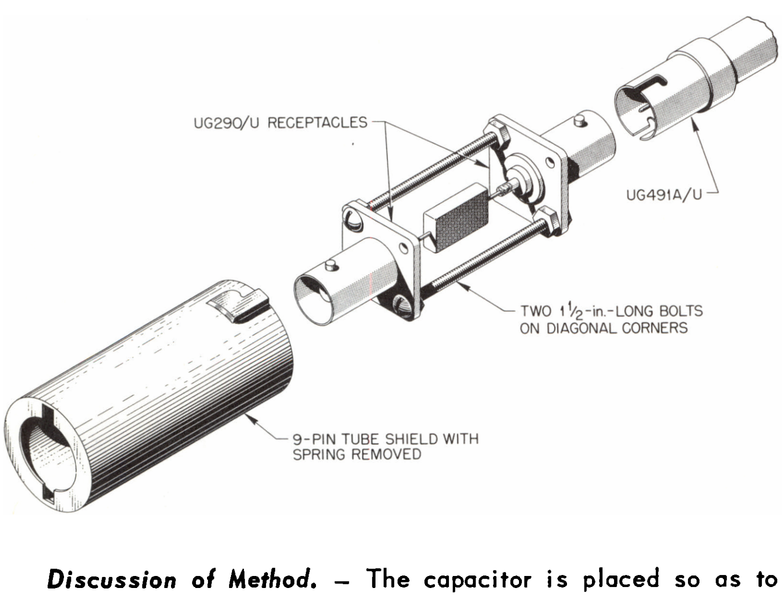

Fig. 1.16. Makeup of a Capacitor Fitting.

Discussion of Method. – The capacitor is placed so as to minimize its capacity to ground. The UG491A/U adaptor permits the fitting to be plugged into a BNC connector on the preamplifier. The nine-pin tube shield is a snug fit over the assembly and prevents unwanted pickup.

With the fitting connected to the amplifier, the amplifier input capacitance is increased by (C + 4.9 pf), where C is the capacitance of the series condenser. The equivalent circuit is shown in Fig. 1.17, which is also the circuit used for measuring the equivalent noise charge of the amplifier. The loading effect on the generator depends on the preamplifier circuit and on the value of C. In no case will the loading exceed C. An example of loading, which is probably extreme, is that in which a charge-sensitive preamplifier is used, C = 1000 pf, and the generator source impedance (including terminating resistor) is 50 ohms. The resulting time constant is 50 nsec.

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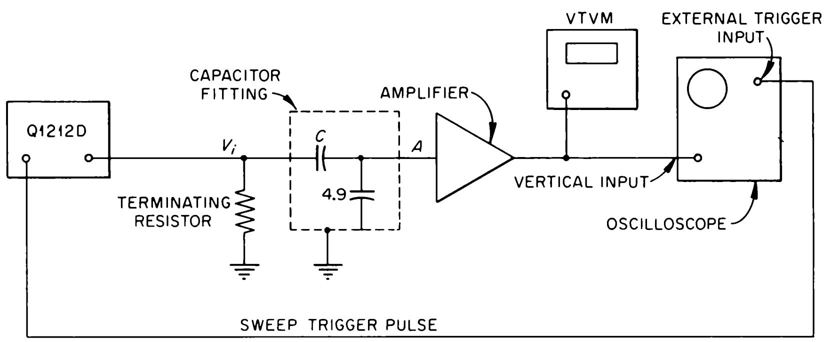

Fig. 1.17. Test Setup for Measuring Amplifier Noise.

In Fig. 1.17, the charge transferred to the input of the amplifier is

$$Q_i \text{ (coulombs)} = V_i C \, , \tag{26}$$

where

V_i = generator peak output signal, v,

C = capacitance, f.

The input capacitance of the amplifier does not affect Q_i. This is proved in the following manner.

Assume for the moment that the generator is grounded and that (C + 4.9) pf represents the detector capacitance. Let C_i be the input capacitance of the amplifier measured at point A in Fig. 1.17. If a quantity of charge Q is released in the detector, it will produce a voltage pulse at point A of magnitude

$$V \text{ (at } A) = A/(C_i + C + 4.9) \; . \tag{27}$$

The generator, on the other hand, will produce a pulse at point A of magnitude

$$V \text{ (at } A) = V_i C/(C_i + C + 4.9) = Q/(C_i + C + 4.9) \; . \tag{28}$$

By equating Eqs. (27) and (28), it is seen that the generator will produce the same signal as a detector pulse or a noise pulse if it is assumed that the charge transferred from the generator is $Q = V_i C$.

Procedure. – The noise test is made as follows:

1. With the generator on, adjust the signal level to produce a convenient output signal from the amplifier and measure this signal with the oscilloscope. Record the generator signal and the output pulse height from the amplifier.

2. Turn off the driving voltage to the relay; read and record the voltmeter deflection.
3. Compute the equivalent noise charge from the equation

$$Q_e \text{ (coulombs)} = (1.13\, v_n)\, V_i C/V_o \,, \tag{29}$$

where 1.13 is the constant which relates the actual voltmeter reading, v_n, to the rms value of a random noise voltage; V_i is the generator voltage; C is the series capacitance, in farads; and V_o is the peak voltage measured at the amplifier output with the oscilloscope.

A group of typical test readings obtained with an A8 amplifier (see Sec 2.3) is shown in Table 1.6.

Table 1.6. Typical Values for a Noise Measurement

C (pf)	V_i (mv)	Q_i (coulombs)	V_o (v)	$1.13v_n$ (v)	Q_e (coulombs)	N_e
		$\times 10^{-14}$			$\times 10^{-17}$	
10	5	5.00	108	1.42	66.0	4140
33	1	3.30	60	1.66	91.5	5720
100	0.5	5.00	90	2.39	133.0	8320

The equivalent noise in electrons is obtained by dividing Q_e by 1.6×10^{-19}, the charge per electron. The spectral line width (full width at half maximum) is obtained from

$$E_r \text{ (kev)} = 2.35 W N_e \times 10^{-3} \,, \tag{30}$$

or

$$E_r \text{ (kev)} = 1.47 W Q_e \times 10^{16} \,, \tag{31}$$

where W is the mean energy in electron volts necessary to produce an electron at the output terminal of the detector, and N_e and Q_e are, respectively, the equivalent noise of the amplifier in electrons and in coulombs.

Typical values of W are shown in Table 1.7.

Table 1.7. Typical Values of W

Detector	W (ev)
Gas-filled chamber, 90% argon and 10% methane	25
Gas-filled chamber, 99+% argon	30–35
Scintillation detector, NaI(Tl)	700–2000
Silicon detector	3.5

Measuring Amplifier Noise with a Pulse-Height Analyzer

This method is the same as the foregoing method except that a pulse-height analyzer replaces the vtvm.

A spectrum of the pulse-generator signal is obtained from which the equivalent noise voltage is computed by dividing the fwhm by 2.35 (see Sec 1.5.2).

1.6 OVERLOAD TESTS

The overload performance of most amplifiers can be checked with a Q-1212C or Q-1212D mercury relay pulse generator (see *Instruction Manual, Mercury Relay Pulse Generator, Model Q-1212C*, ORNL CF-60-9-67).

A typical test arrangement is shown in Fig. 1.18. With the generator terminated as shown, the maximum pulse height is 5.0 v, the impedance seen by the amplifier is 50 ohms, and the decay-time constant is 600 μsec. The maximum pulse height and source impedance are doubled if the DIRECT OUTPUT terminal is terminated instead. (The decay time remains the same.) If both terminals are terminated, the maximum pulse is 5 v, the fall-time constant is 300 μsec, and the source resistance is 50 ohms.

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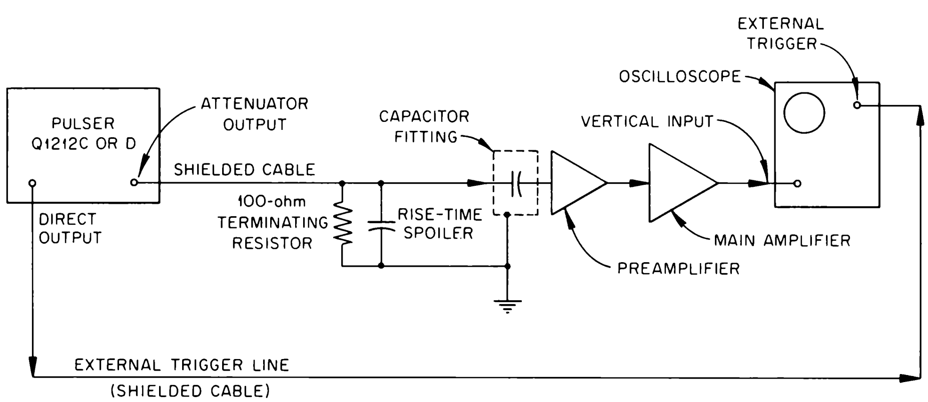

Fig. 1.18. Test Setup for Measuring Overload Performance. See Fig. 1.16 for makeup of the capacitor fitting.

The rise-time spoiler makes it possible to simulate the leading edge of a slow signal such as that from a sodium iodide scintillation detector (time constant equals 0.25 μsec; with a 100-ohm source resistance, C equals 0.0025 μf). The spoiler may be located at either end of the cable. To prevent ringing, the leads to the spoiling capacitor should be kept short, and the cable from the generator should be terminated at the amplifier end.

The series capacitor between generator and amplifier simulates the detector capacitance. In the absence of specific information, a value of 33 pf should be used and the capacitor fitting should be connected directly to the preamplifier input connector.

The rise time of the oscilloscope should be faster than that of the amplifier.

By triggering the oscilloscope externally, the position of the leading edge of the input pulse is made independent of the signal amplitude and attenuator setting of the generator; the position will change if the generator pulse-height control is varied, but in general it is unnecessary to change the pulse-height control during an overload test.

Capacitive feedback between output and input (via the oscilloscope and pulser) may cause instability when the amplifier is operated near full gain. When this occurs, the external trigger line must be removed and the internal triggering mode of the oscilloscope used. The effect of the external trigger connection on the system performance should always be checked to determine the safe operating range when an overload test is started.

With some amplifiers and pulsers, two artifacts may appear on the amplifier output signal under conditions of extreme overload. The first occurs 5 to 20 μsec from the leading edge of the pulse and appears as a small-amplitude satellite pulse having a lot of time jitter. This is caused by the closure of the second of the pair of bifurcated contacts in the mercury relay. The second artifact occurs halfway between main pulses and signifies the return of the relay armature to the charge position.

All signal-carrying leads should be shielded, and all test instruments should be plugged into the same group of power-line outlets.

1.6.1 Definition of Overload Factor

Times-one overload (or 1$\times$ overload) is considered to be rated amplifier output. It follows that 5$\times$ overload is that condition which would result in 5$\times$ rated output if the amplifier had no overload limit, etc.

1.6.2 Overload Tests with Input Pulses Greater Than 10 v

There are times when it may be necessary to test an amplifier with input signals of more than 10-v amplitude. When this is the case, it will probably be necessary to use a pulse having an exponential decay characteristic of time constant more than 600 μsec (to keep the generator pulse shape from affecting the overload response). These characteristics cannot be obtained from a standard Q-1212C or Q-1212D pulser, and an amplifying and shaping stage must be added (externally).

A satisfactory circuit is shown in Fig. 1.19.

The amplifier is designed for negative signals only. With the input terminated as shown, 50-v pulses can be obtained with an unloaded rise time of approximately 0.07 μsec and a decay time constant of 4 msec. The output may be loaded with 1 kilohm with no loss of performance. If the output is loaded with 100 pf, the rise time will be slowed to approximately 0.1 μsec. Unloaded output pulses of 100-v amplitude can be obtained if the input terminating resistor is removed.

The amplifier should NOT be powered from the amplifier under test if oscillation is to be avoided.

Because of the diode stretcher, the linearity is rather poor. This circuit should not be used for pulses below 1-v output.

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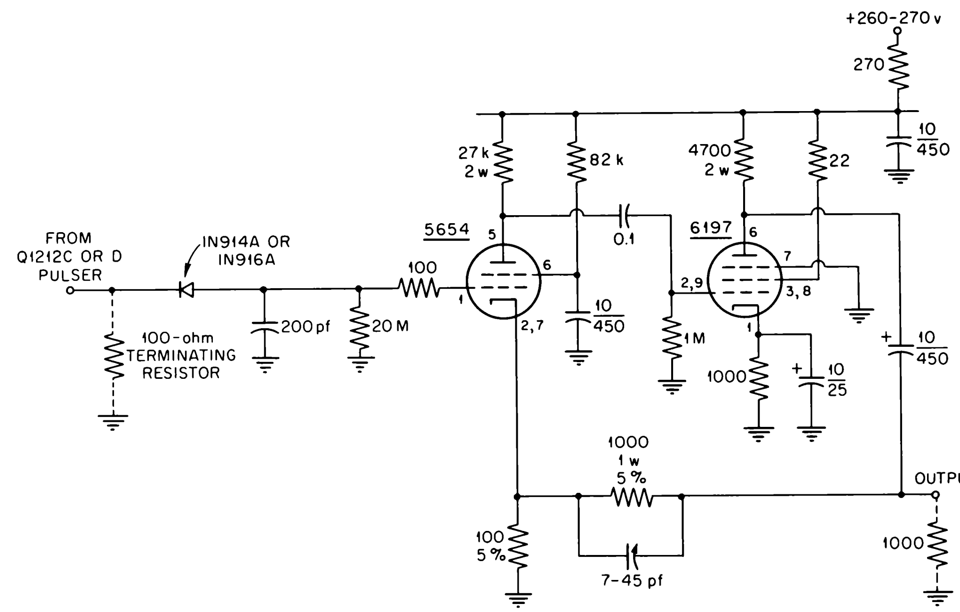

Fig. 1.19. Power Amplifier for Use with a Q-1212C or D Pulser (Gain = 10).

1.7 MEASUREMENT OF POWER-SUPPLY REGULATION

Usually, amplifier power supplies are sufficiently well regulated that a 300-v meter connected between B^+ and ground will show little or no variation when the line voltage changes from 105 to 125 v. To measure the variation, it is necessary to buck out most of the voltage so that one of the low-voltage ranges on the meter may be used. The bucking voltage can be the power supply from another amplifier, batteries, or any other convenient source whose output is independent of the power supply being tested. A typical setup is shown in Fig. 1.20. There is one note of caution to be observed with the setup shown in Fig. 1.20: The voltmeter should be disconnected or switched to a high range before either power supply is switched.

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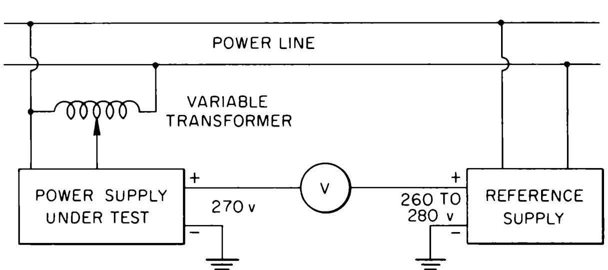


Fig. 1.20. Test Setup for Measuring Power Supply Regulation.

A typical set of measurements is shown below.

Line voltage	105	115	125
Variation, v	– 0.2	0	+ 0.3

The negative sign indicates that the power-supply output has decreased from its 115-v value.

In typical amplifier applications, regulation against variation in load is unimportant. However, it can be measured by the foregoing method if the load is varied (with the aid of loading resistors) instead of the line voltage.

1.8 MEASUREMENT OF AMPLIFIER GAIN SENSITIVITY TO LINE VOLTAGE

The gain variation of the amplifier section must be measured with a pulse-height discriminator which receives its power from a source external to the amplifier. The discriminator may be internal or external to the amplifier chassis.

The block diagram of the test setup is shown in Fig. 1.21. The test is performed as follows:

1. Set the line voltage to the amplifier at 115 v. Note that only the amplifier is connected to the variable line-voltage supply.
2. Set the pulser and discriminator to 500 and the rate meter to 60 cps, full scale. Adjust the amplifier gain and the NORMALIZE control on the pulser for 30 cps (corresponding to the half-triggering-point – htp).
3. Set the pulser to 250 and the discriminator to the htp. If its dial reads 250 ± 20, its zero adjustment is sufficiently close for the remainder of the test.
4. Set the pulser at 500 and the line voltage to the amplifier at 105 v. Permit the amplifier to stabilize for 2 min before taking any readings.
5. Set the discriminator for the htp and note the readings. The amplifier gain at 105 v relative to its gain at 115 v is D_2/D_1, where D_2 is the discriminator reading at 105 v and D_1 is the reading at 115 v.
6. Repeat at 125 v.

An example follows:

At 115 v, the discriminator reading is 500.0.

At 105 v, the reading is 498.

The gain at 105 v relative to 115 v is 498/500 = 0.996, which represents a drop of 0.4%.

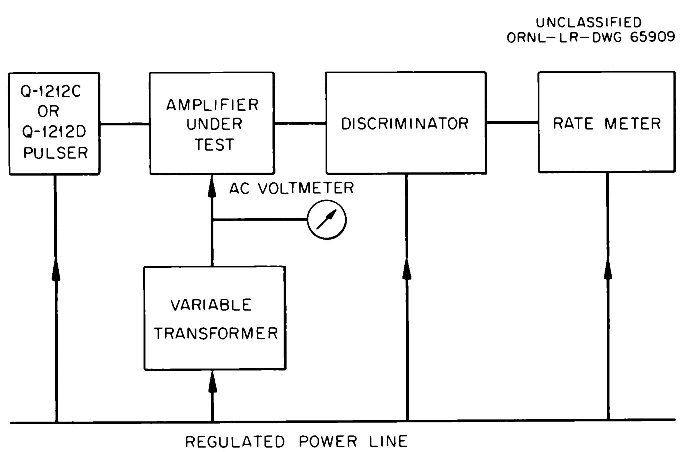


Fig. 1.21. Test Setup for Measuring Gain Sensitivity to Line Voltage.

2. TEST ROUTINES FOR SPECIFIC AMPLIFIERS

2. TEST ROUTINES FOR SPECIFIC AMPLIFIERS

2.1 A1D LINEAR AMPLIFIER, MODEL Q-1326A

2.1.1 General Description

The A1D linear amplifier (see block diagram in Fig. 2.1) is a singly differentiated, general-purpose linear pulse amplifier which can be used with a variety of detectors. It will accept either polarity signal. Clipping times are 0.16 μsec (RC), 1 μsec (delay line), 1.6 μsec (RC), or 16 μsec (RC). Voltage-sensitive preamplifiers are available having gains of 23 and 0.7.

Tolerance to high-duty-cycle operation and the ability to recover rapidly from large overloads were not among the design criteria for this amplifier. Duty cycle should be kept below 2% for best results. In the delay-line operating position, the recovery time (to 10% of rated output) for 10× overload is 25 μsec and for 100× overload is 9000 μsec.

The amplifier is supplied with a built-in integral discriminator.

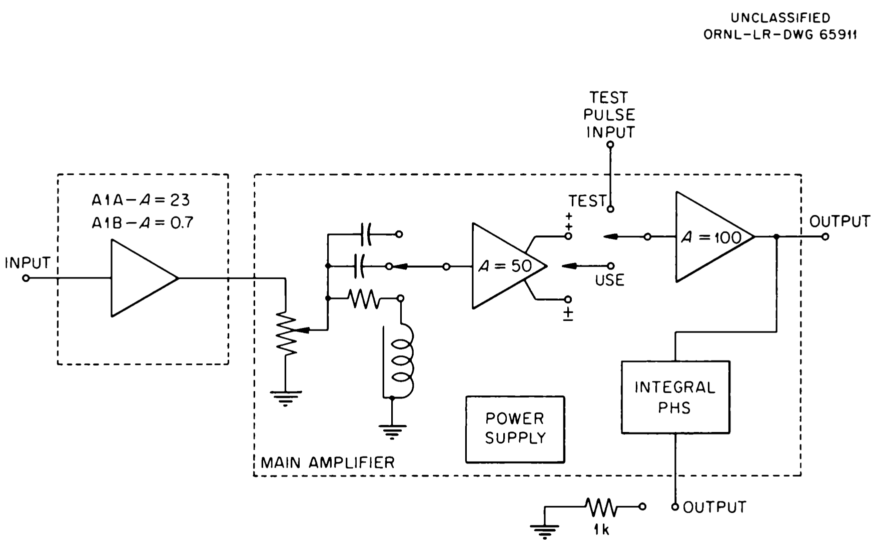


Fig. 2.1. A1D Linear Amplifier.

2.1.2 Specifications for the Main Amplifier

1. Pulse shaping: RC or delay line, unipolar pulse.
2. Pulse characteristics and gain:

Dial Setting	Pulse Width (μsec) Between 50% Points	Pulse Width (μsec) Between 10% Points	Rise Time 10%-90% (μsec)	Fall Time 90%-10% (μsec)	Undershoot (%)	Maximum Gain
Delay line	0.95	1.45	0.3	0.65	0.5	1,920
2-Mc bandwidth	0.4	0.75	0.16	0.45	0.3	2,700
0.5-Mc bandwidth	3.4	6.4	0.7	5.1	3	6,000
0.1-Mc bandwidth	20	200[a]	7.0	45	30	10,040

[a]Due to large undershoot.

3. Gain control –

Coarse:	Six steps by factors of 2, calibrated 2 to 64.
Fine:	0.5 to 1. Should be left at maximum when delay-line pulse shaping is used.

4. Signal polarities: ± input; ± output.
5. Saturated output: 100 v.
6. Integral nonlinearity: 0.1% from 0 to 90 v.
7. Permissible output load: 1000 ohms.
8. Overload performance – delay line only:

Overload Factor	Recovery Time (μsec)	
	To ± 10 v	To ± 5 v
1	1.5	1.5
10	25	35
100	9000	10,500

9. Gain change vs line voltage:

Line Voltage	105	115	125
ΔG	–1%	0	+1.2%

10. Line voltage at which power supply drops out of regulation: 90 v.
11. Power-supply ripple: 20 mv p–p.
12. Permissible power-supply drain: 120 ma at B^+; 13.9 amp at 6.3 v; 3 amp at 5 v.
13. Measured power-supply drain:

Circuit Group	At B^+ (ma)	At 6.3 v (amp)
Main amplifier	61	5.35
Discriminator	17	1.05
Preamplifier, A1A	22	0.6
Preamplifier, A1B	4	0.15

14. Power-line drain –

Without preamplifier:	1.08 amp.
With A1A preamplifier:	1.21 amp.

15. Front panel size: $8\frac{3}{4}$ by 19 in.

16. Connectors on front panel –

Amplifier output:	Low – UHF or BNC; high – UHF or BNC.
Test pulse input:	UHF or BNC.
PHS output:	UHF or BNC.

17. Connectors on rear apron –

Input:	UHF or BNC.
Preamplifier:	AN-3102A-16S-2S, seven-pin female.
Power:	Three-wire NEMA.

18. Tube complement: 1 – 6SJ7; 4 – 6AC7; 1 – 6197; 1 – 5687; 4 – 403B/5591; 1 – 6080; 1 – 5U4GB; 1 – 6AU6; 2 – 85A2.

2.1.3 Specifications for an A1A Preamplifier, Model Q-1326A-10-R_0

1. Circuit configuration: Voltage-sensitive feedback (triple) plus cathode-follower output.
2. Voltage gain: 23.
3. Feedback factor: 16.
4. Signal polarity: Noninverting.
5. Input capacitance: 9 pf.
6. Output impedance: Approximately 300 ohms.
7. Permissible cable length: 12 ft.
8. Rise time with 5 ft of cable: 0.1 μsec.
9. Saturated preamplifier
   Output voltage –
   - Single pulse, long cable: 0.7 v negative; 6 v positive.
   - Pileup waveform: 8 v negative; 20 v positive.
10. Noise performance with amplifier at full gain:

Detector Capacity (pf)	Delay Line	Bandwidth (Mc) 2	0.5	0.1
		Equivalent Electrons, rms		
10	2,000	2,380	1,125	910
33	4,250	4,700	2,340	1,440
100	10,000	11,250	5,620	2,900
		Equivalent Charge (coulombs, rms) $\times 10^{-17}$		
10	32	38	17.8	14.6
33	68	75	37.5	23
100	160	180	90	46.3

11. Nearest main amplifier setting at which main amplifier noise equals preamplifier noise (for detector capacities between 0 and 100 pf):

	Delay Line	Bandwidth (Mc)		
		2	0.5	0.1
Gain	16	16	8	8

12. Discriminator setting for 60-cps noise count (full scale = 1000):

Detector Capacity (pf)	Delay Line	Bandwidth (Mc)		
		2	0.5	0.1
10	26	65	80	85
33	25	57	74	70
100	25	52	70	60

13. Connectors –

Input:	UG-560/U (Amphenol 82-805).
Output:	AN-3102A-16-1S.
High voltage:	UG-931/U (IPC 27,000).

14. Tube complement: 4 – 403B/5591.

2.1.4 Specifications for an A1B Preamplifier

1. Circuit configuration:	Cathode follower.
2. Voltage gain:	0.7.
3. Signal polarity:	Noninverting.
4. Input capacitance:	9 pf.
5. Output impedance:	Approximately 300 ohms.
6. Permissible cable length:	12 ft.
7. Rise time with 5 ft of cable:	0.1 μsec.
8. Saturated preamplifer	
Output voltage –	
Single pulse, long cable:	0.7 v negative; 6 v positive.
Pileup waveform:	8 v negative; 20 v positive.

