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INSTRUCTION MANUAL

435 ACTIVE FILTER AMPLIFIER

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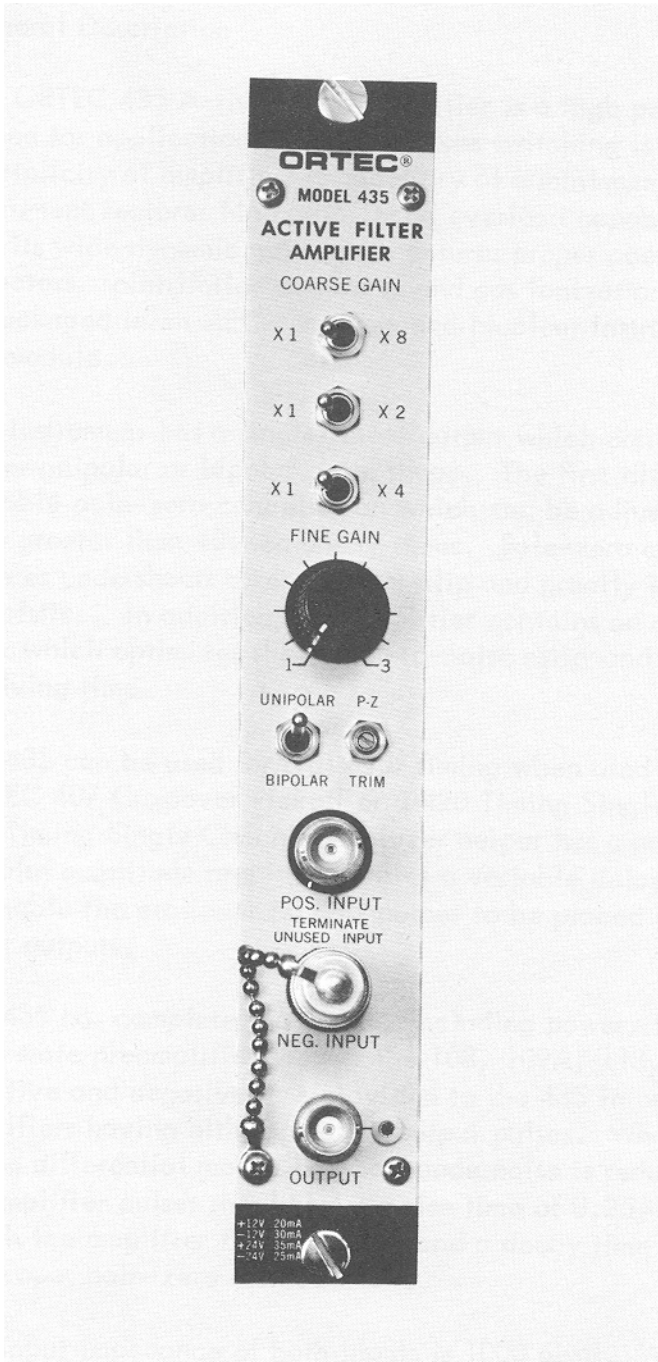
ORTEC®
MODEL 435
ACTIVE FILTER
AMPLIFIER
COARSE GAIN
X 1 X 8
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FINE GAIN
1 3
UNIPOLAR P-Z
BIPOLAR TRIM
POS. INPUT
TERMINATE UNUSED INPUT
NEG. INPUT
OUTPUT
+12V 20mA
−12V 30mA
+24V 35mA
−24V 25mA

ORTEC 435
ACTIVE FILTER AMPLIFIER

1. DESCRIPTION

1.1 General Description

The ORTEC 435 Active Filter Amplifier is a high performance amplifier designed for applications where bandpass switching is not required and where a multiplicity of amplifiers is necessary at a minimum of cost and space. The instrument features high gain, good overload capability, and high resolution; and its wide dynamic gain range ensures proper operation with semiconductor detectors, scintillation counters, and gas ionization chambers. The amplifier is packaged in an AEC-recommended Nuclear Instrument Module (NIM) standard module.

The instrument has a single linear output which can be switch selected for either unipolar or bipolar pulse shape. The first differentiation network has variable pole-zero cancellation which can be adjusted to match preamplifiers with greater than 40 μsec decay times. Pole-zero cancellation dramatically reduces undershoots after the first clip and greatly improves the overload characteristics. In addition, the amplifier contains an active filter shaping network which optimizes the signal-to-noise ratio and minimizes the overall resolving time.

The 435 can be used for crossover timing when used in conjunction with an ORTEC 407 Crossover Pickoff or a 420 Timing Single Channel Analyzer. The 420 Timing Single Channel Analyzer output has a minimum of walk as a function of pulse amplitude and incorporates a variable delay time on the output pulse to enable the crossover pickoff output to be placed in time coincidence with other outputs.

The 435 has complete provisions, including power, for operating any ORTEC solid state preamplifier such as the 108, 109A, 113, and 118A. Two inputs (positive and negative) are provided to the 435 in order to accommodate preamplifiers having either polarity output pulses. When the two inputs are used in the differential mode, common mode noise is reduced by a factor of 200. Preamplifier pulses should have a rise time of 0.25 μsec or less, to properly match the amplifier filter network and a decay time of greater than 40 μsec for proper pole-zero cancellation.

The input impedance of both inputs is 1000 ohms. When long preamplifier cables are used, the cables can be terminated in series at the preamplifier end or in shunt at the amplifier end with the proper resistance.

The output impedance of the 435 is about 0.5 ohm. The output can be connected to other equipment by either a single cable going to all equipment and shunt terminated at the far end (and series terminated at the amplifier if reflections are a problem) or separate cables to each instrument with each cable series terminated at the amplifier. (See Section 3.6.)

Gain changing is accomplished by constant impedance T attenuators. By using this technique, the bandwidth of the feedback amplifier stages involved in gain switching remains constant regardless of gain and therefore rise time changes with gain switching (which cause crossover walk) are limited to only small capacitance effects across the attenuators. A special low temperature coefficient constant impedance potentiometer is used as a Fine Gain control for the same reason.

The 435 is one NIM standard module wide. The unit has no self-contained power supply; power is obtained from a NIM standard Bin and Power Supply such as the ORTEC 401A/402A. The 435 design is consistent with other modules in the ORTEC 400 Series, i.e., it is not possible to overload the bin power supply with a full complement of modules in the bin. Since twelve 435 amplifiers can be contained and powered in one Bin and Power Supply, the 435 is particularly suited to experiments requiring many amplifiers.

1.2 Pole-Zero Cancellation

Pole-zero cancellation is a method for eliminating pulse undershoot after the first differentiating network. The technique employed is described by referring to the waveforms and equations shown in Figures 1-1 and 1-2. In a non-pole-zero cancelled amplifier, the exponential tail on the preamplifier output signal (usually 50 to 500μsec) causes an undershoot whose peak amplitude is roughly:

$$\frac{\text{Undershoot Amplitude}}{\text{Differentiated Pulse Amplitude}} = \frac{\text{Differentiation Time}}{\text{Preamplifier Pulse Decay Time}}$$

For a 1μsec differentiation time and a 50μsec preamplifier pulse decay time, the maximum undershoot is 2% and decays with a 50μsec time constant. Under overload conditions, this undershoot is often sufficiently large to saturate the amplifier during a considerable portion of the undershoot causing excessive deadtime. This effect can be reduced by increasing the preamplifier pulse decay time (which reduces the counting rate capabilities of the preamplifier) or compensating for the undershoot by using pole-zero cancellation.

Pole-zero cancellation is accomplished by the network shown in Fig. 1-2. The pole $(\frac{1}{S + 1/T_o})$ due to the preamplifier pulse decay time is cancelled by the zero $(S + K/R_2C_1)$ of the network. In effect, the dc path across the differentiation capacitor adds an attenuated replica of the preamplifier pulse to just cancel the negative undershoot of the clipping network.

Total pole-zero cancellation requires that the preamplifier output pulse decay time is a single exponential decay and matched to the pole-zero cancellation network. The variable pole-zero cancellation network allows accurate cancellation for all preamplifiers having 40µsec or greater decay times. The network is factory adjusted to 50µsec which is compatible with all ORTEC FET preamplifiers. Improper matching of the pole-zero cancellation network will degrade the overload performance and cause excessive pile-up distortion at medium counting rates. Improper matching causes either an undercompensation (undershoot is not eliminated) or an overcompensation (output after the main pulse does not return to the baseline and decays to the baseline with the preamplifier time constant). The pole-zero trim is accessible from the front panel of the 435 and can easily be adjusted by observing the baseline with a monoenergetic source or pulser having the same decay time as the preamplifier.

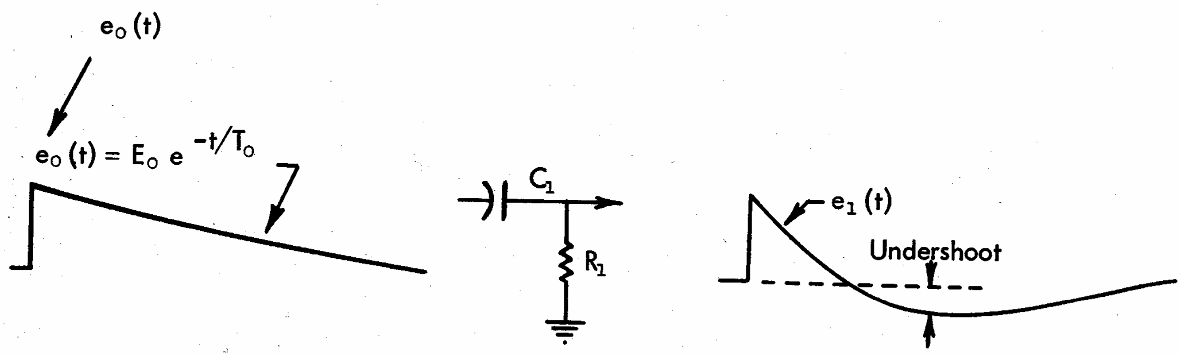
1.3 Active Filter

When only grid current and shot noise (gate current and drain thermal noise for an FET) are considered, the best signal-to-noise ratio occurs where the two noise contributions are equal for a given pulse shape. Also at this point, there is an optimum pulse shape for the optimum signal-to-noise ratio. Unfortunately, this shape (the Cusp shown in Fig. 1-3) is not physically realizable and very difficult to simulate. A pulse shape that can be simulated (the Gaussian in Fig. 1-3), requires a single RC differentiate and n equal RC integrates where n approaches infinity. The Laplace transform of this transfer function is:

$$G(S) = \frac{S}{(S + 1/RC)} \quad \frac{1}{(S + 1/RC)^n} \qquad n \rightarrow \infty$$

Where the first term is the single differentiate and the second term is the n integrates. The 435 Active Filter attempts to simulate this transfer function with the simplest possible circuit.