9. Noise performance with main amplifier gain at maximum:

Detector Capacity (pf)	Delay Line	Bandwidth (Mc)	
		0.5	0.1
	Equivalent Electrons, rms		
10	22,000	5,600	5,100
33		11,200	7,100
100	96,000	25,000	16,400
	Equivalent Charge (coulombs, rms)		
		$\times 10^{-15}$	
10	3.5	0.9	0.82
33		1.8	1.13
100	15.4	4	2.63

10. Connectors –
    - Input: UG-560/U (Amphenol 82-805).
    - Output: AN-3102A-16-1S.
    - High voltage: UG-931/U (IPC 27,000).
11. Tube complement: 1 – 403B/5591.

2.1.5 Specifications for the Discriminator

1. Type: Integral.
2. Resolving time: Duration of drive pulse plus 0.1 μsec.
3. Pulse-height range: 85 v.
4. Output pulse: 30-v negative pulse into 1000-ohm load; 0.5 μsec wide at the base.
5. Hysteresis: Less than 0.1 v.
6. Zero shift vs line voltage:

	Line Voltage (v)		
	105	115	125
Zero shift, minor dial divisions	–15	0	+5

7. Current drain: 17 ma at B^+; 1.05 amp at 6.3 v.
8. Tube complement: 4 – 403B/5591; 1 – 6AC7; 1 – 6AL5 (ordinarily not used).

2.1.6 Tube-Voltage Charts

Typical tube voltages for A1D amplifier tubes and for high-gain preamplifier tubes are given in Tables 2.1 and 2.2, respectively.

Table 2.1. Typical Tube Voltages for A1D Amplifier Tubes

The line voltage was 115 v ac; the unregulated B^+ was 400 v

Tube Number	Pin Number, Tube Element (e),* and Voltage (v)**																	
	1		2		3		4		5		6		7		8		9	
	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v
V1, 6SJ7	Sh	0	H	0	Su	0	G	0	C	1.5	Sc	115	H	0 6.45 ac	P	115	[a]	
V2, 6AC7	Sh	0	H	0	Su	0	G	0	C	1.7	Sc	115	H	0 6.45 ac	P	100		
V3, 6AC7	Sh	0	H	0	Su	0	G	0	C	4.1	Sc	186	H	0 6.45 ac	P	183		
V4, 6AC7	Sh	0	H	0	Su	0	G	0	C	1.34	Sc	93	H	0 6.45 ac	P	111		
V5, 6197	C	2.47	G	0	Sc	120	H	0 6.4 ac	H	0	P	144	Su	0	Sc	120	G	0
V6, 5687	P	256	G	0	C	15.2	H	0	H	0	C	15.2	G	0	H	0 6.4 ac	P	256
V7, 5U4GB			H[b]	425 5.0 ac			P	0 365 ac			P	0 365 ac			H[b]	425 5.0 ac		
V8, 6080	G	203	P	400	C	263	G	203	P	400	C	263	H[b]	258 6.25 ac	H[b]	258 6.25 ac		
V9, 6AU6	G	–[c]	Su	166	H[b]	257 6.25 ac	H[b]	257 6.25 ac	P	203	Sc	207	C	166				
V10, 85A2	P	166	C	84			C	84	P	166			C	84				
V11, 85A2	P	84	C	0			C	0	P	84			C	0				
V12, 6AL5			P	0 to 84	H	0	H	0 6.45 ac	C	–								
V13, 403B/5591	G	–	C	99	H	0	H	0 6.4 ac	P	262	Sc	263	C	99				
V14, 403B/5591	G	95	C	99	H	0	H	0 6.4 ac	P	263	Sc	263	C	99				
V15, 403B/5591	G	–	C	45	H	0	H	0 6.4 ac	P	120	Sc	144	C	45				
V16, 403B/5591	G	36	C	45	H	0	H	0 6.4 ac	P	144	Sc	144	C	45				
V17, 6AC7	Sh	0	H	0	Su	6.0	G	0	C	6.0	Sc	249	H	0 6.4 ac	P	253		

*The tube elements are indicated by letters: G, grid; C, cathode; H, heater; P, plate; Sc, screen grid; Su, suppressor grid; and Sh, shield.

**All voltages were measured with a 20,000-ohms/v multimeter. The voltages are positive with respect to ground and are dc voltages unless specified as ac.

[a] Blanks indicate unused pins.

[b] The ac voltage was measured between heater pins.

[c] Dashes indicate that impedance levels were too high for a meaningful multimeter reading.

Table 2.2. Typical Tube Voltages for High-Gain Preamplifier Tubes

Tube No.[a]	Pin Number, Tube Element,[b] and Voltage (v)[c]						
	1 G (v)	2 C (v)	3 H (v)	4 H (v)	5 P (v)	6 Sc (v)	7 C (v)
1	0	1.14	0	0 5.95 ac	78	77	1.14
2	0	1.95	0	0 5.95 ac	68	98	1.95
3	*d*	16	0	0 5.95 ac	98	98	16
4	16	16	0	0 5.95 ac	98	98	16

[a]All tubes are 403B/5591.

[b]The tube elements are indicated by letters: G, grid; C, cathode; H, heater; P, plate; and Sc, screen grid.

[c]All voltages were measured with a 20,000-ohms/v multimeter. The voltages are positive with respect to ground and are dc voltages unless specified as ac.

[d]The impedance level was too high for a meaningful multimeter reading.

2.1.7 Adjustments

The test setup shown in Fig. 2.2 should be connected. Neither the capacitor fitting (shown in Fig. 1.16) nor the rise-time spoiler are needed; the generator may be connected directly to the preamplifier. The 10× probe on the oscilloscope should be used.

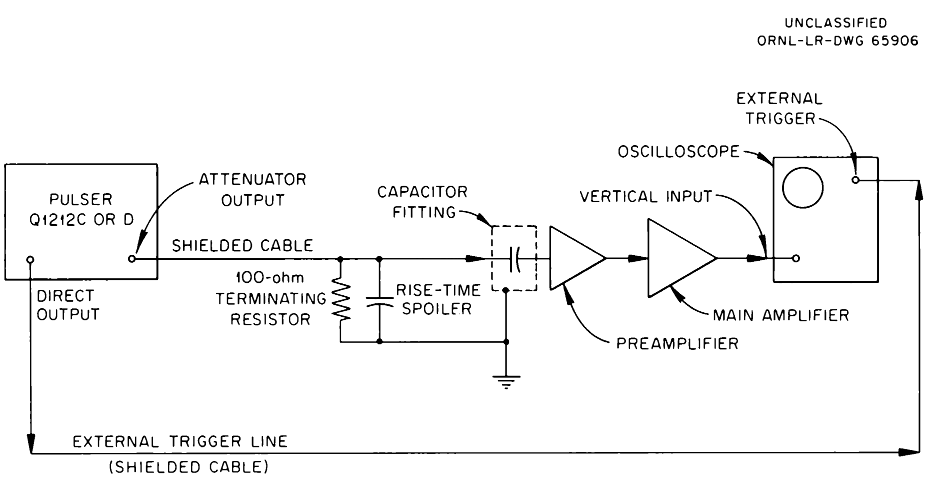


Fig. 2.2. Test Setup for Adjustment of an A1D Amplifier.

Transient Response of a High-Gain Preamplifier

The test instrument settings should be:

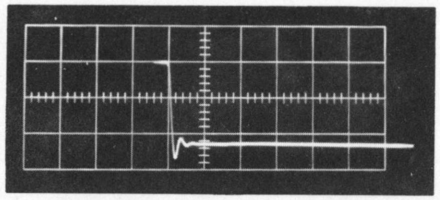
1. Generator: 0.05 v, negative.
2. Main amplifier bandwidth: Delay line.

3. Main amplifier gain –

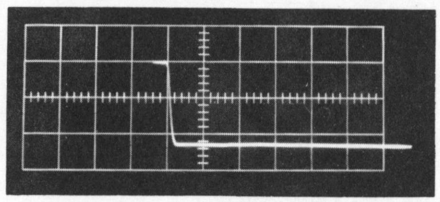
Coarse:	Minimum.
Fine:	Maximum.

4. Observation point: Arm of fine gain control (point B on R1).
5. Oscilloscope: 0.5 v/cm; 0.5 μsec/cm. (The 0.5 v/cm refers to the sensitivity as observed on the face of the tube. With a 10× probe the dial setting on the oscilloscope should be 0.05 v/cm.)

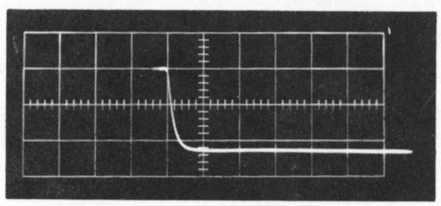
C3 should be adjusted to obtain optimum transient response as illustrated by Fig. 2.3.



(*a*) Trimmer capacity too low



(*b*) Trimmer capacity optimum



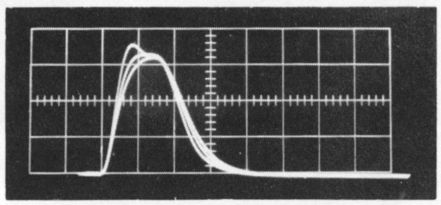
(*c*) Trimmer capacity too high

Fig. 2.3. Waveforms for Adjustment of C3.

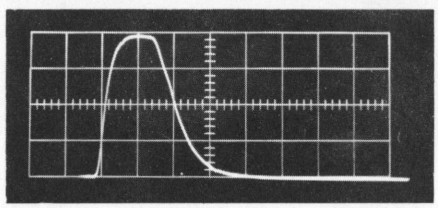
Transient Response of the Main Amplifier

The test instruments should be set as follows:

1. Generator:	0.05 v.
2. Bandwidth:	Delay line.
3. Main amplifier gain –	
Coarse:	Minimum.
Fine:	Maximum.
4. Observation point:	Main Amplifier output (high).
5. Oscilloscope:	20 v/cm; 0.5 μsec/cm.



(*a*) Composite oscillogram



(*b*) Middle trace of *a*

Fig. 2.4. Waveforms for Adjustment of C17. (*a*) Shows a composite oscillogram with three settings of C17. The middle trace represents the optimum.

Overload Performance

The instrument settings are:

1.	Generator:	As needed.
2.	Bandwidth:	Delay line.
3.	Main amplifier gain:	Maximum.
4.	Amplifier loading:	1000 ohms.
5.	Observation point:	Main amplifier output.
6.	Oscilloscope:	50 v/cm; 5 μsec/cm and 2 msec/cm, Fig. 2.5.
7.	Preamplifier:	High gain.
8.	Input time constant:	200 μsec.

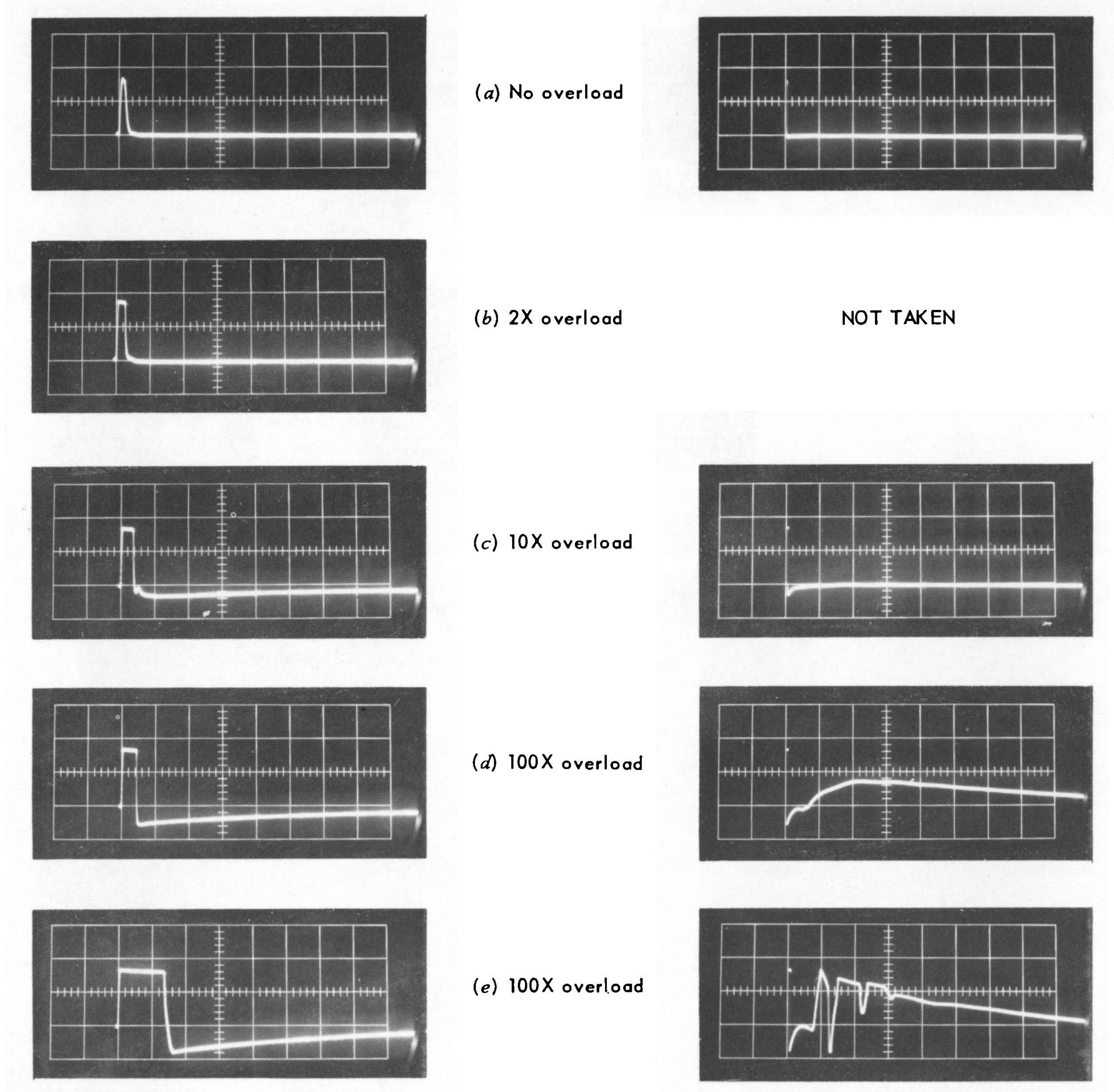


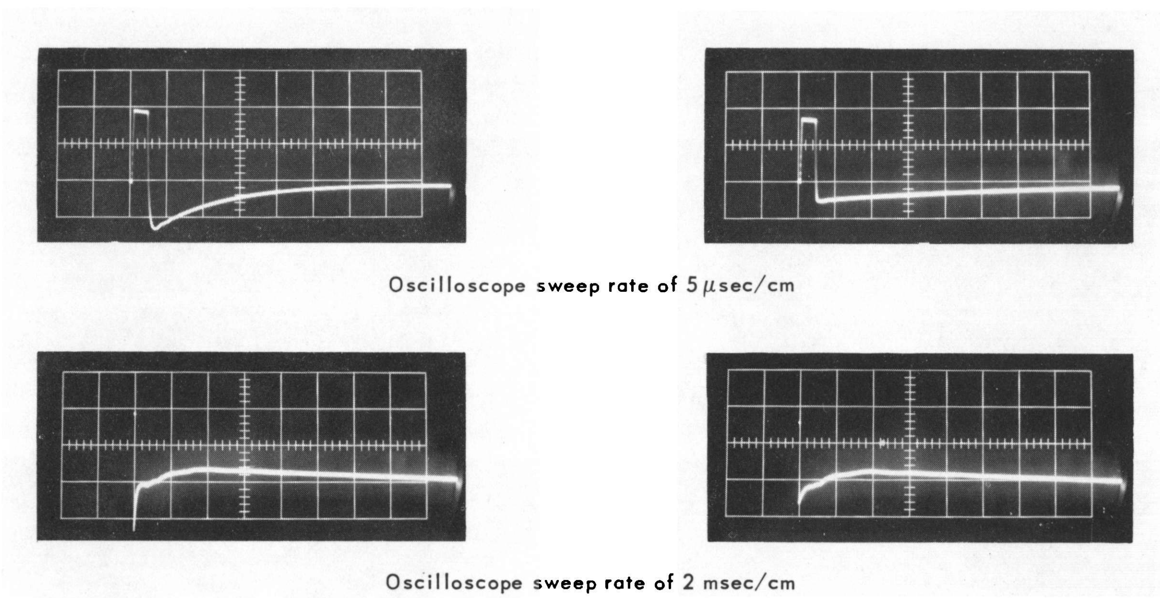
Fig. 2.5. Waveforms of Overload Performance. Waveforms in left and right columns were taken with oscilloscope sweep rates of 5 μsec/cm and 2 msec/cm, respectively.

Effect of Amplifier Load on Waveform (Fig. 2.6)

The principal effects of the 1000-ohm load resistor on the amplifier output signal are: (1) a reduction of approximately 10% in effective amplifier gain (due to the internal resistance of 100 ohms, as determined by R46); and (2) a reduction of more than half in the magnitude of the undershoot. Failure to use this resistor in normal operation will have little if any effect on the accuracy of spectral measurements.

The test-instrument settings are:

1. Generator	As needed.
2. Bandwidth:	Delay line.
3. Main amplifier gain:	Maximum.
4. Observation point:	Amplifier output.
5. Oscilloscope:	50 v/cm.
6. Preamplifier:	High gain.
7. Input time constant:	100 μsec.
8. Overload:	50×.



Oscilloscope sweep rate of 5 μsec/cm

Oscilloscope sweep rate of 2 msec/cm

(*a*) Without 1000-ohm load (*b*) With 1000-ohm load

Fig. 2.6. Effect of Amplifier Load on Waveform.

Effect of Input Time Constant on Amplifier Performance

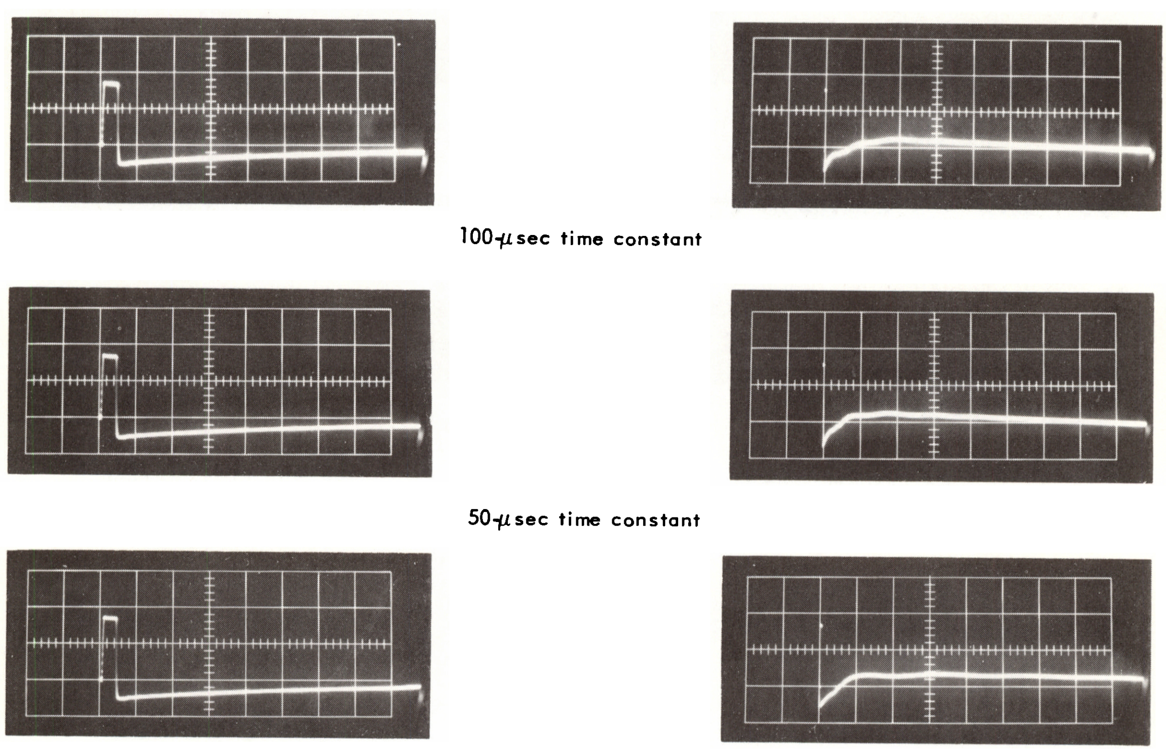
Because of pulse pileup, the permissible counting rate is inversely proportional to the square root of the input time constant. As low a time constant as is consistent with overload recovery time should be used.

To vary the time constant, a 100-pf capacitor was used to couple the signal into the preamplifier. C2 (390 pf) was removed from the input circuit, and various resistors were substituted for $R1$ such that the product 100 $R1$ was the desired time constant, where $R1$ is measured in megohms.

Changing the time constant has no perceptible effect on the undershoot when viewed on the 5-μsec/cm scale (Fig. 2.7). The effect is small, but perceptible, when viewed on the 2-msec/cm scale. A 50-μsec time constant is recommended when using a 1-μsec delay line.

The following are the test-instrument settings:

1. Pulse generator:	Set for 50× overload.
2. Main amplifier bandwidth:	Delay line.
3. Main amplifier gain:	Maximum.
4. Main amplifier loading:	1000 ohms.
5. Observation point:	Amplifier output.
6. Oscilloscope:	50 v/cm; 5 μsec/cm and 2 msec/cm (Fig. 2.7*b*).



100-μsec time constant

50-μsec time constant

20-μsec time constant

(*a*) Oscilloscope sweep rate of 5 μsec/cm

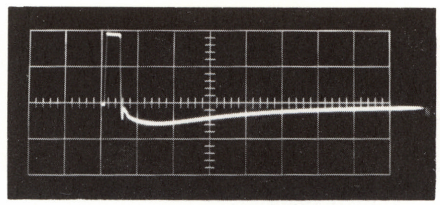
(*b*) Oscilloscope sweep rate of 2 msec/cm

Fig. 2.7. Waveforms Showing Effect of Input Time Constant on Amplifier Performance.

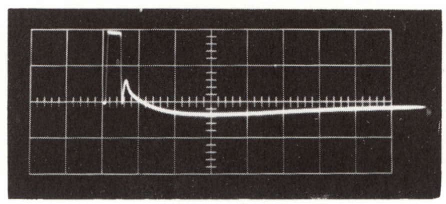
Effect of Fine-Gain Control on Delay-Line Termination (Fig. 2.8)

The instrument settings are:

1.	Pulse generator:	Set for 20× overload.
2.	Main amplifier bandwidth:	Delay line.
3.	Main amplifier loading:	Open circuit.
4.	Observation point:	Amplifier output.
5.	Oscilloscope:	50 v/cm; 5 μsec/cm.



(*a*) Coarse gain, 32; fine gain, 10



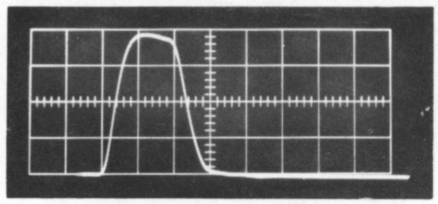
(*b*) Coarse gain, 64; fine gain, 0.5

Fig. 2.8. Waveforms Showing the Effect of Fine-Gain Control on Delay-Line Termination.

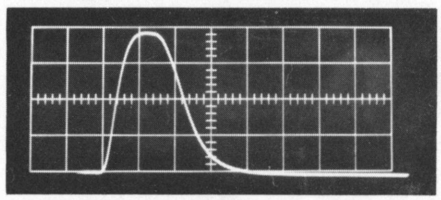
Effect of Compensating Termination 2 ft from End of Line (Figs. 2.9 and 10)

The settings on the test instruments are:

1. Pulse generator:	As needed.
2. Amplifier output:	80 v.
3. Amplifier gain –	
Coarse:	Minimum.
Fine:	Maximum.
4. Observation point:	Amplifier output.
5. Oscilloscope:	20 v/cm; 0.5 μsec/cm.

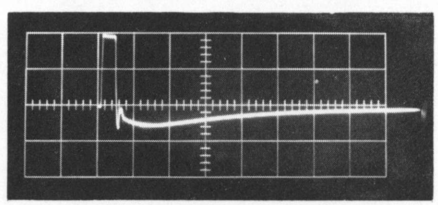


(*a*) Without termination

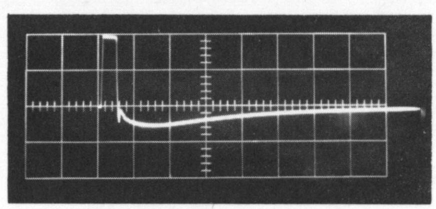


(*b*) With termination

Fig. 2.9. Effect of Termination.