The 435 active filter circuit is shown in Fig. 1-4. The major attraction of the active RC filter is the simple synthesis of a complex pulse shape resulting in a significant reduction in size, complexity and cost. For a given resolving time



Preamplifier Output X First Amplifier Differentiation Network = Differentiated Pulse with Undershoot

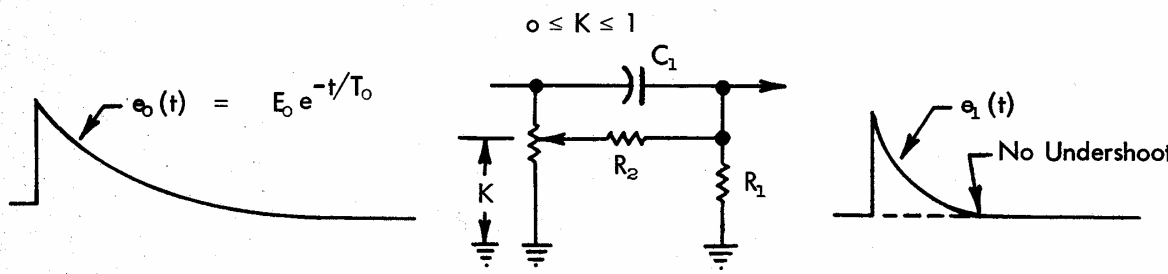
Equations:

$$E_o\, e^{-t/T_o} \quad \times \quad G(t) \quad = \quad e_1(t)$$

$$(E_o)\left(\frac{1}{S + 1/T_o}\right) \quad \times \left(\frac{S}{S + 1/R_1 C_1}\right) \quad = \quad e_1(S) \; ; \quad \text{Laplace Transform}$$

$$\frac{E_o}{T_o - T_1}\left[T_o\, e^{-t/T_1} - T_1\, e^{-t/T_o}\right] \quad = \quad e_1(t) \; ; \quad T_1 = R_1 C_1$$

Figure 1-1 Differentiation in a Non-Pole-Zero Cancelled Amplifier



Preamplifier Output	X	Pole Zero Cancelled Differentiation Network	=	Differentiated Pulse Without Undershoot

$$E_0 e^{-t/T_0} \quad \times \quad G(t) \quad = \quad e_1(t)$$

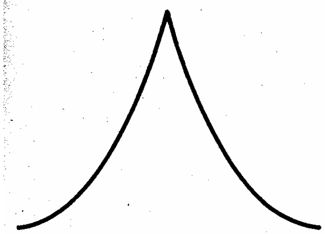
$$E_0 \frac{1}{\left(S + \frac{1}{T_0}\right)} \quad \times \quad \frac{S + \frac{K}{R_2 C_1}}{S + \frac{R_1 + R_2}{R_1 R_2 C_1}} \quad = \quad e_1(s)\text{ ; Laplace Transform}$$

Pole Zero cancel by letting $S + \frac{1}{T_0} = S + \frac{K}{R_2 C_1}$ or:

$$\frac{E_0}{S + \frac{R_1 + R_2}{R_1 R_2 C_1}} = \frac{E_0}{S + \frac{1}{R_p C_1}} = e_1(s)\text{ ; where } R_p = \frac{R_1 R_2}{R_1 + R_2}$$

$$E_0 e^{-t/R_p C_1} = e_1(t)$$

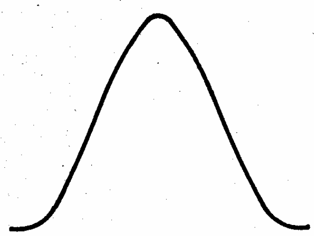
Figure 1-2 Differentiation in a Pole-Zero Cancelled Amplifier



CUSP

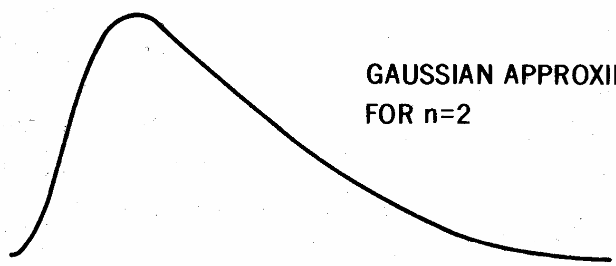
$e^{-t/RC}, t>o$

$e^{t/RC}, t<o$



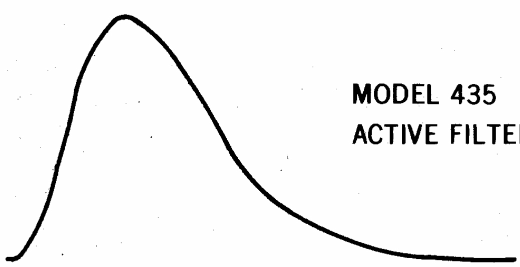
GAUSSIAN

$$\frac{S}{(S + 1/RC)} \; \frac{1}{(S + 1/RC)^n} \qquad n \rightarrow \infty$$



GAUSSIAN APPROXIMATION
FOR n=2

$$\frac{S}{(S + 1/RC)} \; \frac{1}{(S + 1/RC)^2}$$



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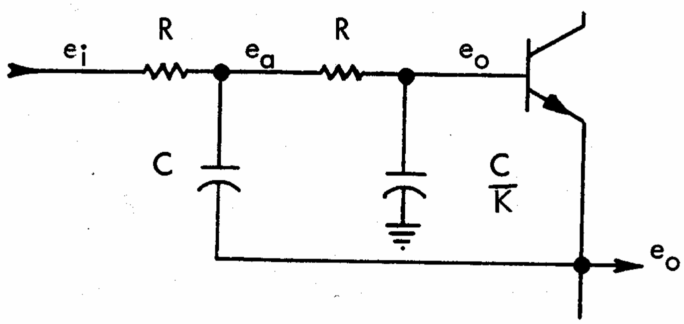
$$\frac{S}{(S + 1/RC)} \; \frac{1}{\left(S + \frac{1-j}{RC}\right)\left(S + \frac{1+j}{RC}\right)} \qquad j = \sqrt{-1}$$