(*a*) Without termination

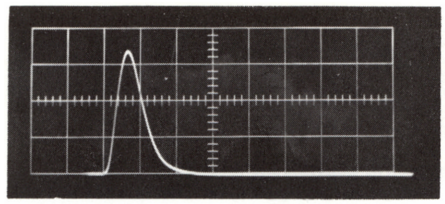


(*b*) With termination

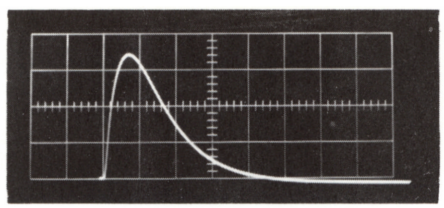
Fig. 2.10. Same Conditions as Shown in Fig. 2.9 Except That Generator Is Set for 20X Overload and Sweep Is Set for 5 μsec/cm.

Waveforms of RC Clipping Positions on Bandwidth Switch (Fig. 2.11)

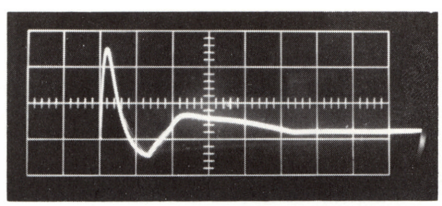
1. Main amplifier gain –
   Coarse: Minimum.
   Fine: Maximum.
2. Main amplifier loading: Open circuit.
3. Observation point: Amplifier output.
4. Oscilloscope: 20 v/cm.
5. Preamplifier: High gain.



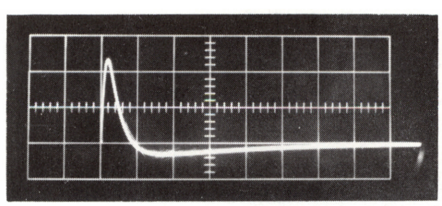
(*a*) 2-Mc bandwidth, RC clipping, oscilloscope 0.5 μsec/cm, pulser 25 mv



(*b*) 0.5-Mc bandwidth, oscilloscope 50 μsec/cm, pulser 4 mv



(*c*) 0.1-Mc bandwidth, oscilloscope 50 μsec/cm, pulser 4 mv



(*d*) Same as *c*, but with C34 removed

Fig. 2.11. Waveforms of RC Clipping Positions.

In Fig. 2.11, waveform (*d*), it is shown that removal of C34 is essential for correct operation in the 0.1-Mc bandwidth position.

Output Noise

1. Main amplifier gain –
   Coarse: 2.
   Fine: Maximum.
2. Observation point: Amplifier output.
3. Oscilloscope: 0.5 v/cm; 10 msec/cm.
4. Preamplifier: High gain.

The waveform in Fig. 2.12 indicates that both 60- to 120-cps components are present and that they have an amplitude at least as great as the rms value of the random noise. The 120-cps component is independent of the bandwidth.

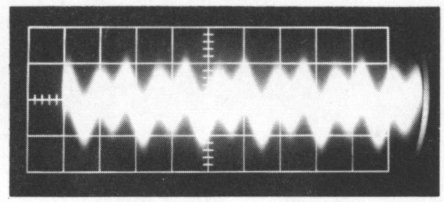
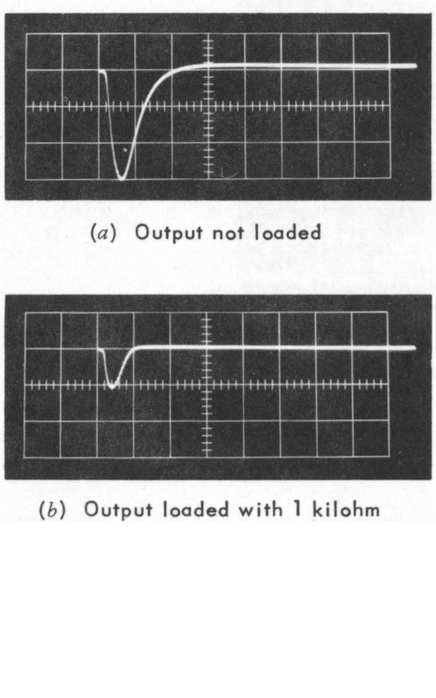


Fig. 2.12. Waveform Showing Effect of Output Noise.

Discriminator Output (Fig. 2.13)

1. Pulse generator: As needed to trigger the discriminator.
2. Amplifier gain: Minimum.
3. Observation point: PHS output.
4. Oscilloscope: 50 v/cm; 0.5 μsec/cm.



(*a*) Output not loaded

(*b*) Output loaded with 1 kilohm

Fig. 2.13. Discriminator Output Waveforms.

2.1.8 Measurement of Feedback Factors

The test instruments should be set as follows:

1. Pulse generator:	As needed to produce 20 to 80 v output.
2. Main amplifier gain:	Minimum.
3. Observation point:	Amplifier output.
4. Termination:	Open circuit.
5. Bandwidth:	Delay line.
6. Oscilloscope:	5 to 20 v/cm; 0.5 μsec/cm.
7. Preamplifier:	High gain.

Note the generator dial reading and the amplifier output. Shunt a 25-μf capacitor across $R3$ in the preamplifier. Readjust the generator to produce the original output voltage. The ratio of dial readings is the feedback factor; a ratio of 16 is typical.

Ground pin 5 of tube V1 in the main amplifier through the 25-μf capacitor. Reset the generator dial for the original output voltage and determine the new ratio. A typical value is 8.

Move the capacitor to pin 5 of tube V4 and repeat. The feedback factor should be about 16.

2.1.9 Linearity

Use the setup shown in Fig. 2.2, but feed the generator signal directly into the main amplifier. Follow the procedure in Sec 1.3.6, "Measurement of Integral Nonlinearity, Method I." Nonlinearity should be no greater than 1 minor dial division between 0 and 1000.

2.1.10 Regulation of Power Supply

Follow the procedure described in Sec 1.7, "Measurement of Power Supply Regulation." Regulation should be ±0.7 v or less in the range of 105 to 125 v.

2.1.11 Gain Dependence on Line Voltage

Follow the procedure described in Sec 1.8, "Measurement of Amplifier Gain Sensitivity to Line Voltage." Gain variation should be less than 3% in the range from 105 to 125 v.

2.1.12 Noise Measurements

Noise measurements may be made by following the procedure described in Sec 1.5.4, "Measuring Amplifier Noise with a VTVM." See the list of specifications for typical noise values.

2.1.13 Circuit Diagrams

The circuit diagram for the A1D main amplifier is shown in Fig. 2.14. The circuit diagrams for the A1A preamplifier and the A1B preamplifier (cathode follower) are shown in Figs. 2.15 and 2.16, respectively.

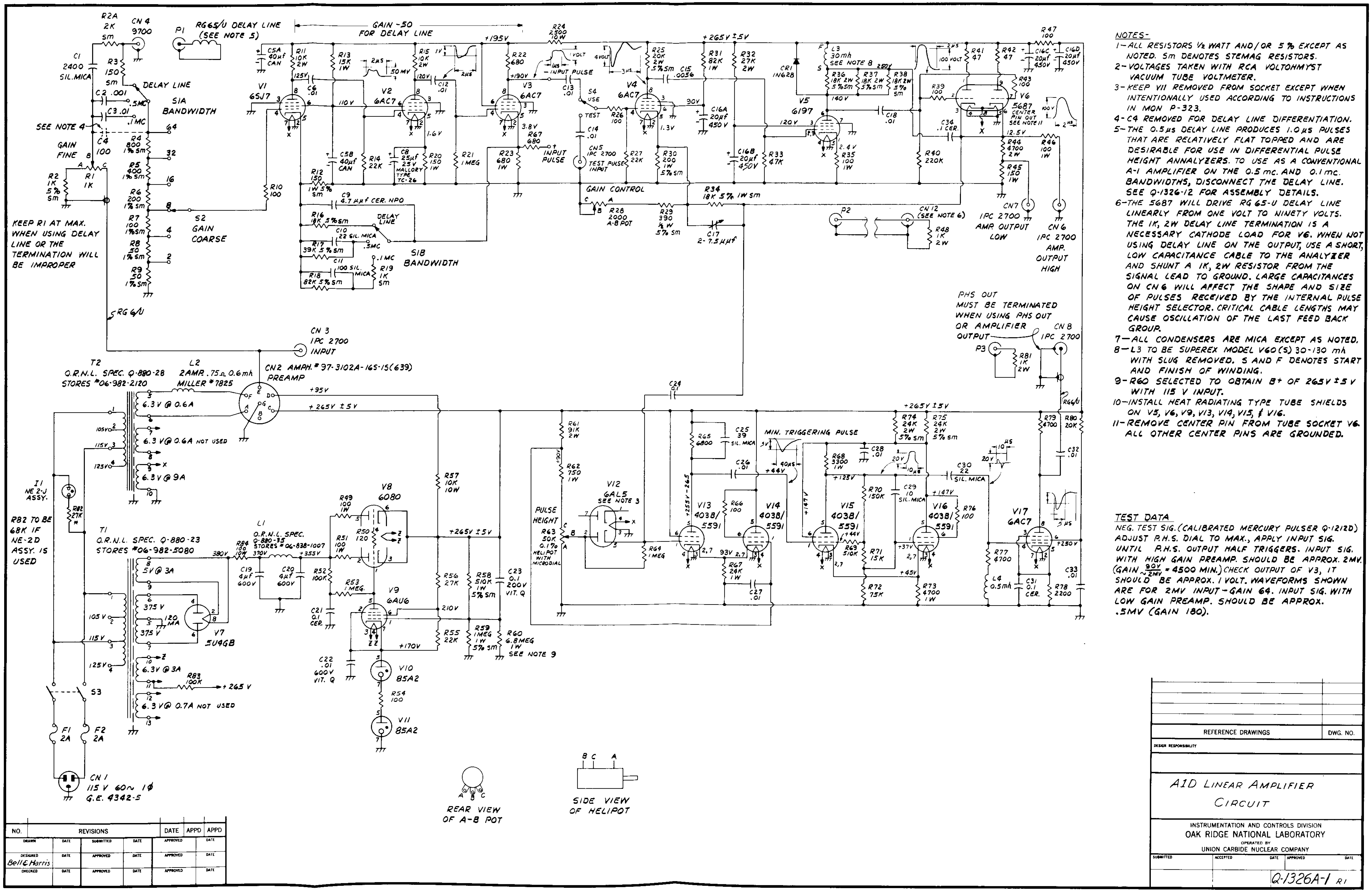


Fig. 2.14. A1D Linear Amplifier Circuit.

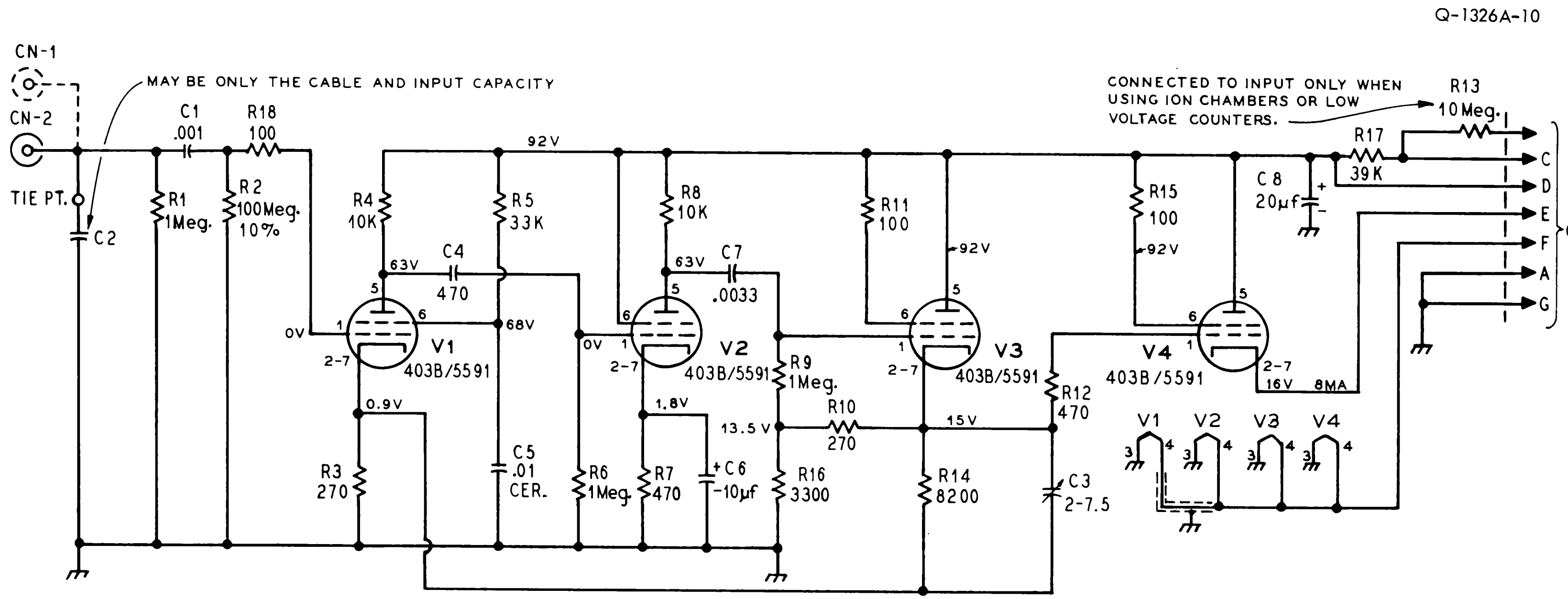


Fig. 2.15. A1D Preamplifier Circuit.

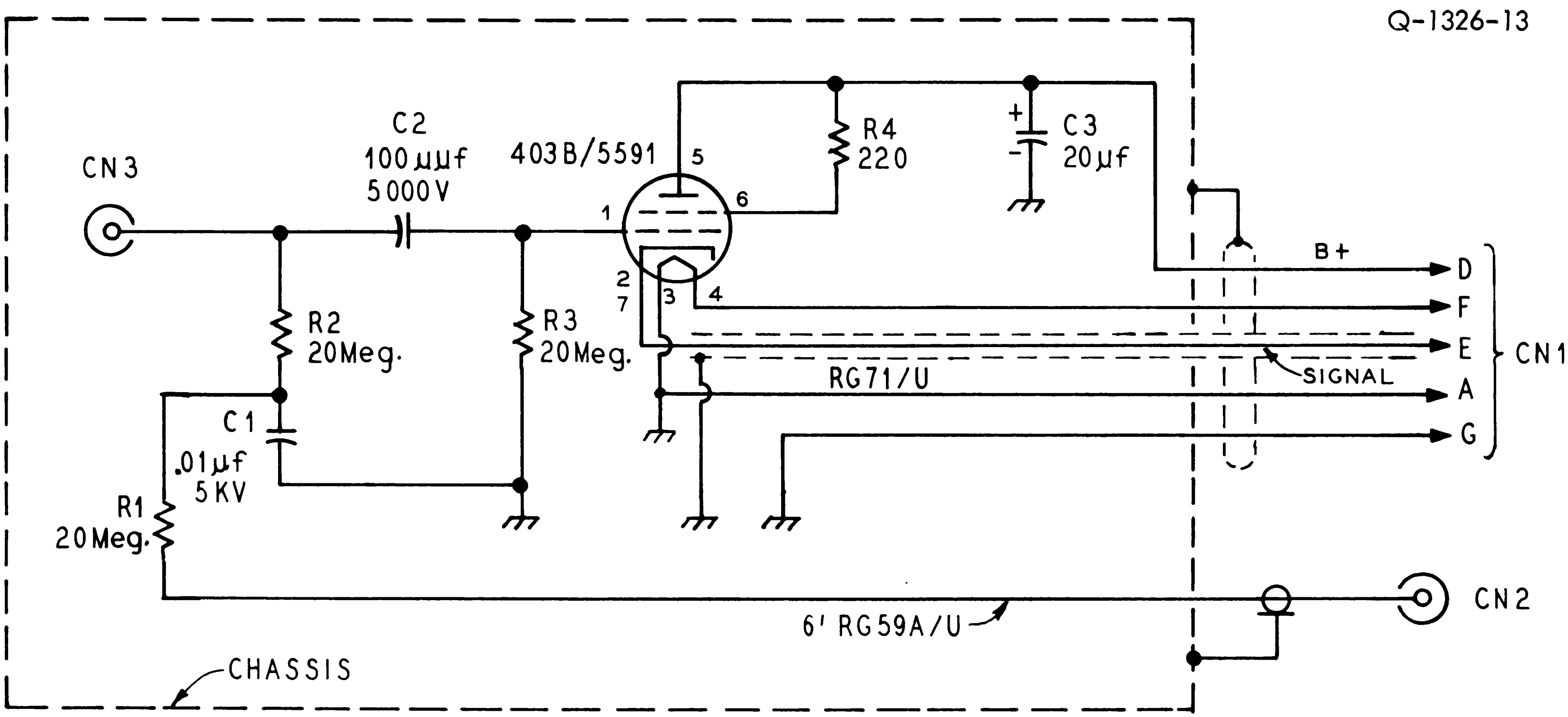


Fig. 2.16. A1D Amplifier: Cathode-Follower Circuit Details and Assembly.

2.2 DD2 LINEAR AMPLIFIER, MODEL Q-1593

2.2.1 General Description

The DD2 linear amplifier (see block diagram in Fig. 2.17) is a double-delay-line, nonblocking amplifier designed for use with scintillation counters. Its particular attributes are its ability to recover rapidly from large overloads (recovery to 10% of rated output in 25 μsec from a 400× overload) and its tolerance of very high counting rates (100,000 counts/sec). It may be used with other radiation detectors if their output signal is compatible with a fixed 1.2-μsec clipping time, a voltage gain range of 50 to 50,000, and a best rms noise level equivalent to 5750 electrons with 10-pf external capacitance.

The amplifier has a built-in differential and integral discriminator.

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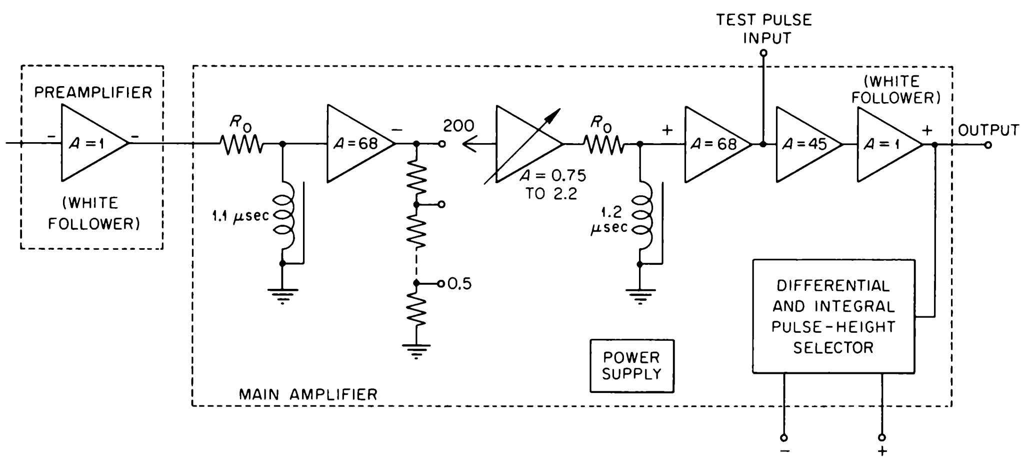


Fig. 2.17. DD2 Linear Amplifier.

2.2.2 Specifications for the Main Amplifier

1. Pulse shaping:	Double delay-line, bipolar pulse.
2. Pulse width –	
Between half-height points, primary pulse:	1.2 μsec.
Between 10% points, total pulse:	2.9 μsec.
3. Rise time from 10% to 90%:	Depends on output pulse height – at 25 v rise time is 0.16 μsec, at 100 v rise time is 0.4 μsec.
4. Fall time from 90% to 10%:	0.1 μsec.
5. Signal polarities –	
Input:	Negative.
Output:	Positive.
6. Gain:	50,000.
7. Gain control –	
Coarse:	0.5 to 200 in a 1-2-5 sequence (relative values).
Fine:	0.4 to 1.0.

8. Saturated output: 115 v.
9. Integral nonlinearity: 0.1% from 10 to 100 v; 0.2% from 0 to 100 v.
10. Loading: 8000 ohms minimum resistance with standard circuit; 4000 ohms with modified circuit.
11. Overload performance:

Overload Factor	Recovery Time (μsec)	
	To $\pm$ 10 v	To $\pm$ 5 v
1	2.9	3.0
10	5	5.5
100	7	18
400	25	35

12. Gain change vs line voltage:

Line Voltage	105	115	125
ΔG	−1.2%	0	+0.8%

13. Line voltage at which power supply drops out of regulation: 94 v.
14. Power-supply ripple: 50 mv p-p.
15. Permissible power-supply drain: 200 ma at B^+; 14.2 amp at 6.3 v.
16. Measured power-supply drain:

Circuit Group	At B^+ (ma)	At 6.3 v (amp)
Main amplifier	166	5.3
Pulse-height selector	34	2.6
Preamplifier	10	0.45

17. Power-line drain –
    - Without preamplifier: 1.56 amp at 115 v.
    - With preamplifier: 1.62 amp at 115 v.
18. Front panel size: $8\frac{3}{4}$ by 19 in.
19. Connectors on front panel –
    - Input: UG-290/U (BNC).
    - Test input: UG-290/U (BNC).
    - Output: UG-290/U (BNC).
    - PHS output: Positive (BNC); negative (BNC).
20. Connectors on rear apron –
    - Power: 3-pin NEMA.
    - External PHS control: 3-pin, male, AN-3102A-10S1-3P.
    - Preamplifier: 7-pin, female, AN-3102A-16S-1S.
21. Tube complement: 8 – 5965; 10 – 403B/5591; 1 – 6AK6; 2 – 6AN5; 1 – 6AU6; 3 – 6197; 1 – 6080; 1 – OA2; 2 – 5651.

2.2.3 Specifications for the Preamplifier

1. Circuit configuration:	White cathode follower.
2. Gain:	0.95.
3. Signal polarities:	Negative input; negative output.
4. Output impedance:	Approximately 100 ohms.
5. Permissible cable length:	May be any length if the heater voltage is $\geq$5.5 v and the cable is correctly terminated at the preamplifier end.
6. Saturated preamplifier output voltage –	
Single pulse, long 93-ohm cable:	8 v.
Pileup signal:	30 v.

7. Noise performance (with the main amplifier at maximum gain):

Detector Capacity (pf)	Noise (equivalent electrons, rms)
10	5,750
33	10,200
100	23,300

8. Nearest main amplifier gain setting at which main amplifier and preamplifier noise contributions are equal:	10.

9. Discriminator setting for 60-cps noise count:

Detector Capacity (pf)	Setting (minor dial divisions)
10	106
33	84
100	76

10. Connectors –	
Input:	UG-290/U (BNC).
Output:	5-pin male, AN-3102A-16S-1S.
11. Tube complement:	1 – 5965.

2.2.4 Specifications for the Discriminator

1. Type:	Differential or integral.
2. Resolving time:	Duration of drive pulse plus 1.2 μsec.
3. Pulse-height range:	5 to 100 v.
4. Slit-width range:	0 to approximately 10 v.
5. Output pulse:	30 v, 1 μsec positive; 30 v, 1 μsec negative (available simultaneously).
6. Hysteresis:	30 minor dial divisions of PHS control.

7. Zero shift vs line voltage:

	Line Voltage (v)		
	105	115	125
Zero shift, minor dial divisions	–4.0	0	+4

8. Slit-width stability: ±10 mv per day when the instrument is operated from a regulated power line in a temperature-controlled (±5°F) environment.

9. Current drain: 34 ma maximum at B^+; 2.6 amp at 6.3 v.

10. Tube complement: 1 – 6197; 4 – 403B/5591; 3 – 5965.

2.2.5 Tube-Voltage Chart

Typical tube voltages for DD2 amplifier tubes are given in Table 2.3.