Figure 1-3 Pulse Shapes for Good Signal-to-Noise Ratios



Equations:

$$e_o = e_a \frac{K/SC}{R + K/SC} = e_a \frac{K}{K + SRC}$$

$$\frac{e_i - e_a}{R} = \frac{e_a - e_o}{R} + \frac{e_a - e_o}{1/SC}$$

Eliminating e_a and solving for the transfer function:

$$\frac{e_o}{e_i} = \frac{K/R^2C^2}{S^2 + \frac{2S}{RC} + \frac{K}{R^2C^2}}$$

Figure 1-4 ORTEC 435 Active Filter

(RC), the time response of the filter network depends only on K (see the circuit equations in Fig. 1-4). For K = 1, the transfer function simplifies to:

$$\frac{e_o}{e_i} = \frac{1/R^2C^2}{S^2 + \frac{2S}{RC} + \frac{1}{R^2C^2}} = \frac{1/R^2C^2}{(S + 1/RC)^2}$$

which is an n = 2 approximation to the Gaussian pulse shape (see Fig. 1-3). For K = 4 (the actual case for the 435), the transfer function becomes:

$$\frac{e_o}{e_i} = \frac{4/R^2C^2}{S^2 + \frac{2S}{RC} + \frac{4}{R^2C^2}} = \frac{4/R^2C^2}{\frac{(S + 1 + j)}{RC}\ \frac{(S + 1 - j)}{RC}} \quad j = \sqrt{-1}$$

In this case, the complex roots cause an underdamped effect which reduces the resolving time and results in a more symmetrical pulse shape (see Fig. 1-3).

The 435 is manufactured with 1 μsec time constants (R = 1000Ω, C = 1000 pF, C/K = 500 pF). These time constants can be changed to suit the experiment. See Section 5.3 for details on this modification.

2. SPECIFICATIONS

2.1 General Specifications

2.1.1 The 435 is intended for use with a Nuclear Standard Bin such as the ORTEC 401A/402A. Two watts of power are required for the operation of the 435 in the quiescent condition. The 401A/402A can be operated on either 115 or 220 volts ac, 50-60 cps; if it is used with 220 volts ac, the manual for the 401A/402A must be referred to in order to ensure that correct connections have been made before operation on 220 volts ac is attempted. The instrument is supplied from the factory wired for 115 volts ac operation. The power input connector to the 401A/402A is a NEMA standard 3-wire grounding type.

Preamplifier power of ±12V and ±24V is available on the 435 rear panel connector, CN4, an Amphenol 17-10090. All signal inputs and outputs are on BNC connectors which are mounted on the front panel.

2.1.2 The instrument is intended for rack mounting in a 401A/402A Nuclear Standard Bin, but the Nuclear Standard Bin is suitably packaged for cabinet installation if desired. The weight of the 435 is 2 pounds and its outside dimensions are approximately 8.75 inches high by 1.35 inches wide by 9.75 inches deep.

2.2 Amplifier Specifications

INPUTS: Positive and negative, less than 0.25μsec rise time for best filter action, 40μsec minimum decay time for pole-zero cancellation, 12V maximum, 6V maximum to prevent saturation before differentiation, 1000 ohms input impedance both inputs.

OUTPUT: Positive or bipolar, 0-10V linear with 11.5V saturation into 1000 ohms, 0-9V linear with 10V saturation into 100 ohms, 0.5 ohm output impedance.

SHAPING: Active network filter resulting in approximately Gaussian shape, peak amplitude at 1.5μsec for unipolar and 1.1μsec for double clip, crossover at 2.5μsec for bipolar.

MAXIMUM GAIN: 1600

GAIN CHANGE RANGE: 192:1 total, 64:1 in factors of 2 by switches with 1% accuracy, 3:1 continuous fine control

LINEARITY: ±0.075% over specified linear range

NOISE: 10μV at maximum gain and single clip, 12μV at maximum gain and double clip, both referred to input

SHORT CIRCUIT LIMITS: The amplifier will sustain a direct short on the output for an indefinite period for counting rates up to 10^4 cps

COUNTING RATE: Less than 0.5% gain shift and 0.25% resolution spread at half maximum for a pulser peak above a 50K cu ts/sec ^{137}Cs background

OVERLOAD: Recovery within 2% of rated output from 1000 times overload in 2.5 non-overloaded pulse widths (20μsec) at maximum gain and specified input

COMMON MODE CANCELLATION: Better than 200:1 between inputs for positive input pulses only

CROSSOVER WALK: Less than 10 nsec crossover shift from maximum to minimum gain, or 50:1 change in pulse amplitude including ORTEC 420 Timing Single Channel Analyzer shift

OPERATING TEMPERATURE: 0 to 50°C

TEMPERATURE STABILITY: 0.01%/°C

POWER REQUIRED: 435 power supplied by 401A/402A Power Supply

DC input voltage	Quiescent current	Current with 50,000 pulses per second, each pulse 8V into 100 ohms
+24V	35 mA	35 mA
-24V	25 mA	25 mA
+12V	20 mA	26 mA
-12V	30 mA	36 mA

3. INSTALLATION

3.1 General Installation Considerations

The 435, used in conjunction with a 401A/402A Bin and Power Supply, is intended for rack mounting, and therefore it is necessary to ensure that vacuum tube equipment operating in the same rack with the 435 has sufficient cooling air circulating to prevent any localized heating of the all-semiconductor circuitry used throughout the 435. The temperature of equipment mounted in racks can easily exceed 120°F (50°C) unless precautions are taken.