Table 2.3. Typical Tube Voltages for DD_2 Amplifier Tubes

The line voltage was 115 v ac, unregulated B^+ was 380 v, unregulated B^- was –350 v, and the ripple voltage of the regulated B^+ was 0.2 v p-p. The dc regulation from 95 to 125 v $\stackrel{\sim}{=}$ 1 v

Tube No.	Pin Number, Tube Element (e),* and Voltage (v)**																	
	1		2		3		4		5		6		7		8		9	
	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v
V0, 5965	P	229	G	–[a]	C	135	H[b]	112 6.0 ac	H[b]	112 6.0 ac	P	135	G	30	C	33	H[b]	112 6.0 ac
V1, 5965	P	125	G	0	C	1.63	H[b]	0 5.9 ac	H[b]	0 5.9 ac	P	125	G	0	C	1.63	H[b]	0 5.9 ac
V2, 403B/5591	G	–1.65	C	0	H[c]	0 5.9 ac	H[c]	0 5.9 ac	P	115	Sc	115	C	0	[d]			
V3, 403B/5591	G	–1.65	C	0	H[c]	0 5.9 ac	H[c]	0 5.9 ac	P	145	Sc	105	C	0				
V4, 403B/5591	G	–	C	115	H[c]	115 6.1 ac	H[c]	115 6.1 ac	P	217	Sc	225	C	115				
V5, 5965	P	120	G	1.7	C	1.78	H[b]	0 5.9 ac	H[b]	0 5.9 ac	P	120	G	0	C	1.78	H[b]	0 5.9 ac
V6, 403B/5591	G	–1.65	C	0	H[c]	0 5.9 ac	H[c]	0 5.9 ac	P	120	Sc	125	C	0				
V7, 403B/5591	G	–1.65	C	0	H[c]	0 5.9 ac	H[c]	0 5.9 ac	P	120	Sc	120	C	0				
V8, 6AN5	G	–9.4	Su	2.0	H[c]	0 5.9 ac	H[c]	0 5.9 ac	P	200	Sc	110	C	2.0				
V9, 5965	P	110	G	–	C	–10	H[b]	0 5.9 ac	H[b]	0 5.9 ac	P	135	G	0	C	0	H[b]	0 5.9 ac
V10, 403B/5591	G	–1.65	C	0	H[c]	0 5.9 ac	H[c]	0 5.9 ac	P	105	Sc	113	C	0				
V11, 6AN5	G	–10	Su	1.6	H[c]	0 5.9 ac	H[c]	0 5.9 ac	P	123	Sc	125	C	1.6				

Table 2.3 (continued)

Tube No.	Pin Number, Tube Element (e),* and Voltage (v)**																	
	1		2		3		4		5		6		7		8		9	
	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v
V12, 5965	P	123	G	123	C	245	H^b	165 6.1 ac	H^b	165 6.1 ac	P	245	G	200	C	215	H^b	165 6.1 ac
V13, 6AK6	G	123	Su	133	H^c	165 6.1 ac	H^c	165 6.1 ac	P	380	Sc	245	C	133				
V14, 6197	C	8.6	G	0	Sc	245	H^e	0 5.9 ac	H^e	0 5.9 ac	P	220	Su	8.6	Sc	245	G	0
V15, 6197	C	7.4	G	0	Sc	245	H^e	0 5.9 ac	H^e	0 5.9 ac	P	245	Su	7.4	Sc	245	G	0
V16, 6080	G	215	P	380	C	265	G	225	P	380	C	265	H^f	165 6.0 ac	H^f	165 6.0 ac		
V17, 6AU6	G	165	Su	170	H^c	165 6.0 ac	H^c	165 6.0 ac	P	215	Sc	265	C	170				
V18, 5651	P	170	C	85			C	85	P	170			C	85				
V19, OA2	P	85	C	0			C	0	P	85			C	0				
V20, OA2	P	0	C	− 145			C	− 145	P	0			C	0				
V201, 6197	C	20 to 120	G	–	Sc	264	H^e	85 6.1 ac	H^e	85 6.1 ac	P	264	Su	20 to 120	Sc	264	G	–
V202, 403B/5591	G	20 to 120	C	120 to 130	H^c	85 6.1 ac	H^c	85 6.1 ac	P	264	Sc	264	C	120 to 130				
V203, 403B/5591	G	117 to 126	C	120 to 130	H^c	85 6.1 ac	H^c	85 6.1 ac	P	249	Sc	264	C	120 to 130				
V204, 5965	P	264	G	− 17	C	0	H^b	0 5.9 ac	H^b	0 5.9 ac	P	264	G	− 17	C	0 Diff 45 Int	H^b	0 5.9 ac
V205, 5965	P	264	G	–	C	100	H^b	85 6.1 ac	H^b	85 6.1 ac	P	214	G	90	C	100	H^b	85 6.1 ac
V206, 5965	P	264	G	− 17	C	0	H^b	0 5.9 ac	H^b	0 5.9 ac	P	264	G	− 17	C	0	H^b	0 5.9 ac
V207, 403 B/5591	G	20 to 120	C	120	H^c	85 6.1 ac	H^c	85 6.1 ac	P	264	Sc	264	C	120				
V208, 403B/5591	G	117	C	120	H^c	85 6.1 ac	H^c	85 6.1 ac	P	249	Sc	264	C	120				

*The tube elements are indicated by letters: G, grid; C, cathode; H, heater, P, plate; Sc, screen grid; Su, suppressor grid; and Sh, shield.

**All voltages were measured with a 20,000-ohms/v multimeter. The voltages are positive with respect to ground and are dc voltages unless specified as ac. A negative sign(−) indicates polarity with respect to ground.

[a] Dashes indicate that impedance levels were too high for a meaningful multimeter reading.

[b] The ac heater voltage was measured between heater pins 4–5 and 9.

[c] The ac heater voltage was measured between pins 3 and 4.

[d] Blanks indicate unused pins.

[e] The ac heater voltage was measured between pins 4 and 5.

[f] The ac heater voltage was measured between pins 7 and 8.

2.2.6 Test Points

A legend of the circuit locations for the test points and the design voltages for connected test points on the front panel are given in Tables 2.4 and 2.5, respectively.

Table 2.4. Legend for Test Points

Test Point	Circuit Location
A	Regulated B^+
B	Ground
1	Unregulated B^+
2	B^-
3	Stage 1
4	Stage 2
5	Stage 3
6	Stage 4
7	Limiter

Table 2.5. Design Voltages for Connected Test Points

Connected Test Points		Design Value	Permissible Deviation
+	−	(v)	(± v)
A	B	270	10
1	A	105	35
B	2	145	5
A	3	20	5
A	4	20	5
A	5	20	5
A	6	20	5
3	7	125	5

2.2.7 Adjustments

Connect the test setup shown in Fig. 2.18. Trigger the oscilloscope sweep from the pulser. Use a 10× oscilloscope probe for all measurements. Keep the amplifier pulse-shape control at minimum (fully counterclockwise) unless instructed otherwise.

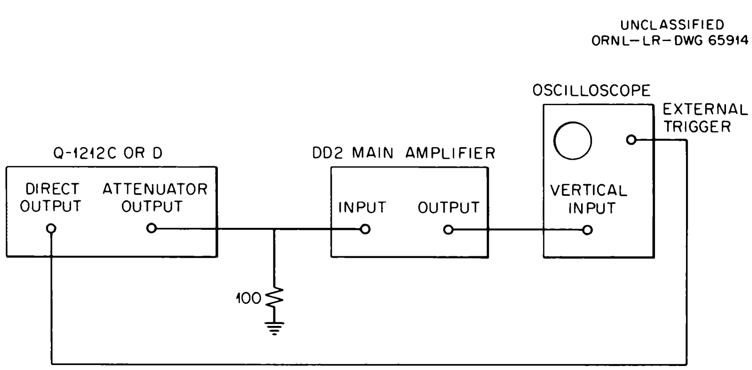


Fig. 2.18. Test Setup for Adjustment of a DD2 Amplifier.

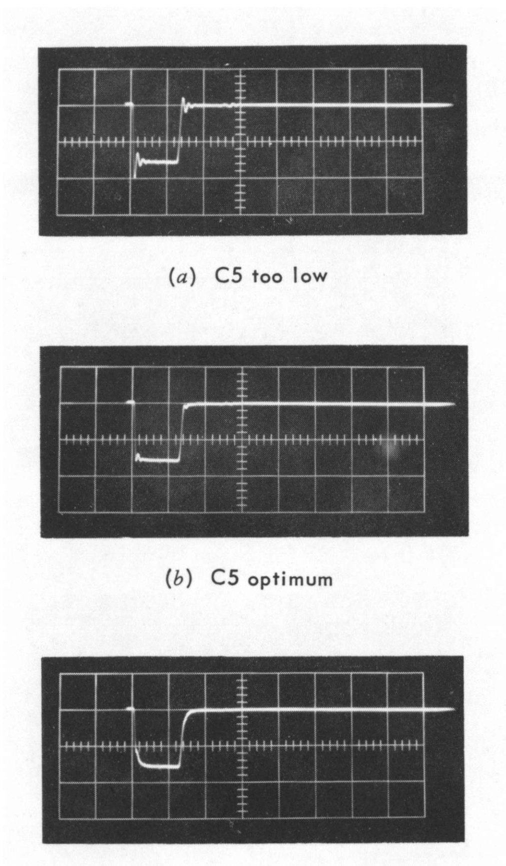
Transient Response of Stage 1

With the instruments set as follows, set C5 for optimum transient response (Fig. 2.19):

1. Generator: 0.5-v output, negative.
2. Main amplifier gain: 0.
3. Observation point: Pin 5 (plate) of V3.
4. Oscilloscope: 10 v/cm; 1 μsec/cm. (The 10 v/cm refers to absolute sensitivity, including probe loss.)

Effect of L1

L1 affects the height and polarity of the pip that occurs at 2.6–2.7 μsec from the start of the pulse (Fig. 2.20). This small pip becomes important under overload conditions. With C4 set for optimum and with the same settings as in the previous section, "Transient Response of Stage 1," the effect of L1 is shown by the waveforms in Fig. 2.20.

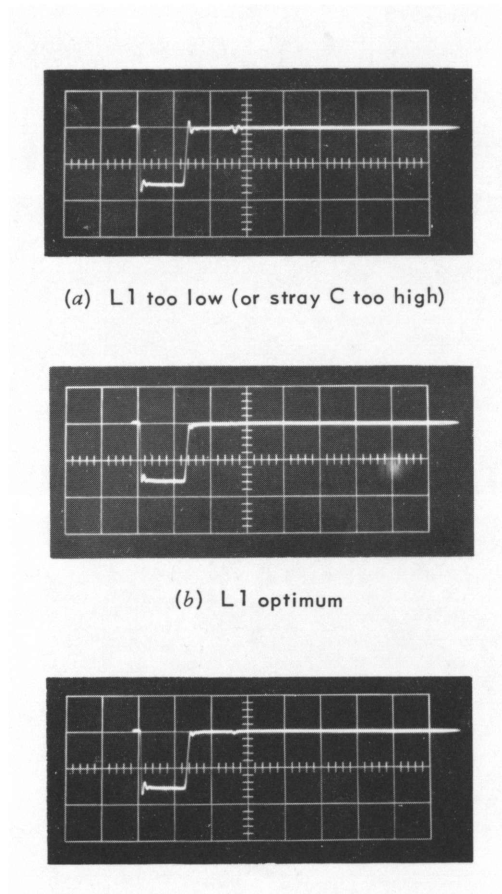


(*a*) C5 too low

(*b*) C5 optimum

(*c*) C5 too high

Fig. 2.19. Effect of C5 on Transient Response of Stage 1.



(*a*) L1 too low (or stray C too high)

(*b*) L1 optimum

(*c*) L1 too high

Fig. 2.20. Effect of L1 on Transient Response of Stage 1.

Effect of R7A

R7A compensates the effect of the delay-line resistance. Figure 2.21 shows the pulse with R7A open-circuited; the pulse fails to recover to the base line (compare this with Fig. 2.20).

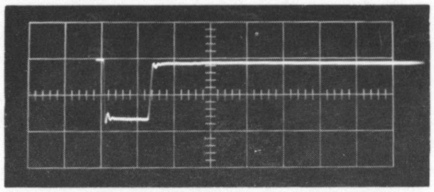
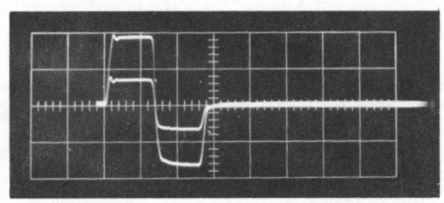


Fig. 2.21. Effect when R7A is Open Circuited.

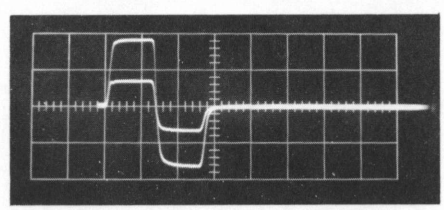
Transient Response of Stage 2

With the following instrument settings, C14 is set for optimum transient response (Fig. 2.22):

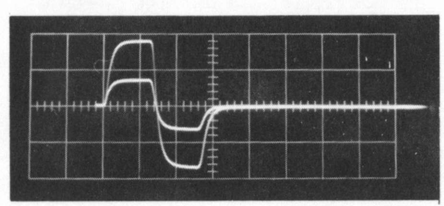
1. Generator:	0.5 v, negative.
2. Main amplifier gain –	
Coarse:	5.
Fine:	See Fig. 2.22.
3. Observation point:	Pin 5 of V7 (plate).
4. Oscilloscope:	10 v/cm; 1 μsec/cm.



(*a*) C14 too low; large trace when fine gain is set at maximum; small trace when set at minimum



(*b*) C14 optimum

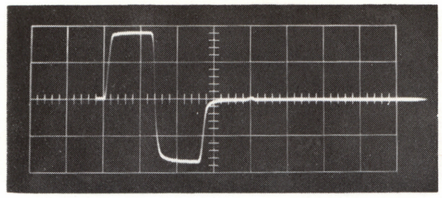


(*c*) C14 too high

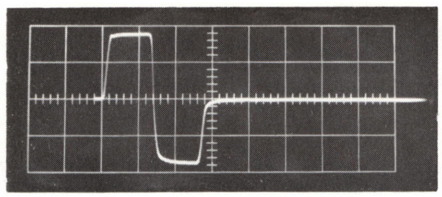
Fig. 2.22. Effect of C14 on Transient Response of Stage 2.

Adjustment of L3

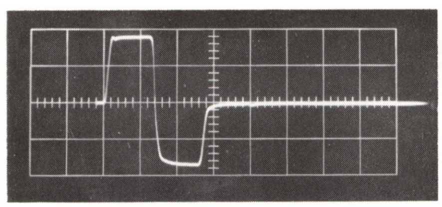
L3 affects the height and polarity of the small pip which occurs 4 μsec from the start of the pulse (Fig. 2.23). This pip becomes important under overload conditions.



(*a*) L3 too low (or stray C too high)



(*b*) L3 optimum



(*c*) L3 too high

Fig. 2.23. Effect of L3.

Final Adjustment of Leading-Edge Characteristics

(*a*) Set the instruments as follows:

1. Generator:	10 mv.
2. Main amplifier gain –	
Coarse:	5.
Fine:	1.0.
3. Observation point:	Output terminal.
4. Oscilloscope:	10 v/cm; 1 μsec/cm.

(*b*) Set C4 and C14 for minimum capacity (Fig. 2.24).

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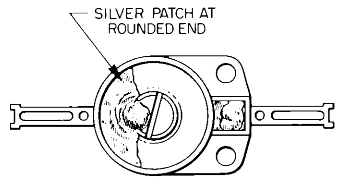


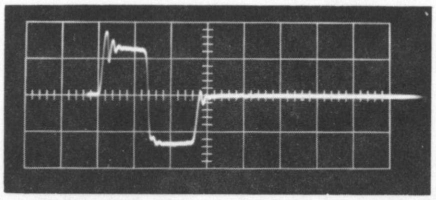
Fig. 2.24. C4 and C14 Set for Minimum Capacity.

(*c*) Set C20 for optimum transient response (Fig. 2.25).

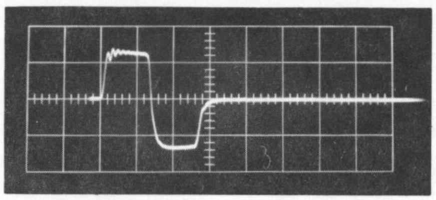
(*d*) After C20 has been set, increase C4 until leading edge is affected slightly. Repeat with C14. The final waveform is shown in Fig. 2.26.

(*e*) The waveforms at 25, 50, 75, 100, and 125 v output are shown in Fig. 2.27.

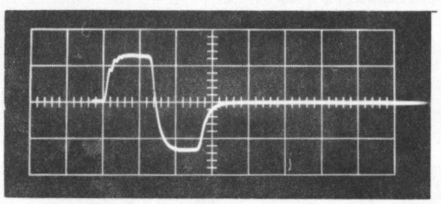
When the generator output terminal is loaded with 0.005 μf (see Sec 1.1.2) to simulate the output of an NaI(Tl) crystal, the waveforms have the appearance shown in Fig. 2.28.



(*a*) C20 set too low



(*b*) C20 optimum



(*c*) C20 set too high

Fig. 2.25. Effect of C20.

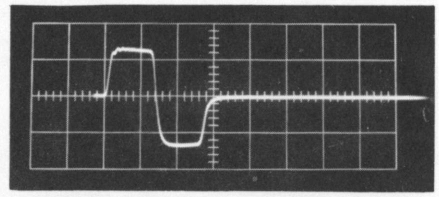


Fig. 2.26. Final Waveform After Adjustment of C14.

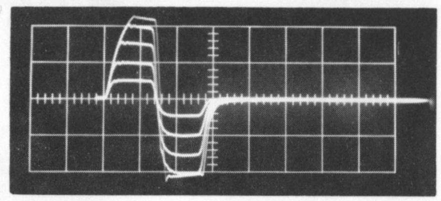


Fig. 2.27. Waveforms at Various Output Voltages.

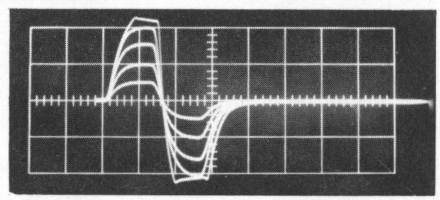


Fig. 2.28. Waveforms when the Output of a NaI(Tl) Crystal Is Simulated.

Pulse-Shape Control

The pulse-shape control affects the shape of the top of the pulse. If the amplifier is used principally with the internal discriminator, the ideal waveform exhibits a 2.5 to 3.5% drop at the top. If the amplifier is used principally with a multichannel analyzer, the ideal waveform rises continuously to the end of the pulse. The latter condition is obtained with the pulse-shape control set fully counterclockwise.

To obtain the required droop for differential discriminator use, use the same test setup as described in the preceding section, "Final Adjustment of Leading-Edge Characteristics," and adjust as follows:

1. Generator:	As needed; use 0.005-μf rise-time spoiler to simulate the output of an NaI(Tl) crystal.
2. Main amplifier gain –	
Coarse:	5 or 10.
Fine:	1.0.
3. Observation point:	Amplifier output terminal.
4. Oscilloscope:	1 μsec/cm; 10 v/cm; depress baseline as needed.

Waveforms at three output levels are shown in Fig. 2.29. The droop should be adjusted for 2.5 to 3 v at 90-v output.

The effect of the rise-time spoiler is shown in Fig. 2.30. The oscilloscope sensitivity is 50 v/cm.

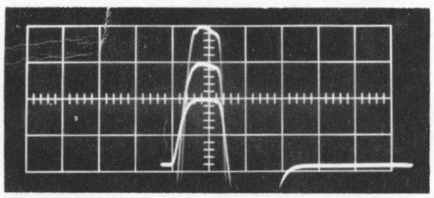


Fig. 2.29. Waveforms at Three Output Levels. Top trace, 90 v; middle trace, 45 v; bottom trace, 18 v output.

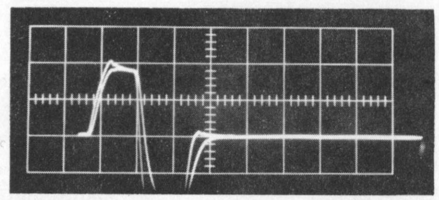


Fig. 2.30. Effect of Rise-Time Spoiler. Upper trace, without rise-time spoiler; lower trace, with rise-time spoiler.

Final Adjustment of Delay-Line Termination

(*a*) Connect the test setup shown in Fig. 2.31. The rise-time spoiler should be 0.005 μf. The capacitor fitting (see Fig. 1.16) may be eliminated, and the generator output signal may be fed directly into the preamplifier.

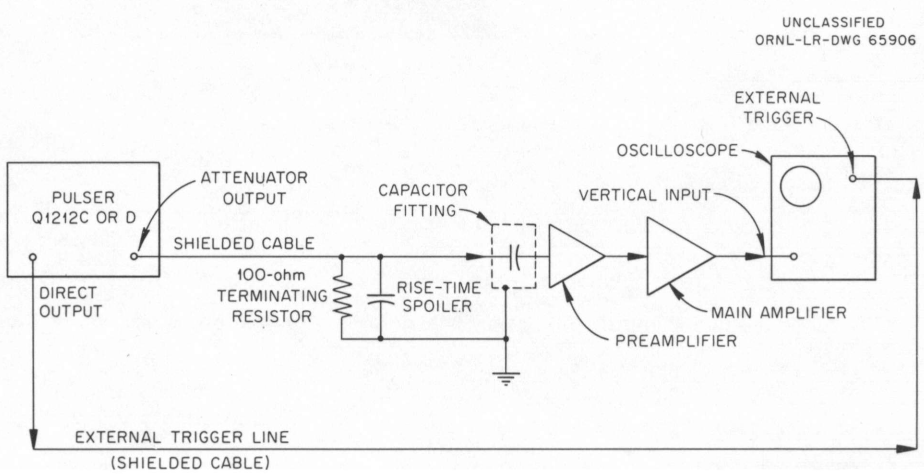
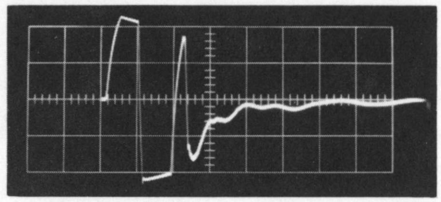


Fig. 2.31. Test Setup for Final Adjustment of Delay-Line Terminator.

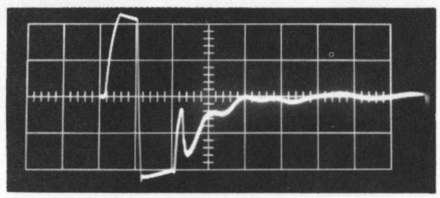
(*b*) Set the test instruments as follows:

1. Generator: As needed for 100× overload.
2. Amplifier gain –
   - Coarse: 50.
   - Fine: 1.0.
3. Oscilloscope: 50 v/cm; 2 μsec/cm.

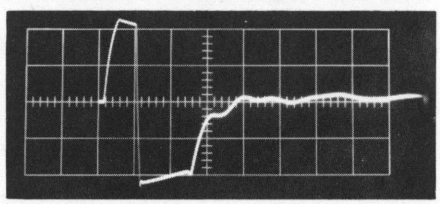
(*c*) Adjust the delay-line termination (front panel) for an incipient secondary overshoot (Fig. 2.32).



(*a*) R2 set too low



(*b*) R2 optimum



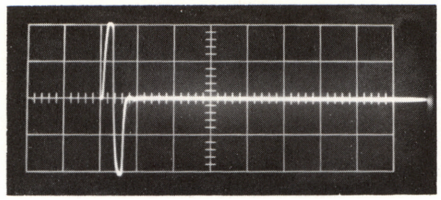
(*c*) R2 set too high

Fig. 2.32. Adjustment of Delay-Line Termination.

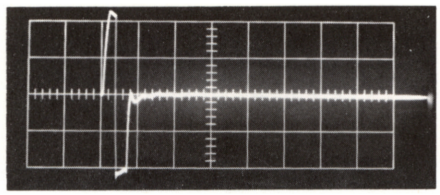
2.2.8 Overload Waveforms

The overload performance is tested with a constant input voltage of approximately negative (–) 1 v (which is the rated maximum input voltage). The degree of overload is varied by the amplifier gain control. A pre-amplifier and rise-time spoiler should be used.

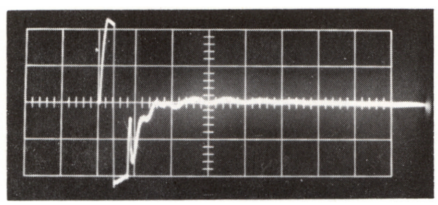
1. Generator: Approximately 1 v.
2. Amplifier gain –
   - Coarse: See Fig. 2.33.
   - Fine: 10.
3. Observation point: Amplifier output.
4. Oscilloscope: 50 v/cm; 5 μsec/cm.



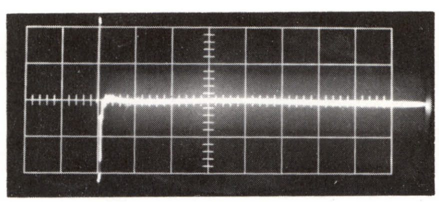
(a) 1X overload; gain = 0.5



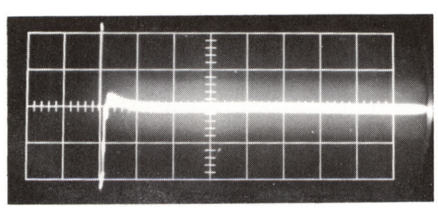
(b) 10X overload; gain = 5



(c) 400X overload; gain = 200



(d) 100X overload; gain = 50; sweep = 50 μsec/cm



(e) 400X overload; gain = 200; sweep = 50 μsec/cm

Fig. 2.33. Waveforms Showing Overload Performance.

2.2.9 Effect of Input Time Constant

The detector load resistor in conjunction with the input capacity (with the detector connected) determines the input time constant. Too long a time constant gives poor tolerance to pileup, while too short a time constant can result in multiple counts in a system having a resolving time of 1 or 2 μsec. Assuming the amplifier to have a clipping time of approximately 1 μsec, the usable dynamic range (the ratio of maximum to minimum input pulse height before multiple counts occur) will approximate T^2, where T is the input time constant in use; for example, with an input time constant of 10 μsec, the dynamic range will be roughly 100:1.