3.2 Termination of Unused Input

The 435 is supplied with a 100 ohm attenuator. This attenuator should be connected to the unused input for proper gain calibration and best signal- to-noise ratio.

3.3 Connection to Preamplifier

The preamplifier output signal can be connected to the 435 via BNC connector CN1 labeled POS INPUT or BNC connector CN2 labeled NEG INPUT, depending on the polarity of the input pulse. The input impedance seen at either input is 1000 ohms and both inputs are dc coupled to ground; therefore, the output of the preamplifier must be either ac coupled or have zero dc voltage under no signal conditions.

The 435 incorporates pole-zero cancellation in order to enhance the overload characteristics of the amplifier. This technique requires matching the network to the preamplifier decay time constant in order to achieve perfect compensation. The network is variable and factory adjusted to 50µsec to match all ORTEC FET preamplifiers. If other preamplifiers or more careful matching is desired, the trim is accessible from the front panel. Adjustment is easily accomplished by using a monoenergetic source and observing the amplifier baseline after each pulse under overload conditions.

Preamplifier power of +24V and -24V is available on the preamp power connector, CN4.

When using the 435 with a remotely located preamplifier (i.e., preamplifier-to-amplifier connection through 25 feet or more of coaxial cable), care must be taken to ensure that the characteristic impedance of the transmission line from the preamplifier output to the 435 input is matched. Since the input impedance of the 435 is 1000 ohms, sending end termination will normally be preferred; i.e., the transmission line should be series terminated at the output of the preamplifier. All ORTEC preamplifiers contain series terminations which are either 91 ohms or variable.

Both inputs of the 435 can be used simultaneously to reduce common mode noise picked up by long cables passing noise generating areas. In this mode of operation, the preamplifier signal is connected to the POS INPUT and a separate identical cable in intimate contact with the first cable (the use of Twinax cable is preferable) is connected from the preamplifier ground to the NEG INPUT. In order to balance the noise concellation, it is sometimes necessary to insert a small variable resistor between the actual preamplifier ground and the center conductor of the second (ground signal) cable.

3.4 Connection of Test Pulse Generator

3.4.1 Connection of Pulse Generator to the 435 Through a Preamplifier

The satisfactory connection of a test pulse generator such as the ORTEC 419 or equivalent depends primarily on two considerations: (1) the preamplifier must be properly connected to the 435 as discussed in Section 3.2, and (2) the proper input signal simulation must be applied to the preamplifier. To ensure proper input signal simulation, refer to the instruction manual for the particular preamplifier being used.

3.4.2 Direct Connection of Pulse Generator to the 435

Since both inputs of the 435 have 1000 ohms input impedance, the test pulse generator will normally have to be terminated at the amplifier input with an additional shunt resistor. In addition, if the test pulse generator has a dc offset, a large series isolating capacitor is also required since the inputs of the 435 are dc coupled to the first amplifier stage. The ORTEC 204 or 419 Test Pulse Generators are designed for direct connection. When either of these units is used, they should be terminated with a 100 ohm terminator at the amplifier input. (The small error due to the finite input impedance of the amplifier can normally be neglected.)

3.4.3 Special Test Pulse Generator Considerations for Pole-Zero Cancellation

The pole-zero cancellation network in the 435 is factory adjusted for a 50μsec decay time to match ORTEC FET preamplifiers. When a tail pulser (such as the 204 or 419) is connected directly to one of the amplifier inputs, the pulser should be modified to obtain a 50μsec decay time if overload tests are to be made (other tests are not affected). See Section 6.2 for the details on this modification.

If a preamplifier is used and a tail pulser connected to the preamplifier pulser input, similar precautions are necessary. In this case, the effect of the pulser decay must be removed, i.e., a step input should be simulated. Details for this modification are also given in Section 6.2.

3.5 Connection to Power — Nuclear Standard Bin, ORTEC 401A/402A

The 435 contains no internal power supply and therefore must obtain power from a Nuclear Standard Bin and Power Supply such as the 401A/402A. It is reccommended that the bin power supply be turned off when inserting or removing modules. The ORTEC 400 Series is designed so that it is not possible to overload the bin power supply with a full complement of modules in the Bin; however, this may not be true when the Bin contains modules other than those of ORTEC design, and in this case, the power supply voltages should be checked after insertion of the modules. The 401A/402A has test points on the power supply control panel to monitor the dc voltages.

3.6 Output Connections and Terminating Considerations

The 435 linear output can be switch selected for either a unipolar or a bipolar output. The unipolar output should be used for high resolution spectrometry applications with semiconductor detectors. The bipolar output should be used in applications requiring high counting rates or crossover timing. Typical system block diagrams for a variety of experiments are described in Section 4.

The source impedance of the 0-10 volt standard linear outputs of most 400 Series modules is approximately 1 ohm. Interconnection of linear signals is, thus, non-critical since the input impedance of circuits to be driven is not important in determining the actual signal span, e.g., 0-10 volts, delivered to the following circuit. Paralleling several loads on a single output is therefore permissible while preserving the 0-10 volt signal span. Short lengths of interconnecting coaxial cable (up to approximately 4 feet) need not be terminated. However, if a cable longer than approximately 4 feet is necessary on a linear output, it should be terminated in a resistive load equal to the cable impedance. Since the output impedance is not purely resistive, and is slightly different for each individual module, when a certain given length of coaxial cable is connected and is not terminated in the characteristic impedance of the cable, oscillations will occasionally be observed. These oscillations can be suppressed for any length of cable by properly terminating the cable, either in series at the sending end or in shunt at the receiving end of the line. To properly terminate the cable at the receiving end, it may be necessary to consider the input impedance of the driven circuit, choosing an additional parallel resistor to make the combination produce the desired termination resistance. Series terminating the cable at the sending end may be preferable in some cases where receiving end terminating is not desirable or possible. When series terminating at the sending end, full signal span, i.e., amplitude, is obtained at the receiving end only when it is essentially unloaded or loaded with an impedance many times that of the cable. This may be accomplished by inserting a series resistor equal to the characteristic impedance of the cable internally in the module between the actual amplifier output on the etched board and the output connector. It must be remembered that this impedance is in series with the input impedance of the load being driven, and in the case where the driven load is 900 ohms, a decrease in the signal span of approximately 10% will occur for a 93-ohm transmission line. A more serious loss occurs when the driven load is 93 ohms and the transmission system is 93 ohms. In this case, a 50% loss will occur. BNC connectors with internal terminators are available from a number of connector manufacturers in nominal values of 50, 100, and 1000 ohms. ORTEC stocks in limited quantity both the 50 and 100 ohm BNC terminators. The BNC terminators are quite convenient to use in conjunction with a BNC tee.

3.7 Amplifier Gain and Noise Considerations

Under normal operating conditions, the signal-to-noise ratio will be determined by detector and preamplifier noise and will be constant for any gain setting. However, in some circumstances, the amplifier may contribute additional noise to the overall system. If this situation exists, the best signal-to-noise performance will be obtained at maximum gain and the worst condition will be when either the X8 or X2 or both gain switches are in the X1 position. The signal-to-noise ratio is only slightly affected by the X4 gain switch position and the Fine Gain control.

3.8 Shorting the Amplifier Output

The output of the 435 is ac coupled with an output impedance of about 0.5 ohm. If the output is shorted with a direct short-circuit and the amplifier counting rate exceeds 10,000 counts per second, the output stage will eventually heat up sufficiently to destroy itself (about one minute for 10^5 cps). The amplifier output may be shorted indefinitely without catastrophic damage at rates below 10^4 cps.

4. OPERATING INSTRUCTIONS

4.1 Controls and Connectors

4.1.1 Description

FINE GAIN: Fine gain control is provided over a range of 3 to 1. The gain change is not very linear with rotation since the control was designed to provide constant impedance attenuation to reduce shape changes with gain rather than linear gain change.

COARSE GAIN: Coarse gain control is provided by three individually switchable, constant impedance "T" attenuators. The gain factors provided by each of the sections are X2, X4, and X8. A better signal-to-noise ratio is obtained when the X2 and X8 switches are in the X2 and X8 position and the X4 and Fine Gain are used for gain reduction.

UNIPOLAR-BIPOLAR: Either unipolar or bipolar output pulses are switch selectable. The gains are matched in both modes to 2.5%.

PZ TRIM: A trim potentiometer is provided to adjust the pole-zero cancellation network for varying preamplifier decay times.

POS. INPUT: The input for preamplifier pulses having a positive polarity. Preamplifier pulses should have less than 0.25μsec rise time and a 50μsec decay time. When this input is not used, it should be terminated with a 100 ohm terminator which is supplied with the amplifier.

NEG. INPUT: The input for preamplifier pulses having a negative polarity. When this input is not used, it should be terminated with a 100 ohm terminator which is supplied with the amplifier.

OUTPUT: A single, low impedance shaped output with a test probe adjacent for oscilloscope monitoring.

4.1.2 Front and Rear Panel Connector Data

Connector See Dwg 435-0101-S1	Generic Designation	Test Point	Output or Input Impedance	Shape and Amplitude Limitations
CN-1	POS. INPUT	No	1000 Ω	Positive only, less than 0.25μsec rise time, 50μsec decay time, 6V maximum linear, 12V maximum

Front and Rear Panel Connector Data (Cont'd)

Connector See Dwg 435-0101-S1	Generic Designation	Test Point	Output or Input Impedance	Shape and Amplitude Limitations
CN-2	NEG INPUT	No	1000 Ω	Same as above but negative only
CN-3	OUTPUT	TP-1	0.5 Ω	Positive or bipolar, 0-10V linear with 11.5V saturation into 1000 Ω, 0-9V linear with 10V saturation into 100 Ω, approximate Gaussian shape
CN-4	PREAMP POWER	No	dc	Pin 1: Gnd Pin 6: -24V Pin 7: +24V Pin 9: -12V Pin 4: +12V

4.2 Initial Testing and Observation of Pulse Waveforms

Refer to Section 6 for information on testing performance and observing waveforms at front panel test points.

4.3 General Considerations for Operation with Semiconductor Detectors

4.3.1 Calibration of Test Pulser

The ORTEC 419 Mercury Pulser, or equivalent, may easily be calibrated so that the maximum pulse height dial reading (1000 divisions) is equivalent to a 10-MeV loss in a semiconductor radiation detector. The procedure is as follows:

(1) Connect the detector to be used to the spectrometer system; i.e., preamp, main amplifier, and biased amplifier.

(2) Allow particles from a source of known energy (α-particles, for example) to fall on the detector.

(3) Adjust the amplifier gains and the bias level of the biased amplifier to give a suitable output pulse.

(4) Set the pulser PULSE HEIGHT potentiometer at the energy of the α-particles striking the detector (e.g., for a 5.1 MeV α-particle, set the dial at 510 divisions).

(5) Turn on the Pulser, use the NORMALIZE potentiometer and attenuators to set the output due to the pulser to the same pulse height as the pulse obtained in (3) above.