The effect of time constant may be studied with a circuit as shown in Fig. 2.34. The instrument settings are:

1. Generator:	Adjust for 100× overload.
2. Amplifier gain:	50.
3. Oscilloscope:	50 v/cm; 10 μsec/cm.

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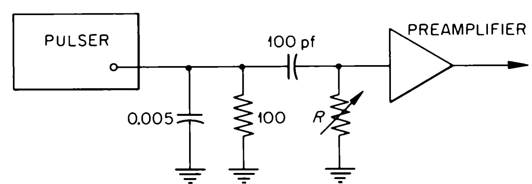
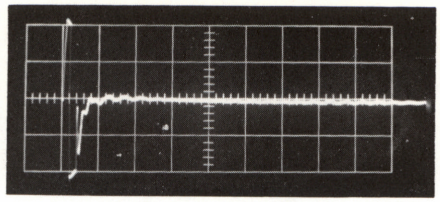


Fig. 2.34. Test Instrument Setup for Studying the Effect of Input Circuit Time Constant.

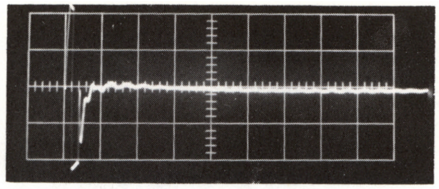
Waveforms resulting from use of the following time constants and detector load values are shown in Fig. 2.35.

Time Constant (μsec)	Load Resistance (ohms)
100	1 M
50	500 k
20	200 k
10	100 k
5	50 k

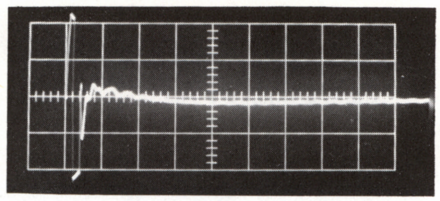
Note that with a 10-μsec time constant, the secondary pulse is 75 v high, resulting in a dynamic range of 75 v/100 v × 100× overload = 75, in approximate agreement with the preceding discussion.



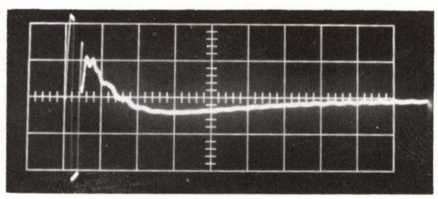
(*a*) 100 μsec



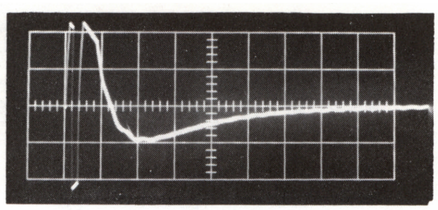
(*b*) 50 μsec



(*c*) 20 μsec



(*d*) 10 μsec



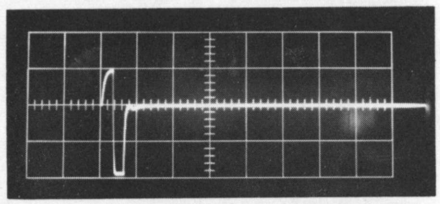
(*e*) 5 μsec

Fig. 2.35. Waveforms Showing the Effect of Time Constants.

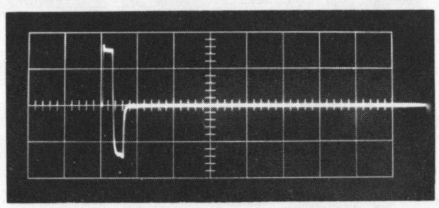
2.2.10 Overload Waveforms at Various Places in the Amplifier

With the instruments set as follows, the overload waveforms at various places in the amplifier are shown in Fig. 2.36.

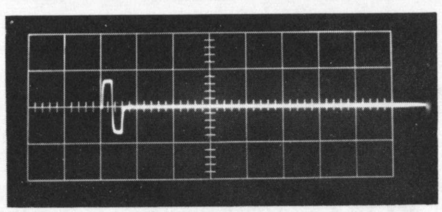
1. Generator: $\sim$1 v.
2. Amplifier gain: Maximum.
3. Oscilloscope: 5 μsec/cm.
4. Overload: 400$\times$.
5. Rise-time spoiler: None.



(*a*) At the input of the third stage, pin 1 of V8, with the oscilloscope set at 50 v/cm



(b) At the input of the second stage, junction of L3 and R36, with the oscilloscope set at 2 v/cm



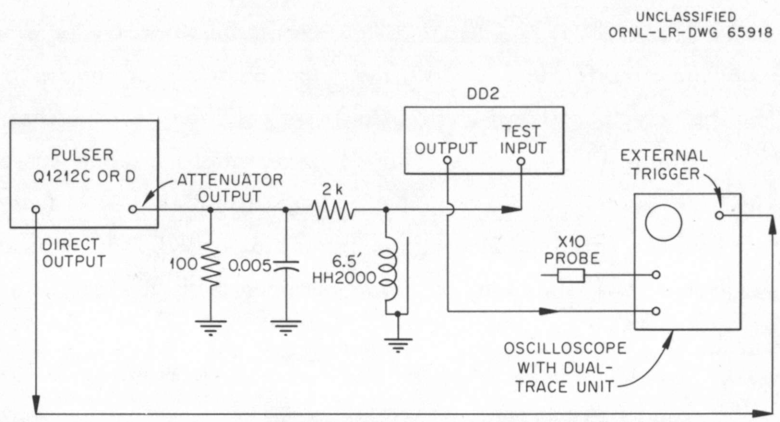
(*c*) Same as *b* but with the amplifier gain set at 20

Fig. 2.36. Waveforms Showing Overload at Various Places in the Amplifier.

2.2.11 Differential Discriminator Waveforms

Connect the circuit of Fig. 2.37. The amplifier test input terminal is used to eliminate noise modulation caused by the input stage.

Fig. 2.37. Circuit for Studying Differential Discriminator Waveforms.



Effect of Pulse-Shape Control on Width of Pulse from the Schmitt Triggers

For a Schmitt trigger to perform accurately its task of pulse-height discrimination, driving pulses that barely cross the trigger threshold should produce full-height output pulses from the Schmitt.[1] This is best accomplished when the driving pulse has a droop somewhat less than the trigger hysteresis. To observe the effect of driving pulse shape, set the controls as follows:

1. Pulse-shape control: 2.5-v droop at 95-v output.
2. Oscilloscope: 0.5 μsec/cm.
3. Oscilloscope –
   - To observe amplifier waveform: 5 v/cm (suppressed baseline).
   - To observe Schmitt-trigger waveform: 10 v/cm.
4. Observation point: Output of lower buffer, pin 3 of V204.

Adjust the PHS dial for the barely triggering condition. The resulting waveforms are shown in Fig. 2.38.

Turn the pulse-shape control counterclockwise until no droop is obtained, and readjust the PHS dial until the Schmitt is barely triggering. This waveform is shown in Fig. 2.39. Note that the Schmitt trigger starts into its quasi-stable state at the peak of the amplifier pulse and that recovery starts when the amplifier pulse has dropped to the hysteresis level.

---

[1] E. Fairstein, *Rev. Sci. Instr.* 27, 483 (1956).

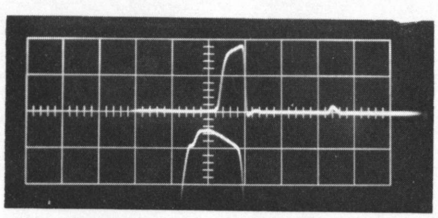


Fig. 2.38. Waveform Obtained When the Schmitt is Barely Triggering. The waveform is the amplifier pulse.

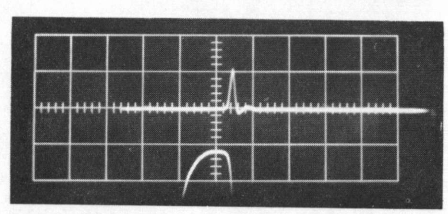


Fig. 2.39. Waveform Obtained with no Droop in the Pulse Shape and the Schmitt Barely Triggering.

In Fig. 2.38, it is seen that the Schmitt output pulse is wide enough to permit reliable operation of following circuits. In Fig. 2.39, the output pulse is very narrow and of partial height.

The waveform shown in Fig. 2.40 was obtained with the instruments set the same as when the waveform shown in Fig. 2.38 was obtained, except that the probe was connected to the plate (pin 6) of V208. The poor rise time is due to probe loading. The precursor is feed-through due to the grid-to-cathode capacity of V207. This is hidden by the cutoff bias of V208 (disturbances up to 12 v are biased off).

Figure 2.41 shows a waveform obtained with the same instrument settings as for Fig. 2.39 but modified as for Fig. 2.40.

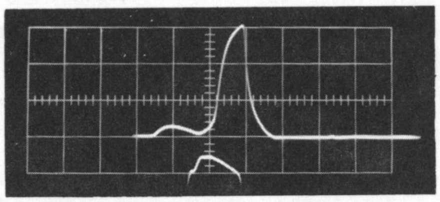


Fig. 2.40. Waveform Taken with the Probe Connected to the Plate of V208 and with the Instrument Settings as for Fig. 2.38.

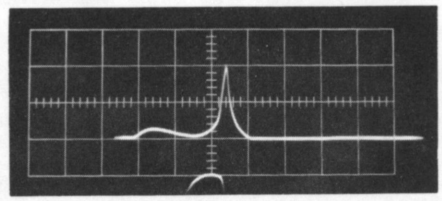


Fig. 2.41. Waveform Obtained with the Probe Connected to the Plate of V208 and with the Instrument Settings as for Fig. 2.39.

Effect of Amplifier Pulse Height on Schmitt Trigger Pulse Width

The instrument settings listed in the preceding section are used here, but the probe is returned to pin 3 of V204. In Fig. 2.42, the lowest amplifier trace corresponds to an amplitude of 94.5 v, which is insufficient to cause triggering. The trace above the lowest has an amplitude of 95 v and barely triggers the Schmitt circuit (the Schmitt output waveform is the narrowest of the upper three waveforms in Fig. 2.42). Successive pulse heights are 96 and 98 v, and the corresponding waveforms are successively wider.

In Fig. 2.42, the width of the Schmitt pulse (at the baseline) is 0.1 μsec less than the time spent by the amplifier pulse between the 95-v level on the rising edge of the pulse and the 91-v level on the falling edge of the pulse. The difference between these levels is the trigger hysteresis. (The apparent pulse width is less than the time spent between the pulse-height levels because of the time taken to traverse the cutoff region of the buffer stage.)

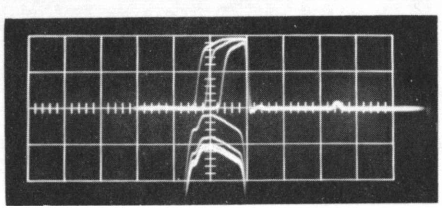


Fig. 2.42. Waveform Obtained with the Probe Connected to Pin 3 of V204.

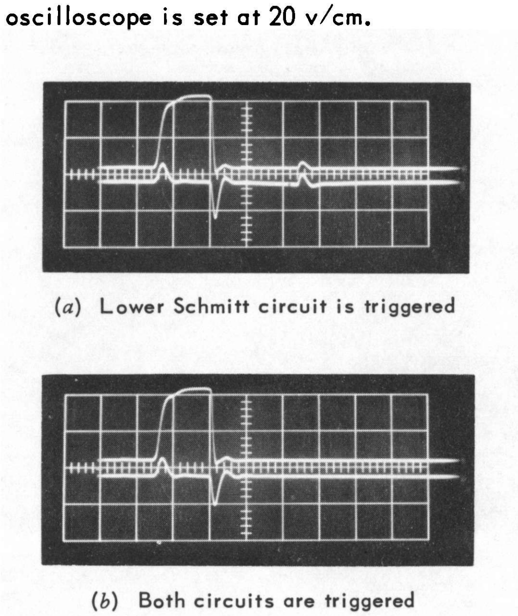
Miscellaneous Discriminator Waveforms

Unless stated otherwise, the top trace of each of the following waveforms is the output of the lower Schmitt trigger circuit as seen at pin 3 of V204. This trace is the timing reference. Calibration is 10 v/cm and 0.5 μsec/cm.

The lower trace in Fig. 2.43*a* and *b* is the interrogating signal[2] at the junction of L301 and R344. In Fig. 2.43*a*, only the lower Schmitt circuit is triggered, while in Fig. 2.43*b* both Schmitt circuits are triggered. The pip that occurs 2 μsec after the start of the pulse in Fig. 2.43*a* is due to feedback from the triggered anticoincidence tube, V205.

The corresponding waveforms at the interrogated grid of the anticoincidence trigger, pin 2 of V205, are shown in Fig. 2.44*a* and *b*. As before, only the lower Schmitt circuit is triggered in Fig. 2.44*a*, and both circuits are triggered in Fig. 2.44*b*.

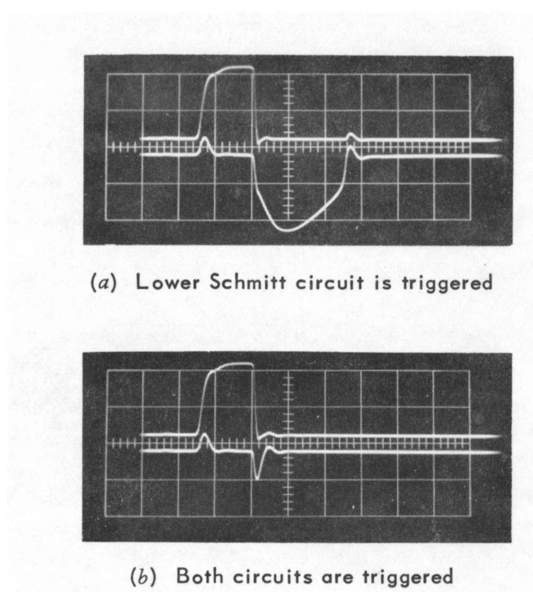
The waveform at the negative output terminal of the differential discriminator is shown in Fig. 2.45. The oscilloscope is set at 20 v/cm.



(*a*) Lower Schmitt circuit is triggered

(*b*) Both circuits are triggered

Fig. 2.43. Waveforms of the Interrogating Signal at the Junction of L301 and R344. The oscilloscope is set at 10 v/cm.



(*a*) Lower Schmitt circuit is triggered

(*b*) Both circuits are triggered

Fig. 2.44. Waveforms at the Interrogated Grid of the Anticoincidence Trigger, Pin 2 of V205. The oscilloscope is set at 20 v/cm.

In Fig. 2.46, the waveform that generates the inhibit signal is shown (this is *not* the inhibit signal). The inflections at the half-height points are produced by L302 and C19 and are important to the operation of the discriminator. The oscilloscope is set at 20 v/cm.

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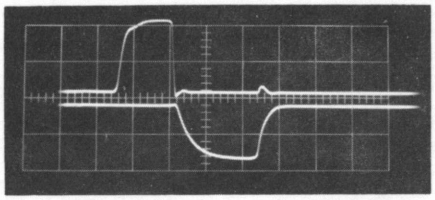
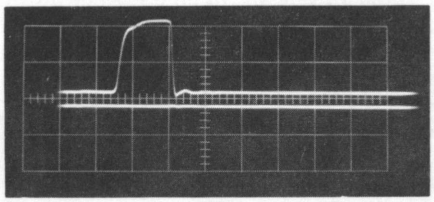
[2]For the theory of operation of this discriminator see E. Fairstein and F. M. Porter, "A Fast Differential and Integral Pulse Height Selector Circuit," *Rev. Sci. Instr.* 23, 650 (1952).

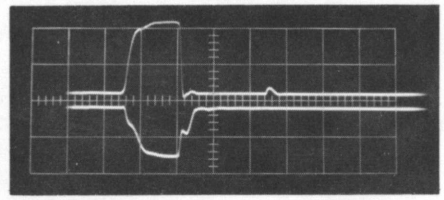


(a) Lower Schmitt circuit is triggered

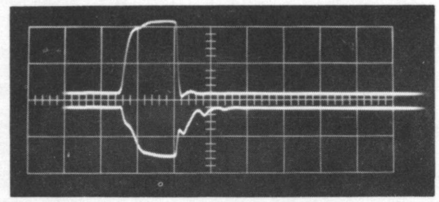


(b) Both circuits are triggered

Fig. 2.45. Waveform at the Negative Output Terminal of the Differential Discriminator.



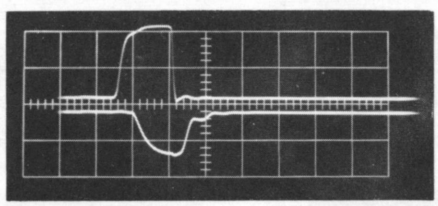
(a) Lower Schmitt circuit is triggered



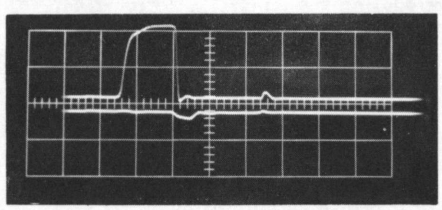
(b) Both circuits are triggered

Fig. 2.46. Waveform that Generates the Inhibit Signal.

In Fig. 2.47 is shown the inhibit signal, which prevents the triggering of the anticoincidence trigger when both Schmitt circuits are triggered. The oscilloscope is set at 10 v/cm.



(a) Lower Schmitt circuit is triggered

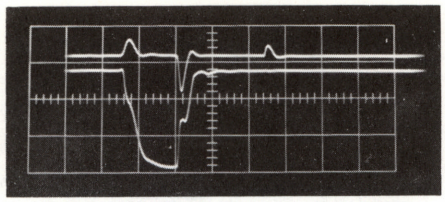


(b) Both circuits are triggered

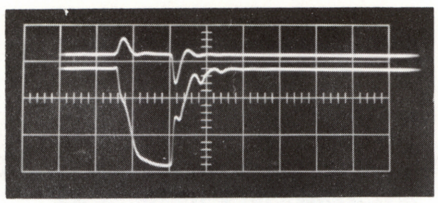
Fig. 2.47. The Inhibit Signal Waveform.

Figure 2.48 shows the time relationship between the interrogating pulse (top trace) and the inhibit-signal-generating pulse (bottom trace). An inhibit signal is generated when the upper Schmitt circuit is triggered. The inhibit signal persists until the inhibit-signal-generating pulse recovers to a point below the level of the inflection point on the trailing edge of the pulse. In Fig. 2.48, it can be seen that this does not occur until the interrogate pulse has suffered appreciable decay.

The oscilloscope is set at 10 v/cm for both traces.



(*a*) Lower Schmitt circuit is triggered

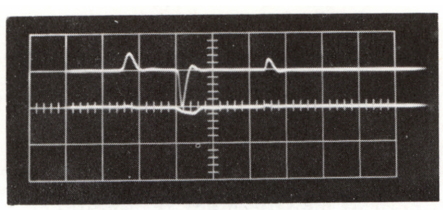


(*b*) Both circuits are triggered

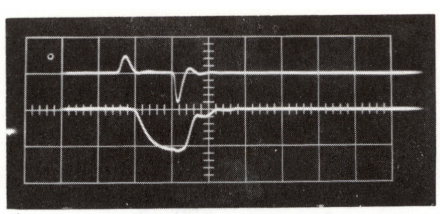
Fig. 2.48. Waveforms Showing the Time Relationship Between the Interrogating Pulse and the Inhibit-Signal-Generating Pulse.

The timing of the inhibit signal with respect to the interrogating signal is shown in Fig. 2.49. In *a*, only the lower Schmitt is triggered. No inhibit signal is generated. In *b*, both Schmitts are triggered. Note that the inhibit signal persists until the interrogate pulse is completely recovered.

The oscilloscope is set at 10 v/cm for both traces.



(*a*) Lower Schmitt circuit is triggered

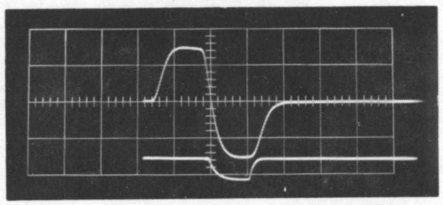


(*b*) Both circuits are triggered

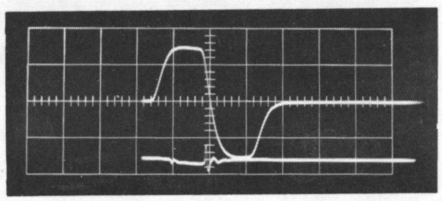
Fig. 2.49. Timing of the Inhibit Signal with Respect to the Interrogating Signal.

The timing of the differential discriminator output with respect to the amplifier signal is shown in Fig. 2.50. In *a*, only the lower Schmitt is triggered, so that an output exists. In *b*, both Schmitts are triggered. The sensitivity for the discriminator signal was increased to 0.5 v/cm to show the magnitude of the feedthrough pulse (0.35 v p-p).

For the upper traces in *a* and *b*, the oscilloscope was set at 50 v/cm and 1 μsec/cm. For the lower trace in *a* and *b*, the oscilloscope was set at 50 v/cm and 0.5 v/cm, respectively.



(*a*) Lower Schmitt circuit is triggered



(*b*) Both circuits are triggered

Fig. 2.50. Timing of the Differential Discriminator Output with Respect to the Amplifier Signal.

2.2.12 Measurement of Feedback Factors

Feedback factors can be measured with the instrument setup shown in Fig. 2.31. A rise-time spoiler of 0.005 μf should be used. The theory of measurement is discussed in Sec 1.4.

First Stage

1. Generator:	As needed to produce a 40- to 80-v output signal.
2. Main amplifier gain:	2 to 10.
3. Observation point:	Output of main amplifier.
4. Oscilloscope:	20 v/cm, 1 μsec/cm.

Note the generator dial reading and the output voltage: it is convenient to set it at some even value, like 0.10 v. Ground pin 7 of V1 and adjust the generator for the same peak output voltage. The ratio of generator voltages is the feedback factor; a ratio of 15 to 20 is normal.

Second Stage

Repeat the preceding instrument settings and procedure, but ground pin 7 of V5 instead. The feedback factor should be 12 to 18.

Third Stage

Follow the same procedure and use the same instrument settings as before, but ground pin 7 of V9 instead. The feedback factor should be 11 to 16.

2.2.13 Linearity

Follow the procedure described in Sec 1.3.6, "Measurement of Integral Nonlinearity, Method I." Nonlinearity should be no greater than 1.5 minor dial divisions between 100 and 1000 or 2 divisions between 50 and 1000.

2.2.14 Regulation of Power Supply

Follow the procedure described in Sec 1.7. Regulation should be ±0.7 v or less in the range 105 to 125 v.

2.2.15 Gain Dependence on Line Voltage

Follow the procedure described in Sec 1.8. Gain variation should be less than 3% in the range 105 to 125 v.

2.2.16 Noise Measurements

Follow the procedure described in Sec 1.5.4. See the list of specifications for the typical noise values.

2.2.17 Circuit Diagrams

The circuit diagram for the DD2 main amplifier is shown in Fig. 2.51, and that for the differential and integral pulse height selector is shown in Fig. 2.52.

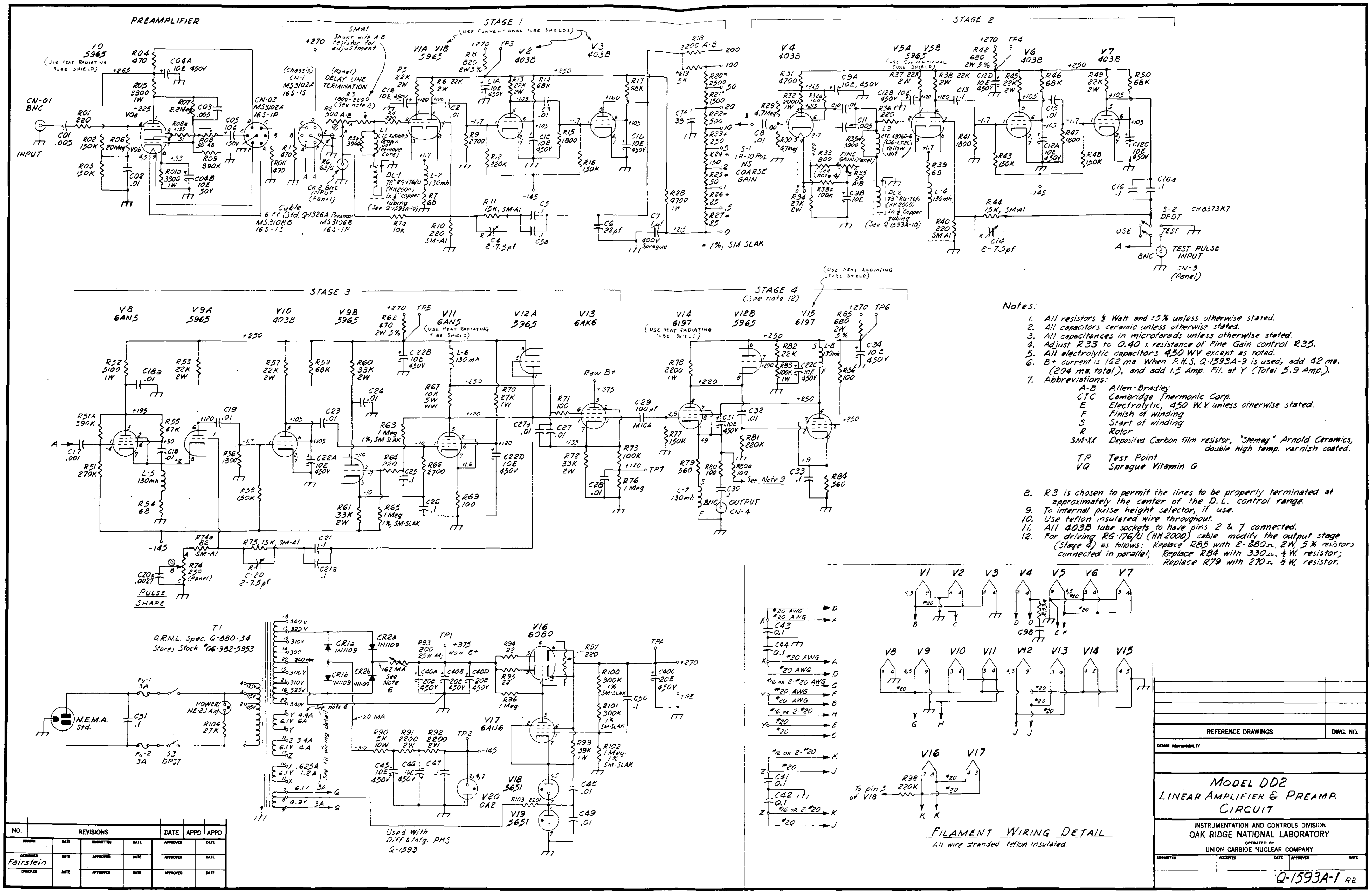


Fig. 2.51. DD2 Linear Amplifier and Preamplifier: Circuit.

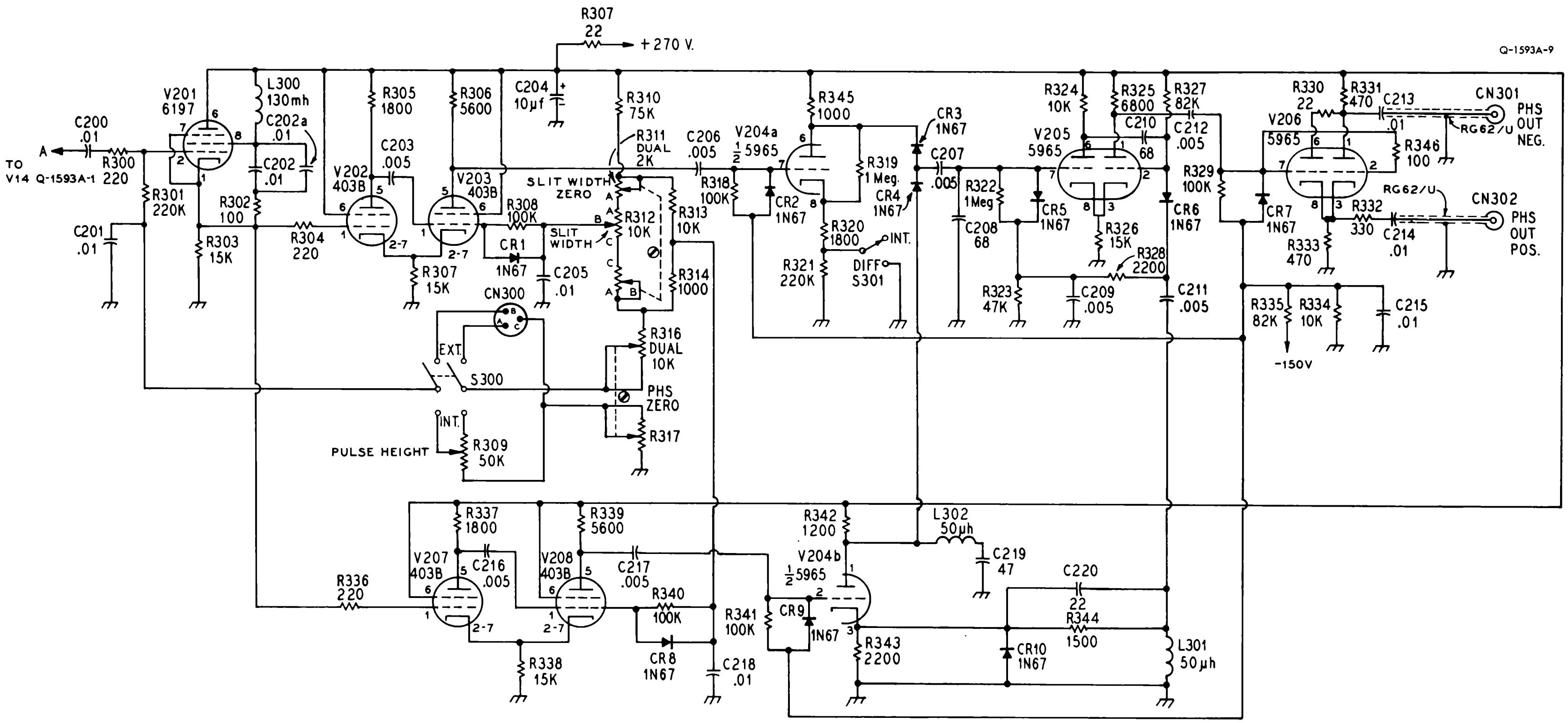


Fig. 2.52. DD2 Linear Amplifier and Preamplifier: Differential and Integral Pulse-Height-Selector Circuit.

2.3 A8 LINEAR AMPLIFIER, MODEL Q-1819-1

2.3.1 General Description

The A8 linear amplifier is a double-delay-line, nonblocking amplifier designed for use with scintillation counters (see block diagram, Fig. 2.53). Its particular attributes are its ability to recover rapidly from very large overloads (recovering to 10% of rated output in 8 μsec from a 4000× overload) and its tolerance of very high counting rates (100,000 counts/sec). It may be used with other radiation detectors if their output signal is compatible with a fixed, 1.1-μsec clipping time, a range of amplifier sensitivity from 0.24 to 370 μv/electron, and a best rms noise level equivalent to 4100 electrons with 10 pf external capacitance.

The amplifier has a built-in integral discriminator.

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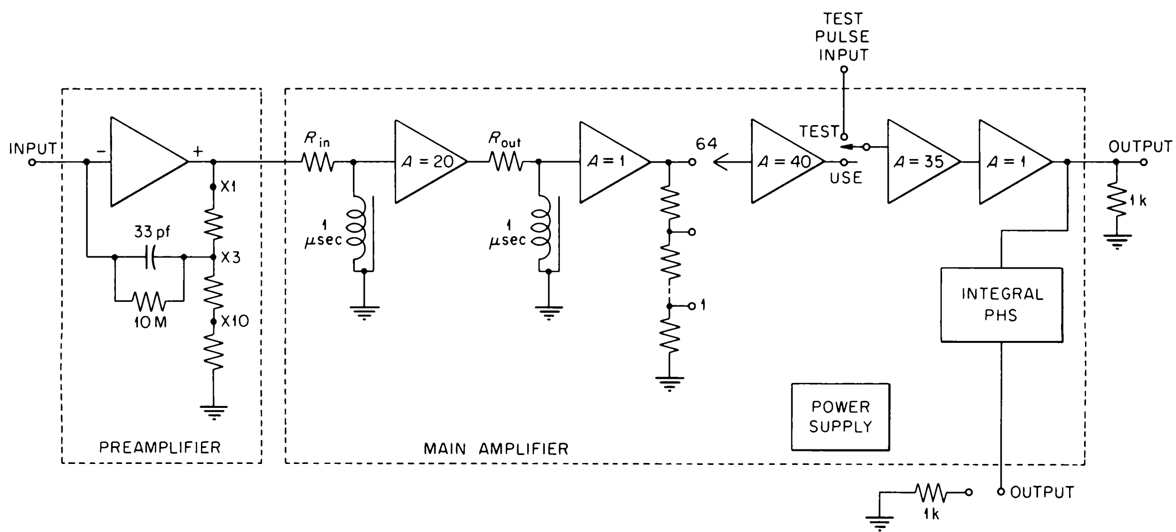

Fig. 2.53. A8 Linear Amplifier.

2.3.2 Specifications for the Main Amplifier

1. Pulse shaping:	Double-delay-line, bipolar pulse.
2. Pulse width –	
Between half-height points, primary pulse:	1.1 μsec.
Between 10% points, total pulse:	2.4 μsec.
3. Rise time from 10 to 90%:	0.24 μsec.
4. Fall time from 90 to 10%:	0.17 μsec.

5. Signal polarities –

Input:	Positive.
Output:	Positive.

6. Gain: 7700.
7. Gain control: Coarse only. In seven steps of 2, calibrated from 1 to 64 (relative values) .
8. Saturated output: 110 v.
9. Integral nonlinearity: 0.1% or less.
10. Loading: 1000 ohms recommended on the amplifier and discriminator outputs.
11. Overload performance:

Overload Factor	Recovery Time (μsec)	
	To ± 10 v	To ± 5 v
1	2.4	2.5
10	3.9	4
100	4	8
1000	7	10

12. Gain change vs line voltage:

Line Voltage	105	115	125
ΔG	– 1.2%	0	+ 0.8%

13. Line voltage at which power supply drops out of regulation: Below 95 v and above 120 v.
14. Power-supply ripple: 50 mv p-p.
15. Permissible power-supply drain: 200 ma at B^+; 18.2 amp at 6.3 v.
16. Measured power-supply drain:

Circuit Group	At B^+ (ma)	At 6.3 v (amp)
Main amplifier	43	6.1
Pulse-height selector	25	1.7
Preamplifier	11	0.6

17. Power-line drain –

Without preamplifier:	1.29 amp at 115 v.
With preamplifier:	1.34 amp at 115 v.

18. Front panel size: $8\frac{3}{4}$ by 19 in.
19. Connectors on front panel –
    - Test input: UG-290/U (BNC).
    - Amplifier output: UG-290/U (BNC).
    - PHS output: UG-290/U (BNC).
20. Connectors on rear apron –
    - Power: 2-pin, male.
    - Preamplifier: AN/MS-3102A-14S-5S, 5-pin, female.
21. Tube complement: 9 – 403B/5591; 1 – 6AU6; 3 – 6197; 1 – 6080; 1 – 5U4GB.

2.3.3 Specifications for the High-Performance Preamplifier

1. Circuit configuration: Charge sensitive.
2. Feedback elements: 33 pf shunted by 10 megohms (time constant = 330 μsec).
3. Charge sensitivity: 3×10^{10} v/coulomb 1×, 3×, or 10×. 4.8×10^{-9} v/electron 1×, 3×, or 10×.
4. Signal polarities: Negative input; positive output.
5. Output impedance: Approximately 100 ohms.
6. Permissible cable length: May be any length if the heater voltage is $\geqq 5.5$ v and the cable is correctly terminated at the preamplifier end.
7. Saturated preamplifier output voltage –
    - Single pulse, long 93-ohm cable: 14 v.
    - Pileup signal: 60 v.
8. Noise performance (with the main amplifier at maximum gain):

Detector Capacity (pf)	Preamplifier Gain		
	1×	3×	10×
	Equivalent Electrons, rms		
10	16,200	7,400	4,140
33	17,800	8,750	5,720
100	19,200	11,600	8,320
	Equivalent Charge (coulombs, rms)		
		$\times 10^{-17}$	
10	260	118	66
33	285	140	92
100	317	186	133

9. Nearest main amplifier gain setting at which the main amplifier and preamplifier noise contributions are equal:

Preamplifier Gain	Amplifier Gain
1×	2
3×	4
10×	8

10. Open-loop gain: 150.
11. Connectors –
    Input: UG-2901U (BNC).
    Output: AN-31024-14S-5P, 5-pin.
12. Tube complement: 1 – 404A/5847; 1 – 417A/5842.

2.3.4 Specifications for the Low-Performance Preamplifier

The low-performance preamplifier differs from the high-performance unit in that a 403B/5591 tube is used in place of the 404A/5847 tube, and a 6197 tube is used in place of the 417A/5842 tube. This results in an open-loop gain of approximately 60. The noise performance is about 40% worse. It is recommended that the maximum cable length be 20 ft. Other characteristics are approximately the same.

2.3.5 Specifications for the Discriminator

1. Type: Integral.
2. Resolving time: Duration of the drive pulse plus 0.1 μsec.
3. Pulse-height range: 2 to 75 v.
4. Output pulse: 40-v negative with 1000-ohm load. The pulse width depends on the width and height of the driving pulse.
5. Hysteresis: 10 minor dial divisions of PHS control.
6. Zero shift vs line voltage:

	Line Voltage (v)		
	105	115	125
Zero shift, minor dial divisions	−4.5	0	+1.2

7. Current drain: 25 ma at B^+; 1.7 amp at 3 v.
8. Tube complement: 1 – 6211; 3 – 403B/5591; 1 – 6201; 1 – 6197.

2.3.6 Tube-Voltage Chart

Typical tube voltages for A8 linear amplifier tubes are given in Table 2.6.

Table 2.6. Typical Tube Voltages for A8 Linear Amplifier Tubes

The line voltage was 115 v ac, unregulated B^+ was 375 v, unregulated ripple voltage 8 v p-p, regulated ripple voltage was 50 mv p-p (both ripple voltages were measured with an oscilloscope), and dc regulation (see Sec 1.7) from 95 v to 105 v input $\cong$ 0.2 v

Tube No.	Pin Number, Tube Element (e),* and Voltage (v)**																	
	1		2		3		4		5		6		7		8		9	
	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v
V1, 404A/5847	G	–[a]	[b]		H	0	C	1.62			P	100			Sc	110	H	0 5.6 ac
V2, 417A/5842	P	250			H	0	G	100			C	103			G	100	H	0 5.6 ac
V3, 403B/5591	G	–	C	110	H	0	H	0 5.4 ac	P	240	Sc	255	C	110				
V4, 403B/5591	G	0	C	1.9	H	0	H	0 5.4 ac	P	110	Sc	135	C	1.9				
V5, 403B/5591	G	0	C	2.0	H	0	H	0 5.4 ac	P	123	Sc	124	C	2.0				
V6, 403B/5591	G	0	C	1.2	H	0	H	0 5.4 ac	P	95	Sc	131	C	1.2				
V7, 403B/5591	G	–	C	99	H	0	H	0 5.4 ac	P	238	Sc	250	C	99				
V8, 403B/5591	G	0	C	1.9	H	0	H	0 5.4 ac	P	99	Sc	82	C	1.9				
V9, 403B/5591	G	0.3	C	1.88	H	0 5.4 ac	H	0	P	109	Sc	117	C	1.88				
V10, 403B/5591	G	0	C	1.08	H	0	H	0 6.0 ac	P	112	Sc	91	C	1.08				
V11, 403B/5591	G	0	C	1.98	H	0	H	0 6.0 ac	P	202	Sc	126	C	1.98				
V12, 6197	C	1.7	G	0	Sc	104	H	0 6.0 ac	H	0	P	170	Su	0	Sc	104	G	0
V13, 6197	C	7.7	G	0.75	Sc	243	H	0 6.0 ac	H	0	P	255	Su	7.7	Sc	243	G	0.75

V14, 6197	C 9.6	G 0	Sc 255	H 0 6.0 ac	H 0	P 243	Su 0	Sc 255	G 0
V15, 5U4GB		H[c] 375 5.4 ac		P 350 ac		P 350 ac		H[c] 375 5.4 ac	
V16, 6080	G1 225	P1 340	C1 258	G2 225	P2 340	C2 258	H[d] 255 6.2 ac	H[d] 255 6.2 ac	
V17, 6AU6	G –	Su 166	H[e] 255 6.2 ac	H[e] 255 6.2 ac	P 225	Sc 205	C 166		
V18, 85A2/OG3	P 165	C 84		C 84	P 165		C 84		
V19, 85A2/OG3	P 165	C 84		C 84	P 165		C 84		
V20, 403B/5591	G –	C 75	H[e] 0 6.1 ac	H[e] 0 6.1 ac	P 227	Sc 195	C 75		
V21, 403B/5591	G –	C 75	H[e] 0 6.1 ac	H[e] 0 6.1 ac	P 225	Sc 195	C 75		
V22, 403B/5591***	G 16	C 16	H[e] 0 6.1 ac	H[e] 0 6.1 ac	P 75	Sc 107	C 16		
V23, 6211	P1 255	G1 –	C1 194	H[f] 0 6.1 ac	H[f] 0 6.1 ac	P2 255	G2 75	C2 150	H[f] 0 6.1 ac
V24, 6197	C 4.8	G 0	Sc 140	H[g] 0 6.1 ac	H[g] 0 6.1 ac	P 235	Su 4.8	Sc 140	G 0
V25, 6201	P1 255	G1 –	C1 80	H[g] 0 6.0 ac	H[g] 0 6.0 ac	P2 –	G2 –	C2 42	H 0

*The tube elements are indicated by letters: G, grid; C, cathode; H, heater; P, plate; Sc, screen grid; and Su, suppressor grid. A number following the letter indicates the sequence of the element, i.e., G1 is the first grid and G2 is the second grid.

**All voltages were measured with a 20,000-ohms/v multimeter. The voltages are positive with respect to ground and are dc voltages unless specified as ac.

***R99 should be changed from 47K to 20K.

[a]Dashes indicate that the impedance levels were too high for a meaningful multimeter reading.

[b]Blanks indicate unused pins.

[c]The ac heater voltage was measured between pins 2 and 8 and not to ground.

[d]The ac heater voltage was measured between pins 7 and 8.

[e]The ac heater voltage was measured between pins 3 and 4.

[f]The ac heater voltage was measured between pins 4–5 and 9.

[g]The ac heater voltage was measured between pins 4 and 5.

2.3.7 Adjustments

Connect the test setup shown in Fig. 2.54. Couple the generator to the preamplifier through a 33-pf capacitor fitting. (See Fig. 1.16 for a description of the capacitor fitting.) Use the 10× oscilloscope probe for all tests. Terminate the main amplifier and PHS outputs in 1000 ohms.

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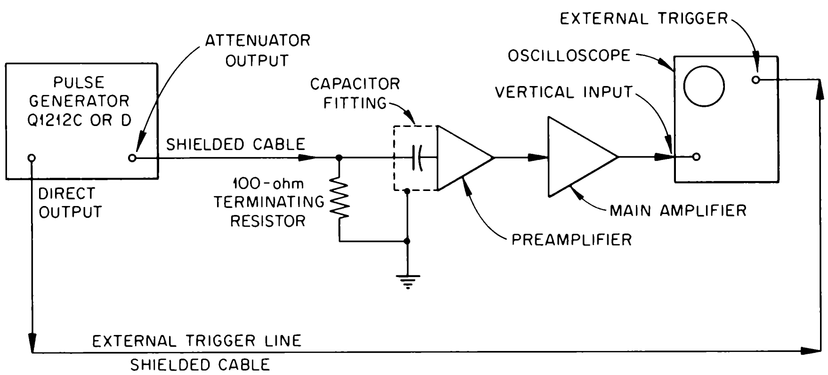


Fig. 2.54. Test Setup for Adjustment of an A8 Linear Amplifier.

Preamplifier Transient Response

(*a*) This procedure applies to the high-performance unit only. The test instruments should be set as follows:

1. Generator:	1-v output, negative.
2. Main amplifier gain:	Minimum.
3. Observation point:	Main amplifier input (junction of R14 and R16).
4. Oscilloscope:	1 v/cm; 1 μsec/cm. (The 1 v/cm refers to the sensitivity as observed on the face of the tube. With a 10× probe the dial setting on the oscilloscope would be 0.1 v/cm.)
5. Preamplifier gain:	

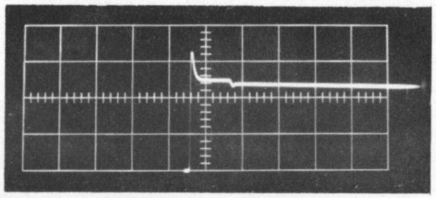
(*b*) Set C5 for optimum transient response (Fig. 2.55).

(*c*) Set the trimming capacitor at its minimum value (Fig. 2.56) on the gain 1× and 3× positions.

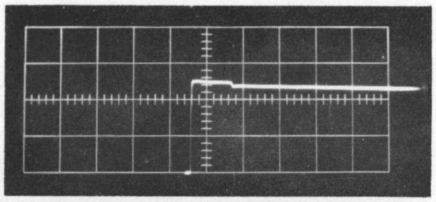
Adjustment of Main Amplifier Transient Response

(*a*) Set the test instruments as follows:

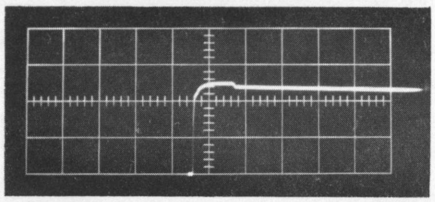
1. Generator:	0.1-v output.
2. Main amplifier gain:	Minimum.
3. Observation point:	Top of gain control (junction of C22 and R49).
4. Oscilloscope:	1 v/cm; 1 μsec/cm.
5. Preamplifier gain:	3× for remainder of tests (transient response optimized).



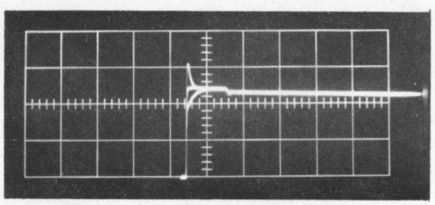
(*a*) C5 set too low



(*b*) C5 optimum



(*c*) C5 set too high



(*d*) Composite of *a*, *b*, and *c*

Fig. 2.55. Waveforms Showing the Effect of C5 on Optimum Transient Response.

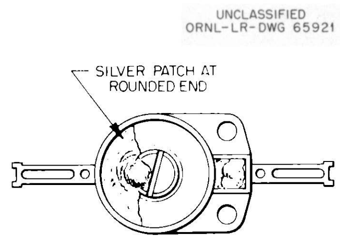
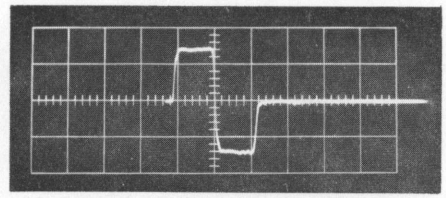
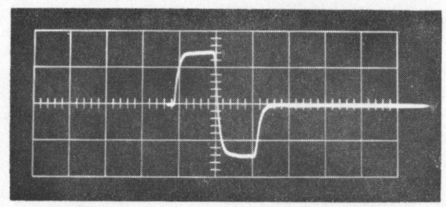


Fig. 2.56. Trimming Capacitor Set at Its Minimum Value.



(*a*) C15 set too low



(*b*) C15 set optimum

Fig. 2.57. Waveforms Showing Effect of C15 on Transient Response.

(*b*) Set C15 to optimize transient response (Fig. 2.57).

(*c*) Connect the oscilloscope probe to the output of the amplifier. Increase the generator output to produce a 90-v signal at the amplifier output terminal (approx. 0.32 v from the generator). Set C27 and C34 to minimum capacitance. The resulting waveform should be that of Fig. 2.58*a*.

(*d*) Set C34 for optimum response as shown in Fig. 2.58*b*. Note that some overshoot occurs on the top and some ringing occurs on the leading edge of the negative pulse.

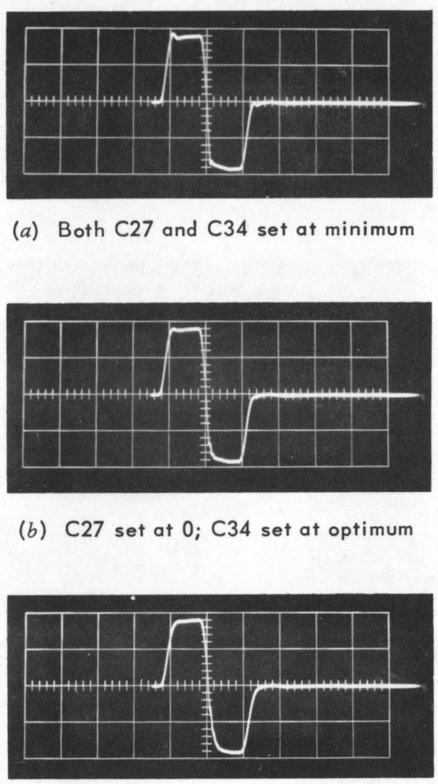
(*e*) Set C27 just high enough to remove the overshoot from the top of the pulse, as shown in Fig. 2.58*c*.

Adjustment of the First Delay-Line Termination

(*a*) The instrument settings are:

1. Generator:	1-v output.
2. Main amplifier gain:	Minimum.
3. Observation point:	Output of first white cathode follower (junction of R23 and R24).
4. Oscilloscope:	0.5 v/cm; 1 μsec/cm.

(*b*) Set the first delay-line termination (R14) for optimum response (Fig. 2.59).

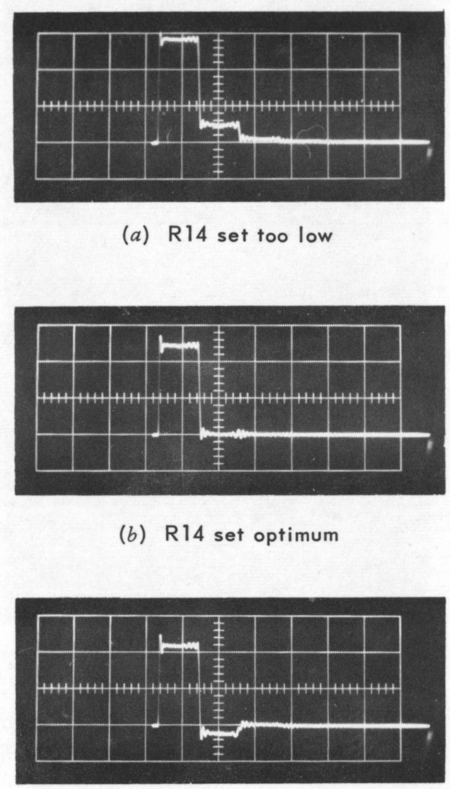


(*a*) Both C27 and C34 set at minimum

(*b*) C27 set at 0; C34 set at optimum

(*c*) Both C27 and C34 set at optimum

Fig. 2.58. Waveforms Showing Capacitance Effects of C27 and C34.