(6) The pulser is now calibrated; the dial reads in MeV if the number of dial divisions is divided by 100.

4.3.2 Amplifier Noise and Resolution Measurements

As shown in Figure 4-1, the preamplifier, amplifier, pulse generator, oscilloscope , and a wide-band rms voltmeter such as the Hewlett-Packard 400D are required for this measurement. Connect a suitable capacitor to the input to simulate the detector capacitance desired. To obtain the resolution spread due to amplifier noise:

(1) Measure the rms noise voltage (E_{rms}) at the amplifier output.

(2) Turn on the 419 mercury relay pulse generator and adjust the pulser output to any convenient readable voltage, E_o, as determined by the oscilloscope.

(3) The full width at half maximum (fwhm) resolution spread due to amplifier noise is then

$$N\ (\text{fwhm}) = \frac{2.660\ E_{rms} E_{dial}}{E_o}$$

where E_{dial} is the pulser dial reading in MeV and the factor 2.660 is the correction factor for rms to fwhm (2.35) and noise to rms meter correction (1.13) for average-indicating voltmeters such as the Hewlett-Packard 400D. A true rms voltmeter does not require the latter correction factor.

The resolution spread will depend upon the total input capacitance, since the capacitance degrades the signal-to-noise ratio much faster than the noise. A typical resolution spread versus external input capacitance for the 118A-435 system is shown in Fig. 4-2.

4.3.3 Detector Noise Resolution Measurements

The same measurement described in Section 4.3.2 can be made with a biased detector instead of the external capacitor used to simulate the detector capacitance. The resolution spread will be larger because the detector contributes both noise and capacitance to the input. The detector noise resolution spread can be isolated from the amplifier noise spread if the detector capacity is known, since