(*a*) R14 set too low

(*b*) R14 set optimum

(*c*) R14 set too high

Fig. 2.59. Waveforms Showing Effect of R14.

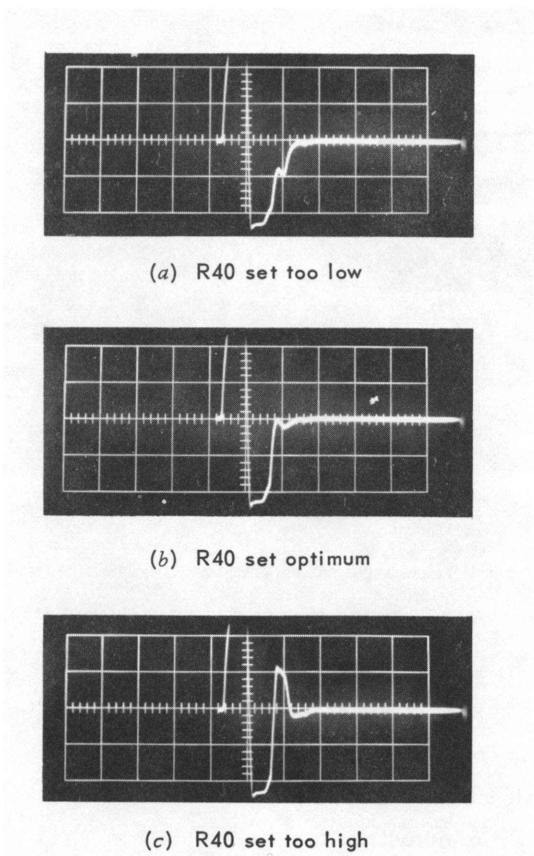
Preliminary Setting of Second Delay-Line Termination

(*a*) Instrument settings:

1. Main amplifier gain:	Maximum.
2. Observation point:	Amplifier output.
3. Oscilloscope:	50 v/cm; 2 μsec/cm.

(*b*) Cover the output section of the amplifier box with a sheet of metal (the PHS box cover may be used). Set the generator for an amplifier output signal of 90 v (approximately 5 mv from the generator). Then increase the generator output 10×. Adjust the terminating resistor (R40) of the second line for optimum response (Fig. 2.60).

(*c*) Reduce the generator signal to its original value and observe the pulse shape. It should appear as in Fig. 2.61*a*. If the shield over the output section of the amplifier is removed, a slight baseline tilt occurs after the pulse (Fig. 2.61*b*), which is evidence of unwanted feedback between output and input.

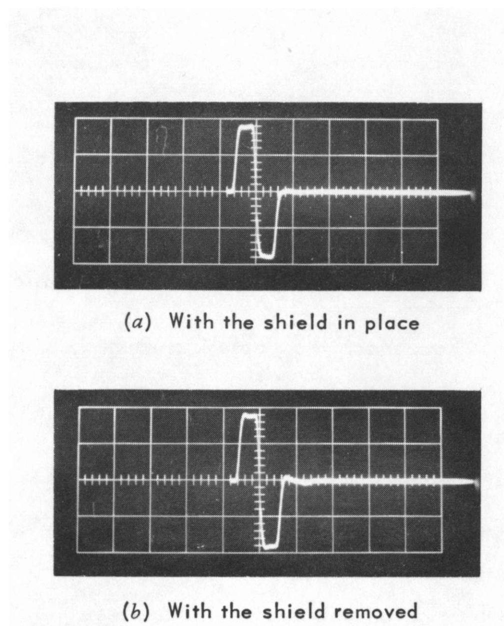


(*a*) R40 set too low

(*b*) R40 set optimum

(*c*) R40 set too high

Fig. 2.60. Waveforms Showing the Effect of R40. All settings are at 10X overload.



(*a*) With the shield in place

(*b*) With the shield removed

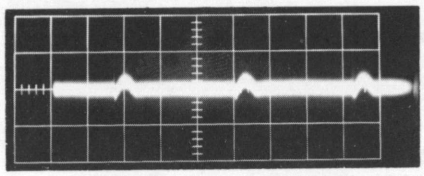
Fig. 2.61. Waveforms Showing the Effect of the Shield Over the Output Section of the Amplifier.

Adjustment of DC Compensation

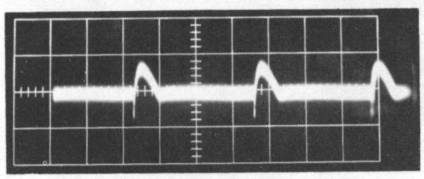
(*a*) Instrument settings:

1. Generator: 5-v output.
2. Main amplifier gain: Maximum.
3. Observation point: Amplifier output.
4. Oscilloscope: 10 v/cm; 5 msec/cm.

(*b*) Remove the cable between the generator and the external trigger input connector of the oscilloscope. Trigger the sweep from the 60-cycle line. Adjust R17 for minimum baseline disturbance. (See Fig. 2.62. The broad trace is due to amplifier noise.)



(*a*) R17 set optimum



(*b*) R17 set at extreme end

Fig. 2.62. R17 Adjusted for Minimum Baseline Disturbance.

Final Setting of Second Delay-Line Termination

(*a*) Reconnect the cable between the pulser and the oscilloscope, and trigger the sweep from the pulser. During part of the test, a low pulser output level is required, and it may be necessary to trigger the sweep internally.

(*b*) Instrument settings:

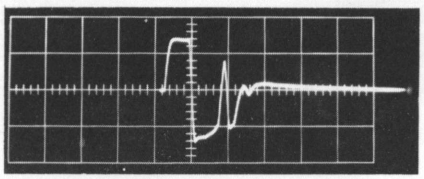
1. Generator: Click in the 5× attenuator switch.
2. Main amplifier gain: Maximum.
3. Observation point: Amplifier output.
4. Oscilloscope: 100 v/cm; 2 μsec/cm.

(*c*) Adjust the generator pulse-height dial for a maximum secondary positive spike from the amplifier (it may be necessary to change the attenuator setting to do this). If no such spike shows, rotate R40 to the point where a positive-going secondary spike is just about to cross the baseline. This is the optimum setting for R40 (Fig. 2.63).

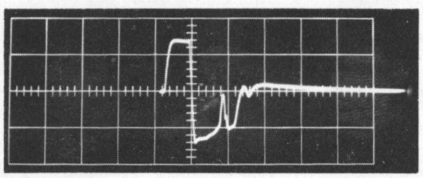
Adjustment of Area Balance

1. Generator: Set for 10× overload.
2. Main amplifier gain: Maximum.
3. Observation point: Main amplifier output.
4. Oscilloscope: 100 v/cm; 2 μsec/cm.

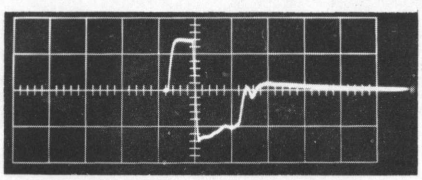
After the instruments are set as listed, R71 should be adjusted to give equal positive and negative pulse heights (Fig. 2.64). Ideally this adjustment should be made under actual operating conditions. The trimmer should be adjusted for minimum peak shift when going from a low to a high counting rate.



(*a*) R40 set too low

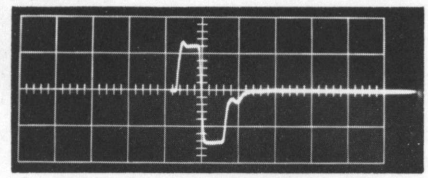


(*b*) R40 optimum

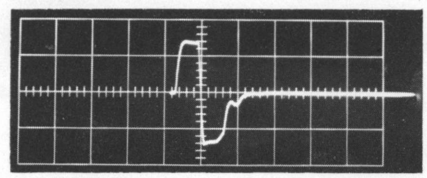


(*c*) R40 set too high

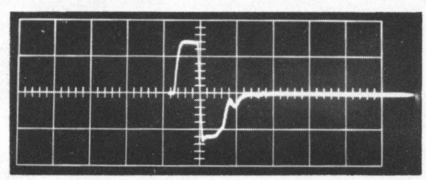
Fig. 2.63. Adjustment of R40 to Optimum.



(*a*) R71 set too low



(*b*) R71 set at optimum



(*c*) R71 set too high

Fig. 2.64. Adjustment of R71.

2.3.8 Overload Waveforms

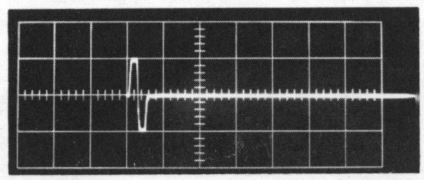
The test setup shown in Fig. 2.54 is used for studying overload waveforms.

The original specification for an A8 linear Amplifier required overload tests to 4000×. Since, however, the worst response (incipient double pulse) occurs at an overload level of less than 1000× (see Sec 2.3.7, "Final Setting of Second Delay-Line Termination"), it may be assumed that if performance is satisfactory up to 1000× overload it will also be satisfactory at 4000× overload. The waveforms resulting from tests up to 1000× overload are shown in Fig. 2.65.

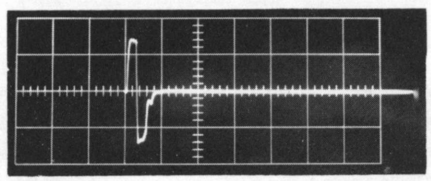
The break in the baseline at 15 μsec from the start of the pulse in Fig. 2.65*d* is a pulse-generator artifact and should be disregarded.

The instrument settings for the test are the following:

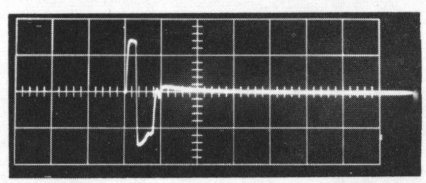
1. Generator:	Set as needed.
2. Main amplifier gain:	Maximum.
3. Observation point:	Main amplifier output.
4. Oscilloscope:	100 v/cm; 5 μsec/cm.



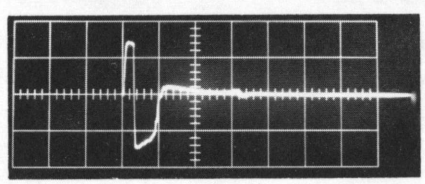
(*a*) Incipient overload



(*b*) 10X



(*c*) 100X



(*d*) 1000X

Fig. 2.65. Waveforms Resulting from Overload.

2.3.9 Miscellaneous Amplifier Overload Waveforms

In all the following tests, the overload levels refer to the main amplifier output terminal.

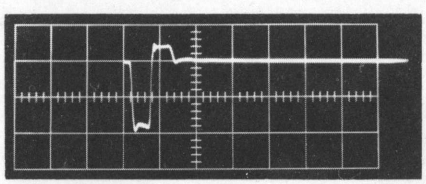
Test 1. – The waveforms from the following instrument settings are shown in Fig. 2.66.

1. Generator:	As needed.
2. Main amplifier gain:	Maximum.
3. Observation point:	Pin 6 of V13 (plate).
4. Oscilloscope:	20 v/cm; 2 μsec/cm.

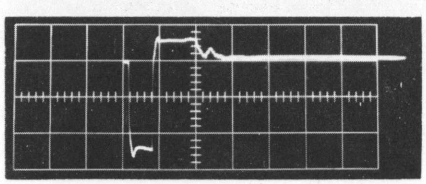
Test 2. – The waveforms from the following instrument settings are shown in Fig. 2.67.

1. Generator:	As needed.
2. Main amplifier gain:	Maximum.
3. Observation point:	Junction of CR8 and CR9.
4. Oscilloscope:	2 v/cm; 2 μsec/cm.

In Fig. 2.67*d*, a generator artifact may be seen at 14 μsec.

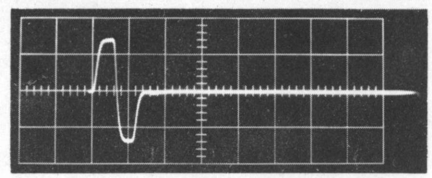


(*a*) Incipient overload

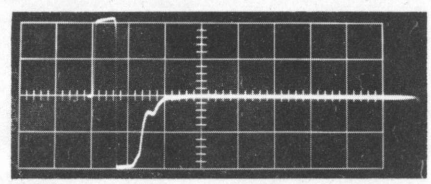


(*b*) 100X overload

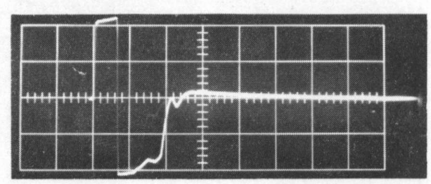
Fig. 2.66. Waveforms Resulting from Test 1.



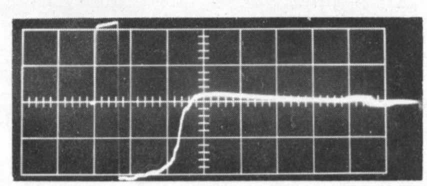
(*a*) Incipient overload



(*b*) 10X



(*c*) 100X



(*d*) 1000X

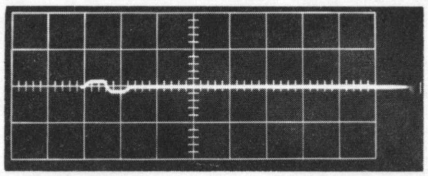
Fig. 2.67. Waveforms Resulting from Test 2.

Test 3. – The waveforms from the following instrument settings are shown in Fig. 2.68.

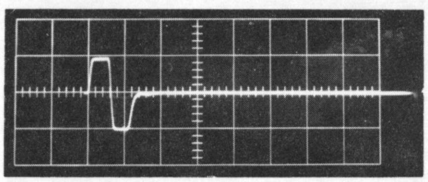
1. Generator: As needed.
2. Main amplifier output: Maximum.
3. Observation point: Pin 1 of V9.
4. Oscilloscope: 0.5 v/cm; 2 μsec/cm.

Test 4. – The waveforms from the following instrument settings are shown in Fig. 2.69.

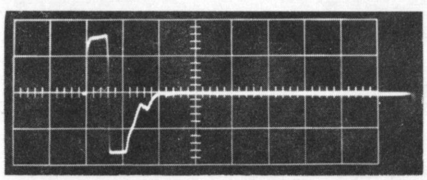
1. Generator: As needed.
2. Main amplifier output: Maximum.
3. Observation point: Junction CR5 and C13.
4. Oscilloscope: 1 v/cm; 2 μsec/cm.



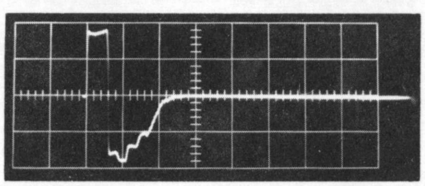
(*a*) Incipient overload



(*b*) 10X

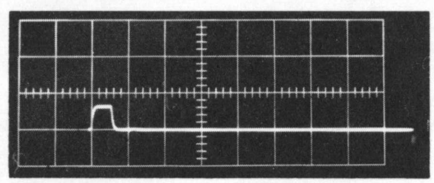


(*c*) 100X

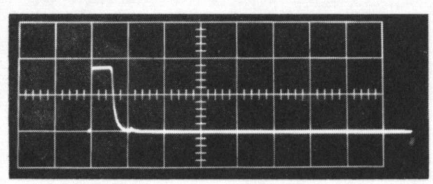


(*d*) 1000X

Fig. 2.68. Waveforms Resulting from Test 3.



(*a*) 100X overload (this point is not saturated)



(*b*) 1000X overload (saturated)

Fig. 2.69. Waveforms Resulting from Test 4.

2.3.10 Discriminator Output Waveform

Figure 2.70 shows a discriminator output waveform. The small pulse is obtained with the output terminated in 1 kilohm, and the large pulse is obtained with the output not terminated. The instruments should be adjusted to the following settings.

1. Generator setting: As needed.
2. Main amplifier output: As needed.
3. Observation point: Discriminator output terminal.
4. Oscilloscope: 100 v/cm; 1 μsec/cm.
5. PHS dial: Set low for positive triggering of PHS.

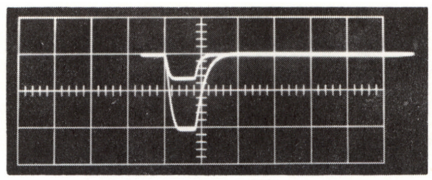


Fig. 2.70. Discriminator Output Waveform.

2.3.11 Measurement of Feedback Factors

The test setup shown in Fig. 2.54 is used to measure feedback factors. A general discussion of measurement of feedback factors is given in Sec 1.4.2.

High-Performance Preamplifier

The coupling capacitor between the pulser and the input to V1 should be removed. If a capacitor (C1) is used in the preamplifier, the capacitor should be short-circuited. The instruments should be set as follows for this measurement.

1. Generator: 0.01 v.
2. Main amplifier gain: Minimum.
3. Observation point: Input to main amplifier (junction of R14 and R16).
4. Oscilloscope: 0.5 v/cm; 2 μsec/cm.

If the preamplifier is working correctly, approximately 1.5 v of signal will be obtained, yielding a μ factor of $1.5/0.01 = 150$.

First Feedback Loop of the Main Amplifier

To prevent oscillations, disconnect the delay line and connect a 1500-ohm resistor between the junction of C8, C9, and ground. Connect the generator to the preamplifier through a 33-pf capacitor. Set the instruments as follows:

1.	Generator:	As needed to produce a 40- to 100-v output signal.
2.	Main amplifier gain:	Minimum.
3.	Observation point:	Output of main amplifier.
4.	Oscilloscope:	20 v/cm; 1 μsec/cm.

Note the generator dial reading: it is convenient to set it at some even value, such as 1.00 v. Shunt R39 with a 100-μf capacitor, and readjust the generator for the same amplifier output voltage. The ratio of generator voltages is the feedback factor; a ratio of 25 is normal.

Second Feedback Loop of the Main Amplifier

Move the capacitor shunt to R68 and repeat the procedure given for the first feedback loop. A normal value for the factor ($\mu\beta + 1$) is 20.

2.3.12 Linearity

Follow the procedure in Sec 1.3.6, "Measurement of Integral Nonlinearity, Method I." Nonlinearity should be no greater than 1 minor dial division.

2.3.13 Regulation of Power Supply

Follow the procedure described in Sec 1.7. Regulation should be ±0.5 v or less.

2.3.14 Gain Dependence on Line Voltage

Follow the procedure described in Sec 1.8. Gain variation should be less than 2.5% from 105 to 120 v.

2.3.15 Noise Measurements

Follow the procedure described in Sec 1.5.4. See the list of specifications for typical noise values.

2.3.16 Circuit Diagram

The circuit diagram for the A8 linear amplifier is shown in Fig. 2.71.

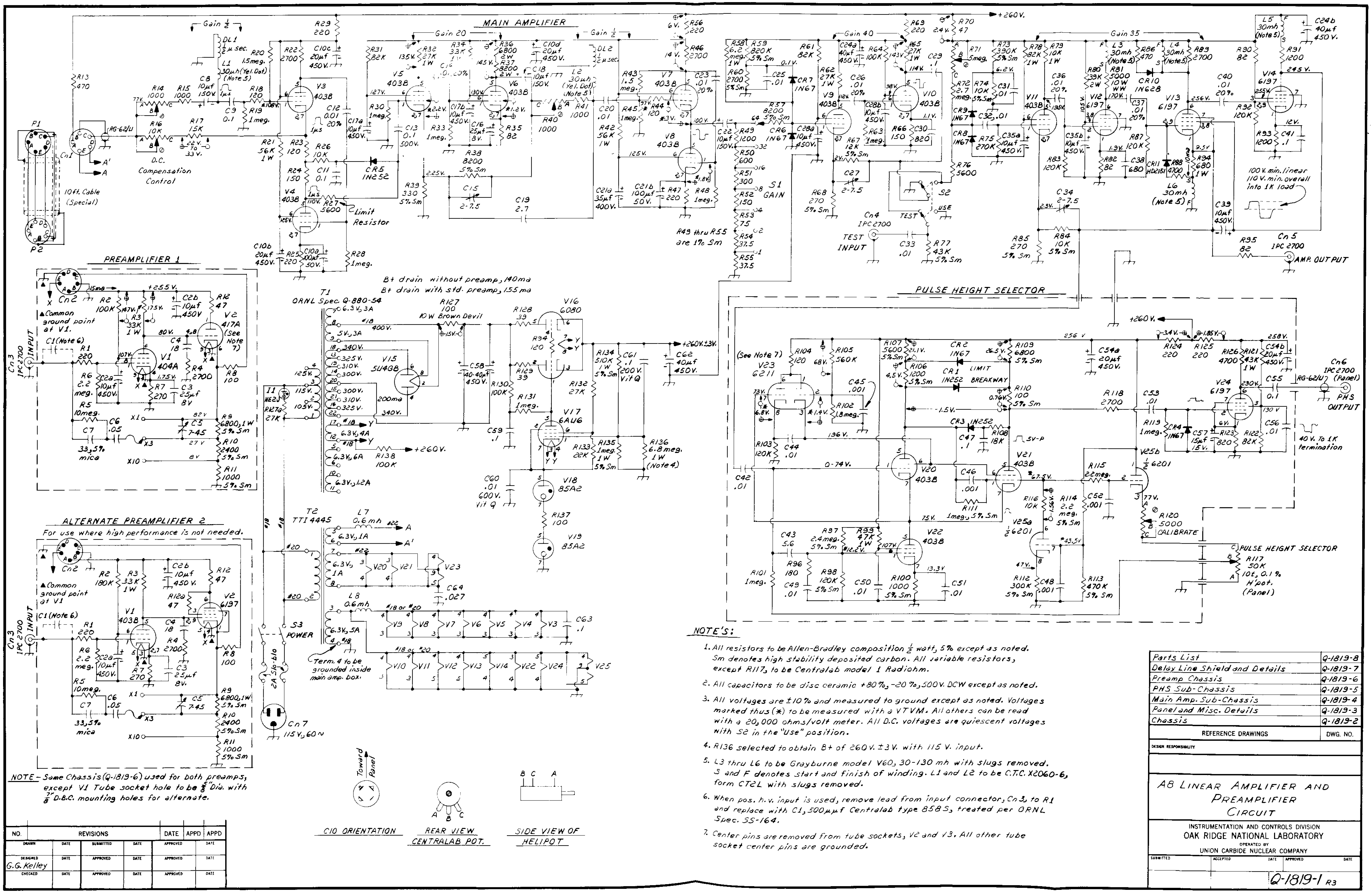


Fig. 2.71. A8 Linear Amplifier and Preamplifier: Circuit.

2.4 A61 LINEAR AMPLIFIER, RIDL MODELS 30-7, 30-8, AND 30-11

2.4.1 General Description

The A61 amplifier (block diagram in Fig. 2.72) was designed for use with scintillation counters. Double RC clipping networks result in good tolerance to high counting rates and large overloads. These characteristics are obtained only when the detector load resistor is adjusted to produce a 1-μsec clipping time at the detector.

The amplifier has a built-in integral discriminator. A pulse generator is also built in for checking the multichannel analyzer with which the amplifier is normally used.

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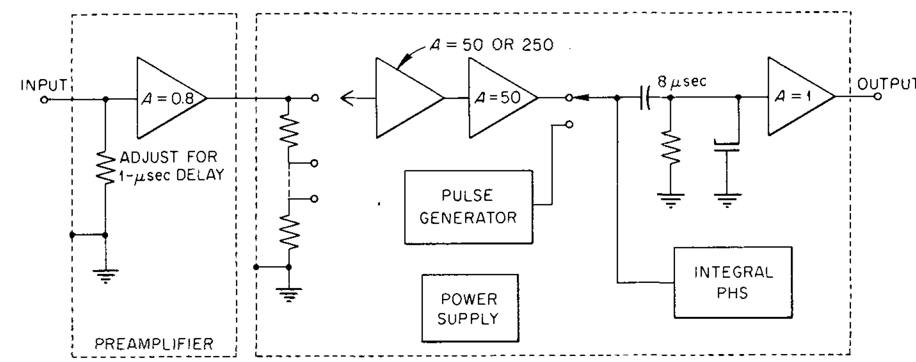

Fig. 2.72. A61 Linear Amplifier.

2.4.2 Specifications for the Main Amplifier

1. Pulse shaping: Double RC.
2. Pulse width –
   - Between half-height points: 1.7 μsec.
   - Between 10% points: 3.4 μsec.
3. Rise time from 10 to 90%: 0.4 μsec.
4. Fall time from 90 to 10%: 2.4 μsec.
5. Signal polarities –
   - Input: Negative.
   - Output: Positive.
6. Gain: 2800.
7. Gain control –
   - Coarse: $\frac{1}{16}$ to 1 in steps of 2, plus a 5× hi-lo gain switch (relative values).
   - Fine: None.

8. Saturated output:	130 v.
9. Integral nonlinearity:	0.1% from 2 to 80 v. Nonlinear below 2 v.
10. Loading:	Approximately 10 kilohms for 1% nonlinearity.
11. Overload performance:	

Overload Factor	Recovery Time (μsec)	
	To $\pm$ 10 v	To $\pm$ 5 v
1	2.4	3.4
10	9	10
100	150	165
200	220	255

12. Gain change vs line voltage:	Less than 0.1% from 105 to 125 v.
13. Line voltage at which power supply drops out of regulation:	80 v.
14. Power-supply ripple:	90 mv p-p.
15. Front panel size:	$8\frac{3}{4}$ by 19 in.
16. Connectors on front panel –	
Output 1:	UG-290/U (BNC).
Output 2:	UG-290/U (BNC).
17. Connectors on rear apron –	
Amplifier input:	BNC.
High voltage input:	UG-931/U (IPC-27,000).
Amplifier output:	UG-290/U (BNC).
PHS output:	Both positive and negative are UG-290/U (BNC).
Preamplifier 1:	AN-3102A-14S-6S.
Preamplifier 2:	AN-3102A-16S-1P
18. Tube complement:	1 – 6AG5; 2 – 6AH6; 2 – 6BK7; 3 – 6CL6; 2 – 6AS5; 1 – 6AU6; 1 – 0A3; 1 – 85A2.
19. Relay:	Clare HG-1003.

2.4.3 Specifications for the Preamplifier

1. Circuit configuration:	Cathode follower.
2. Gain:	0.41 (with differentiator).
3. Input capacity:	Approximately 9 pf.
4. Signal polarities:	Negative input; negative output.
5. Output impedance:	Approximately 200 ohms.

6. Permissible cable length: Approximately 16 ft.
7. Saturated output: 26-v negative.
8. Noise performance:

Detector Capacity (pf)	Noise (equivalent electrons, rms)
5	27,500
15	37,500
100	100,000

9. Nearest main amplifier gain setting at which main amplifier and preamplifier noise contributions are equal: $\frac{1}{2}$.
10. Connectors –
    - Test input: BNC.
    - Power: AN-3102A-14S-6P.
11. Tube complement: 1 – 6AK5.

2.4.4 Specifications for the Discriminator

1. Type: Integral.
2. Resolving time: 1.8 μsec/min.
3. Output pulse –
    - Positive output: 105 v.
    - Negative output: 120 v.
4. Zero shift vs line voltage: Less than 1 minor dial division from 105 to 125 v.
5. Tube complement: 1 – 6AL5; 1 – 6AK5; 1 – 6CF6; 1 – 6CL6.

2.4.5 Tube-Voltage Chart

Typical tube voltages for A61 linear amplifier tubes are given in Table 2.7.

2.4.6 Amplifier Waveforms

Connect the test setup shown in Fig. 2.73. Examine the preamplifier input circuit to be sure that R1 is 100 kilohms; otherwise, change R1. Use a 10× oscilloscope probe for all measurements. See Fig. 1.16 for a description of the capacitor fitting.

Various waveforms are shown in Fig. 2.74.

Table 2.7. Typical Tube Voltages for A61 Linear Amplifier Tubes

The line voltage was 115 v ac, unregulated B^+ before the choke was 408 v, unregulated B^- at the rectifier was – 345 v, the B^- supply was – 277 v, and dc heater supply was 5.4 v

Tube No.	Pin Number, Tube Element (e),* and Voltage (v)**																	
	1		2		3		4		5		6		7		8		9	
	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v
Preamplifier, 6AK5	G	19.2	C	26.2	H	5.3	H	0	P	232	Sc	232	C	26.2	*a*			
V1, 6AG5	G	–[b]	C	0.2	H[c]	0 6.4 ac	H[c]	0 6.4 ac	P	82	Sc	82	C	0.2				
V2, 6AH6	G	82	Su	83	H[c]	0 6.4 ac	H[c]	0 6.4 ac	P	167	Sc	213	C	83				
V3, 6BK7	P	118	G	1.6	C	4.4	H[d]	0 6.4 ac	H[d]	0 6.4 ac	P	213	G	116	C	118		
V4, 6AH6	C	0	Su	1.93	H[c]	0 6.4 ac	H[c]	0 6.4 ac	P	206	Sc	114	C	1.93				
V5, 6CL6	C	3.75	G	0	Sc	148	H[d]	0 6.4 ac	H[d]	0 6.4 ac	P	75	Su	375	Sc	148	G	0
V6, 6CL6	C	81	G	75	Sc	284	H[d]	0 6.4 ac	H[d]	0 6.4 ac	P	284	Su	81	Sc	284	G	75
V7, 6AL5	C1	0.12	P2	0 to 79	H[c]	0 6.4 ac	H[c]	0 6.4 ac	C2[e]	282			P1	85				
V8, 6CL6	C	10.0	G	0.12	Sc	284	H[d]	0 6.4 ac	H[d]	0 6.4 ac	P	284	Su	10	Sc	284	G	0.12

Table 2.7 (continued)

Tube No.	Pin Number, Tube Element (e),* and Voltage (v)**																	
	1		2		3		4		5		6		7		8		9	
	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v	e	v
V9, 6AK5	G	0 to 79	C	85	H[c]	0 6.4 ac	H[c]	0 6.4 ac	P[e]	282	Sc[e]	282	C	85				
V10, 6CF6	G	84	Su	85	H[c]	0 6.4 ac	H[c]	0 6.4 ac	P	255	Sc	245	C	85				
V11, 6CL6	C	5	G	0	Sc	204	H[d]	0 6.4 ac	H[d]	0 6.4 ac	P	278	Su	5	Sc	204	G	0
V12, V13, 6AS5	C	297	G	252	H[c]	235 6.45 ac	H[c]	235 6.45 ac	G	252	Sc	400	P	400				
V14, 6AU6	G	67	Su	73	H[c]	0 6.4 ac	H[c]	0 6.4 ac	P	252	Sc	400	C	400				
V15, OA3			C	0					P	73								
V16, 85A2	P	85	C	0					P	85								

*The tube elements are indicated by letters: G, grid; C, cathode; H, heater; P, plate; Sc, screen grid; and Su, suppressor grid. A number following the letter indicates the sequence of the element; that is, G1 is the first grid and G2 is the second grid.

**All voltages were measured with a 20,000-ohms/v multimeter. The voltages are positive with respect to ground and are dc voltages unless specified as ac.

[a]Blanks indicate unused pins.

[b]Dashes indicate that impedance levels were too high for a meaningful multimeter reading.

[c]The ac heater voltage was measured between pins 3 and 4.

[d]The ac heater voltage was measured between pins 4 and 5.

[e]The PHS dial was set above 100.

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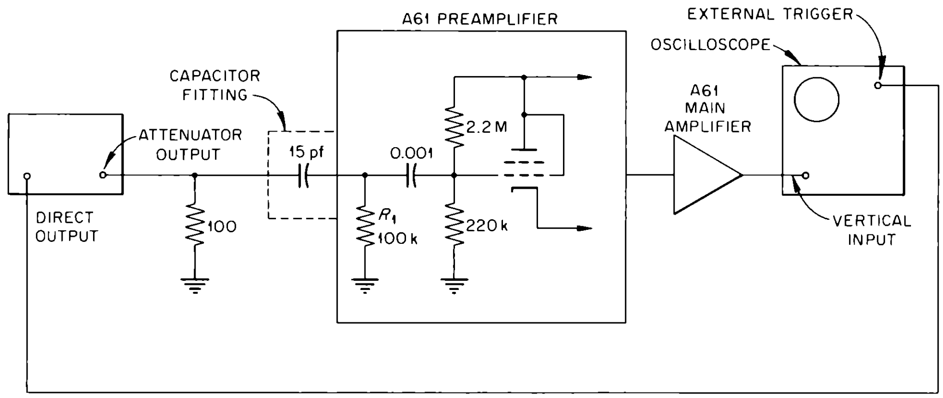
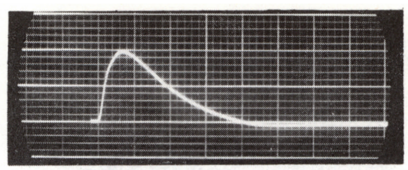
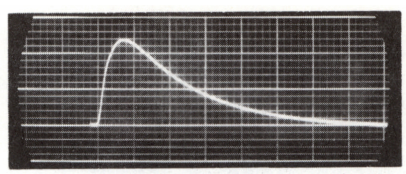


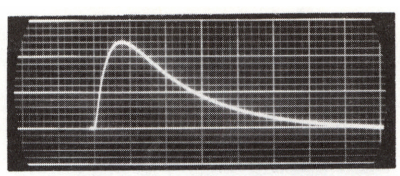
Fig. 2.73. Test Setup for Adjustment of an A61 Linear Amplifier.



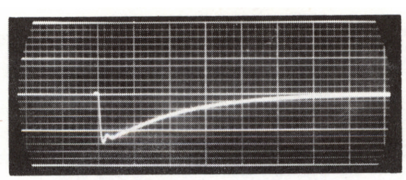
(*a*) At the amplifier output; oscilloscope at 50 v/cm



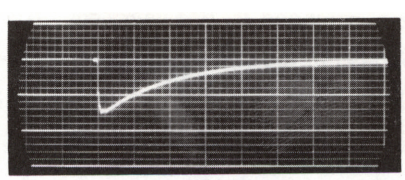
(*b*) At pin 1 of V6; oscilloscope at 50 v/cm



(*c*) At pin 1 of V4; oscilloscope at 1 v/cm



(*d*) At pin 3 of V3; oscilloscope at 0.5 v/cm



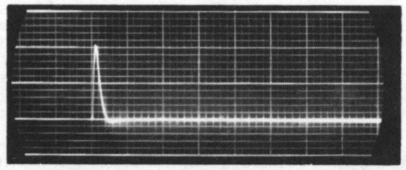
(*e*) At the amplifier input (input end of gain control); oscilloscope at 50 mv/cm; pulse generator at 175 mv

Fig. 2.74. A61 Amplifier Waveforms.

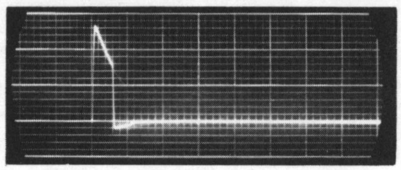
2.4.7 Overload Performance

With the following instrument settings, waveforms at various overloads are shown in Fig. 2.75.

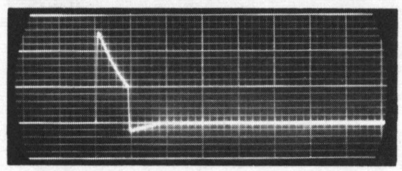
1. Pulse generator:	As needed.
2. Amplifier gain:	Maximum.
3. Observation point:	Main amplifier output.
4. Oscilloscope:	50 v/cm; 10 μsec/cm.



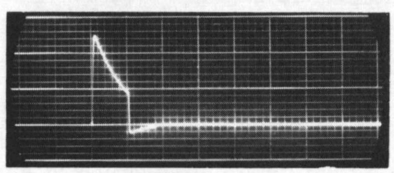
(*a*) 1X overload; generator output is 37 mv



(*b*) 10X



(*c*) 100X



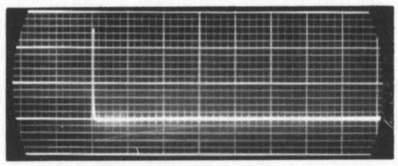
(*d*) 265X

Fig. 2.75. Waveforms Showing Overload Performance.

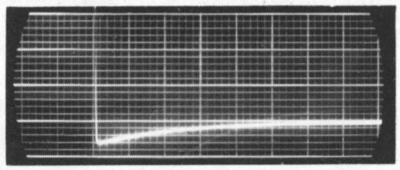
The waveforms of Fig. 2.76 were obtained with the same instrument settings as for Fig. 2.75 except that the probe was connected to the output of the cathode follower preceding the second differentiator (pin 1 of V6), and the sweep speed was changed to 100 μsec/cm. Although the output signal is of relatively short duration for large overloads, the preceding stage is blocked for several hundred microseconds.

With the probe connected to pin 1 of V4 and the oscilloscope set at 5 v/cm and 100 μsec/cm, the waveform shown in Fig. 2.77 was obtained for an overload of 100$\times$.

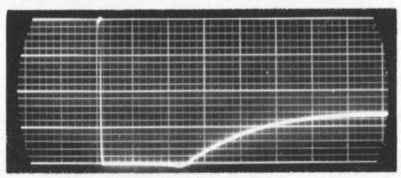
The probe was connected to the plate of the limiter stage (pin 5 of V2), and the oscilloscope was set at 5 v/cm and 5 μsec/cm. The waveform obtained for 100$\times$ overload is shown in Fig. 2.78.



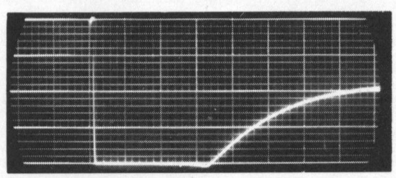
(*a*) 1X overload



(*b*) 10X



(*c*) 100X



(*d*) 265X (10 v input)

Fig. 2.76. Waveforms Showing Overload Performance.

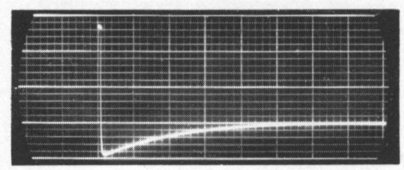


Fig. 2.77. Overload Waveform at Pin 1 of V4.

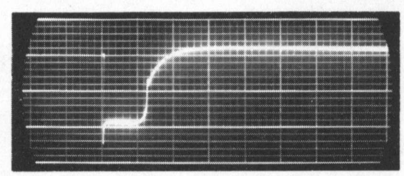
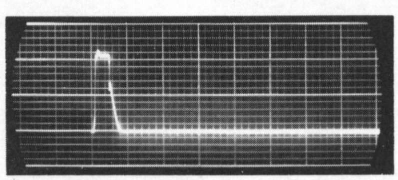


Fig. 2.78. Overload Waveform at Pin 5 of V2.

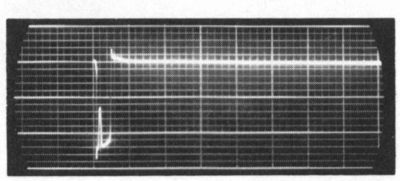
2.4.8 Discriminator Output Waveforms

Waveforms of the discriminator output with the following instrument settings are shown in Fig. 2.79.

1. Pulse generator: As needed to produce 100-v pulse from the amplifier.
2. Discriminator: Maximum.
3. Oscilloscope: 10 v/cm; 5 μsec/cm.



(*a*) Positive output



(*b*) Negative output

Fig. 2.79. Discriminator Output Waveforms.

2.4.9 Measurement of Feedback Factors

First Stage

The test setup shown in Fig. 2.73 is used to measure feedback factors.

Adjust the generator for an amplifier output signal of approximately 50 v. Note the generator dial reading and the output voltage. Ground pin 2 or 7 of V1 and reset the generator for the same output voltage; the ratio of generator readings is the feedback factor. A value of 25 is typical.

Output Group

Repeat the preceding procedure, but ground pin 7 of V4 through a 25-μf capacitor to measure the feedback factor. A value of 16 is typical.

2.4.10 Linearity

Use the setup shown in Fig. 2.73 and follow the procedure described in Sec 1.3.6, "Measurement of Integral Nonlinearity, Method I," except for the section concerned with the zero setting of the PHS. The lowest reading on the PHS of the A61 linear amplifier is 20 dial divisions.

The zero adjustment is made as follows: Set the PHS dial to 520 and the pulse generator dial to 500. Adjust for the half-triggering point (refer to Method I). Now set the PHS dial to 270 and adjust the generator to the half-triggering point. If the generator dial reads 250, the discriminator zero is correctly set. If not, set it according to Method I.

In making the linearity test by Method I, the half-triggering point in a typical amplifier should occur at a dial reading 20 $\pm$ 1 divisions higher than that of the generator, except in the region above 900 on the dial, where the diode clamp in the amplifier causes an error.

2.4.11 Regulation of Power Supply

The regulation afforded by the Sola transformer is such that the variation in B^+ with changes in line voltage is less than 0.025% in the range from 105 to 125 v.

2.4.12 Gain Dependence on Line Voltage

See the preceding paragraph.

2.4.13 Noise Measurements

Follow the procedure described in Sec 1.5.4. See the list of specifications for typical noise values.

2.4.14 Circuit Diagram

The circuit diagram for the A61 linear amplifier is given in Fig. 2.80.

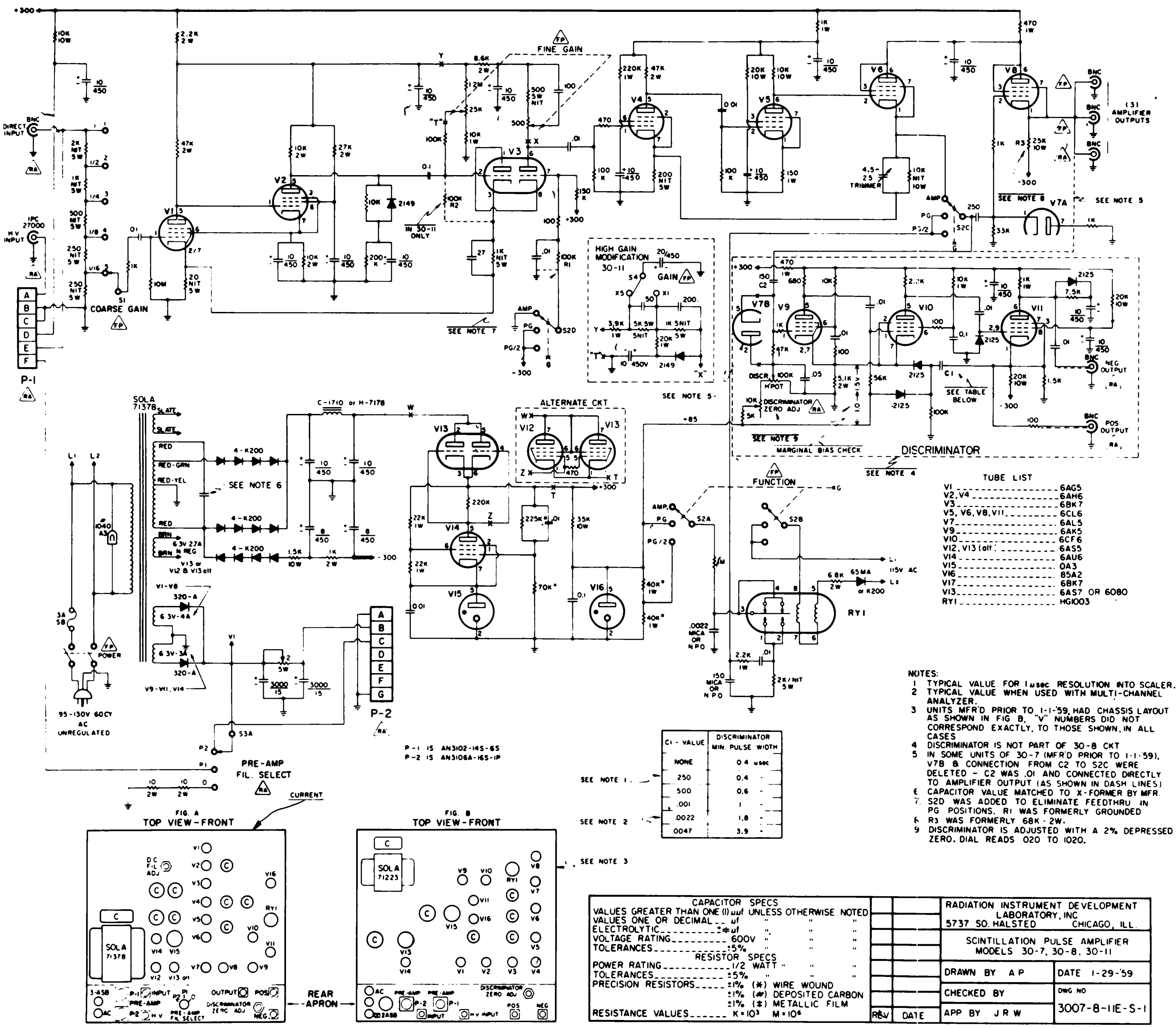


Fig. 2.80. Circuit Diagram for an A61 Linear Amplifier.

2.5 A1 LINEAR AMPLIFIER, MODEL Q-1151

2.5.1 General Description

This amplifier is similar to the A1D linear amplifier, model Q-1326A. It differs from the A1D in that a differential discriminator replaces the integral discriminator usually supplied. Other characteristics are similar. Refer to the section on the A1D amplifier for operating characteristics of the A1 amplifier, and to the section on the DD2 amplifier for operating characteristics of the differential discriminator.

2.5.2 Tube Complement

The A1 linear amplifier has the following tubes:

1 – 6SJ7	1 – 6AH6
3 – 6AC7	2 – 6AL5
2 – 6AG7	1 – 6AS7 or 6080
1 – 6197	1 – 6AU6
4 – 404A/5847	1 – 0A2
2 – 12AT7	1 – 5U4G

2.5.3 Circuit Diagram

The circuit diagram for the A1 linear amplifier is given in Fig. 2.81.

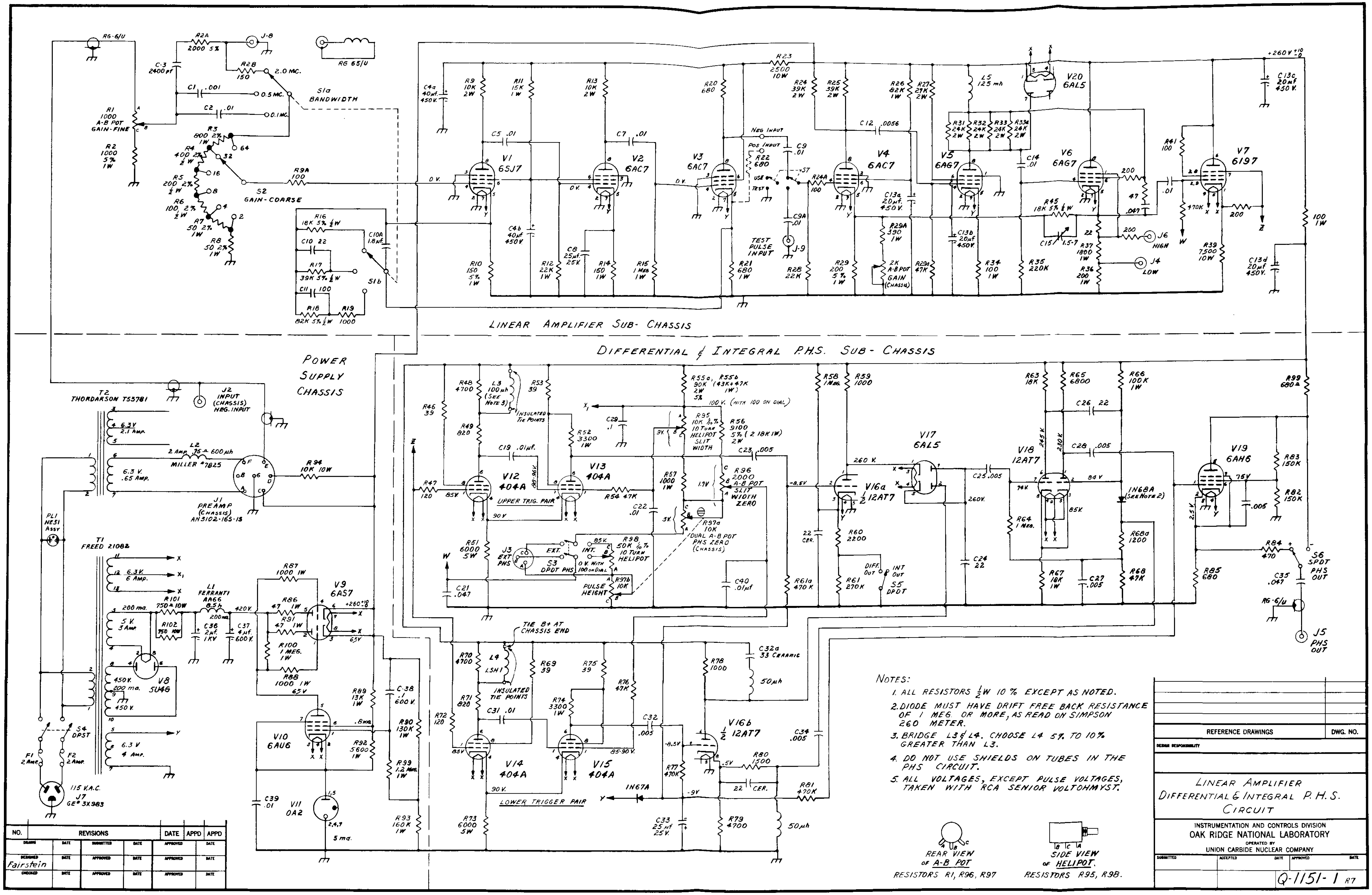


Fig. 2.81. A1 Linear Amplifier Differential and Integral Pulse-Height Selector: Circuit.

3. READING LIST

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