$$N_{det}^2 + N_{amp}^2 = N_{total}^2$$

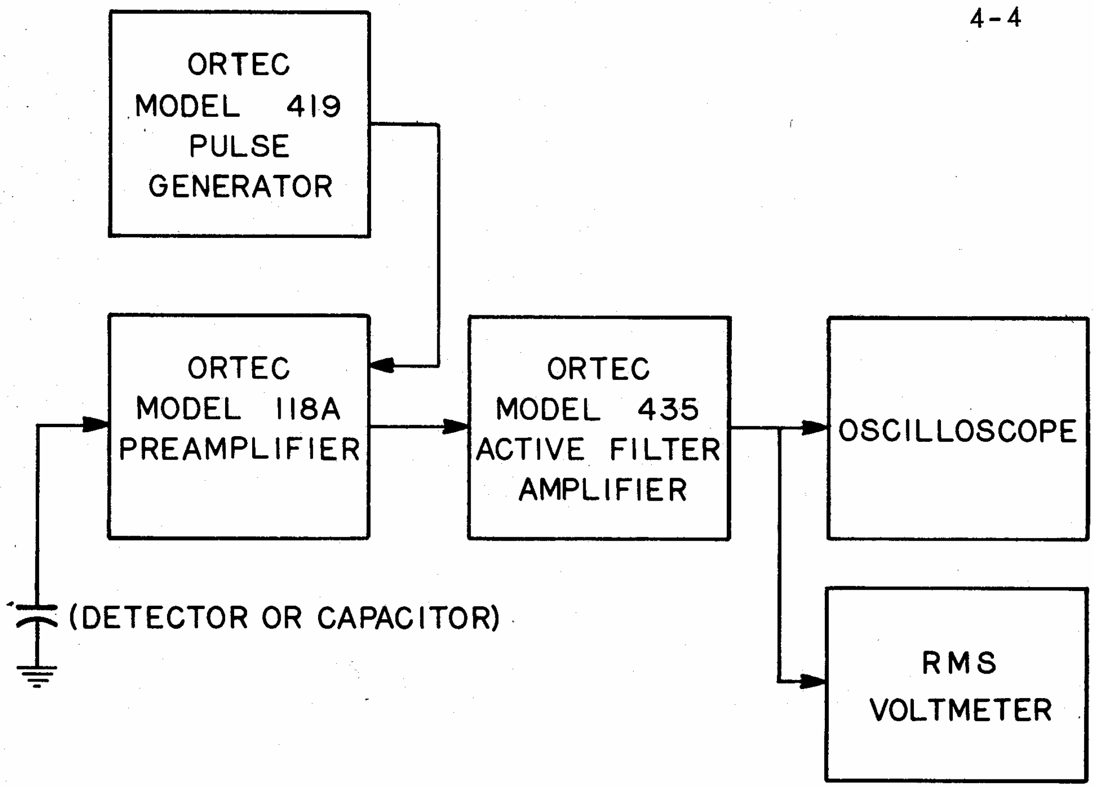


Figure 4-1 Measuring Amplifier and Detector Noise Resolution

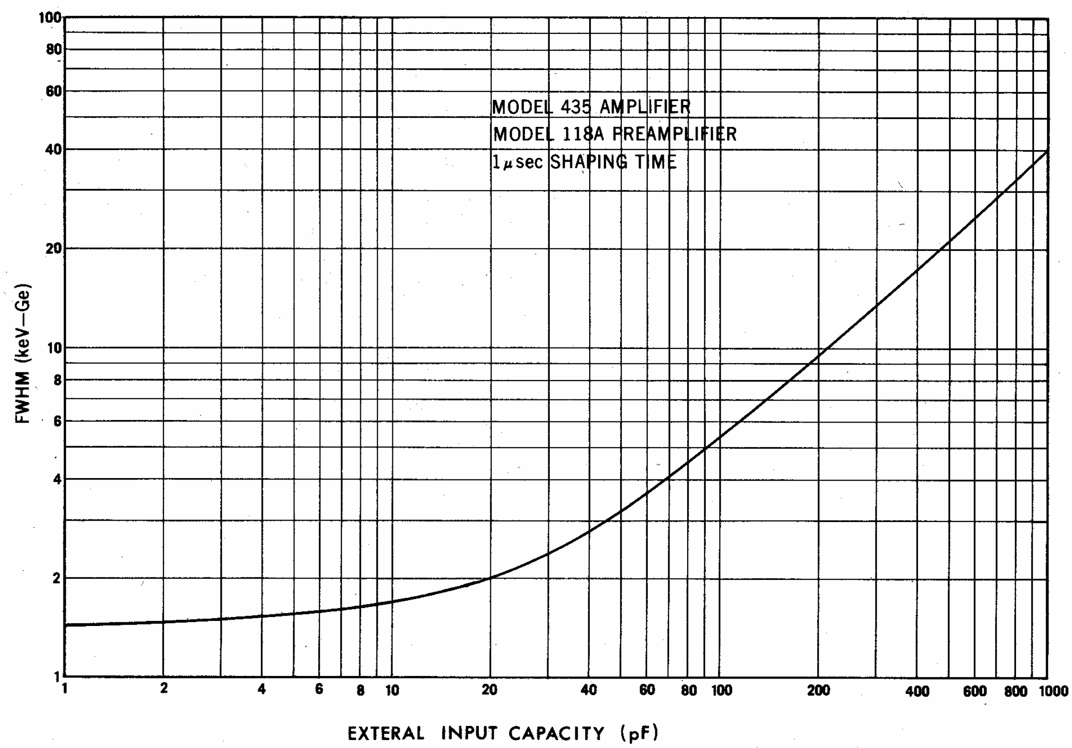


Figure 4-2 Resolution Spread Versus External Input Capacity

where N_{total} is the total resolution spread and N_{amp} is the amplifier resolution spread with the detector replaced by its equivalent capacitance.

The detector noise tends to increase with bias voltage, but the detector capacitance decreases, thus reducing the resolution spread. The overall resolution spread will depend upon which effect is dominant. Figure 4-3 shows curves of typical total noise resolution spread versus bias voltage, using the data from several ORTEC silicon semiconductor radiation detectors.

4.3.4 Amplifier Noise and Resolution Measurements Using a Pulse Height Analyzer

Probably the most convenient method of making resolution measurements is with a pulse height analyzer as shown by the set-up illustrated Figure 4-4.

The amplifier noise resolution spread can be measured directly with a pulse height analyzer and the mercury pulser as follows:

(1) Select the energy of interest with an ORTEC 419 Pulse Generator, and set the Active Filter Amplifier and Biased Amplifier GAIN and BIAS LEVEL controls so that the energy is in a convenient channel of the analyzer.

(2) Calibrate the analyzer in keV per channel, using the pulser (full scale on the pulser dial is 10 MeV when calibrated as described in Section 4.3.1).

(3) The amplifier noise resolution spread can then be obtained by measuring the full width at half maximum of the pulser spectrum.

The detector noise resolution spread for a given detector bias can be determined in the same manner by connecting a detector to the preamplifier input. The amplifier noise resolution spread must be subtracted as described in Section 4.3.3. The detector noise will vary with detector size, bias conditions, and possibly with ambient conditions.

4.3.5 Current-Voltage Measurements for Silicon and Germanium Detectors

The amplifier system is not directly involved in semiconductor detector current-voltage measurements, but the amplifier serves well to permit noise monitoring during the measurements. The detector noise measurement is a more sensitive method of determining the maximum detector voltage which should be used, because the noise increases

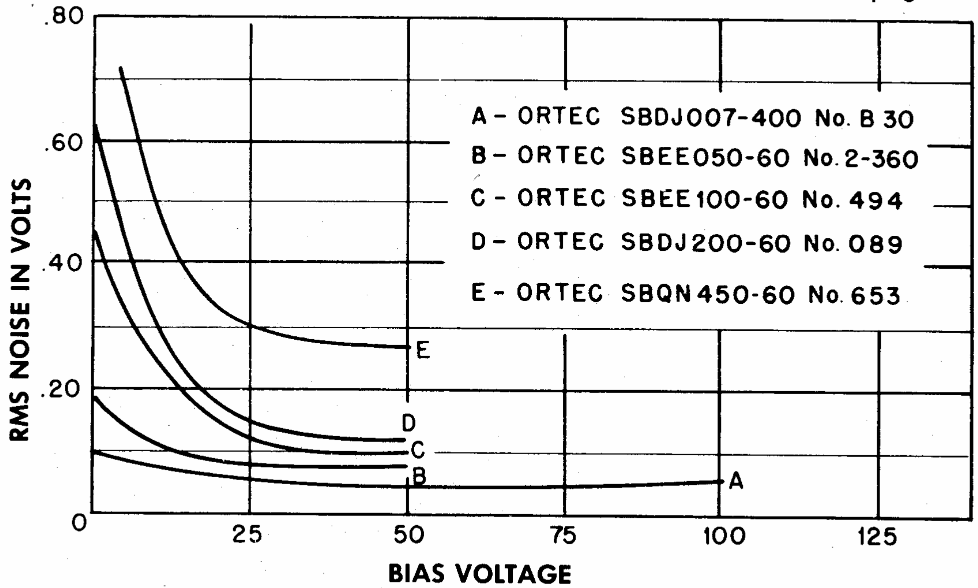


Figure 4-3 Amplifier and Detector Noise Versus Bias Voltage

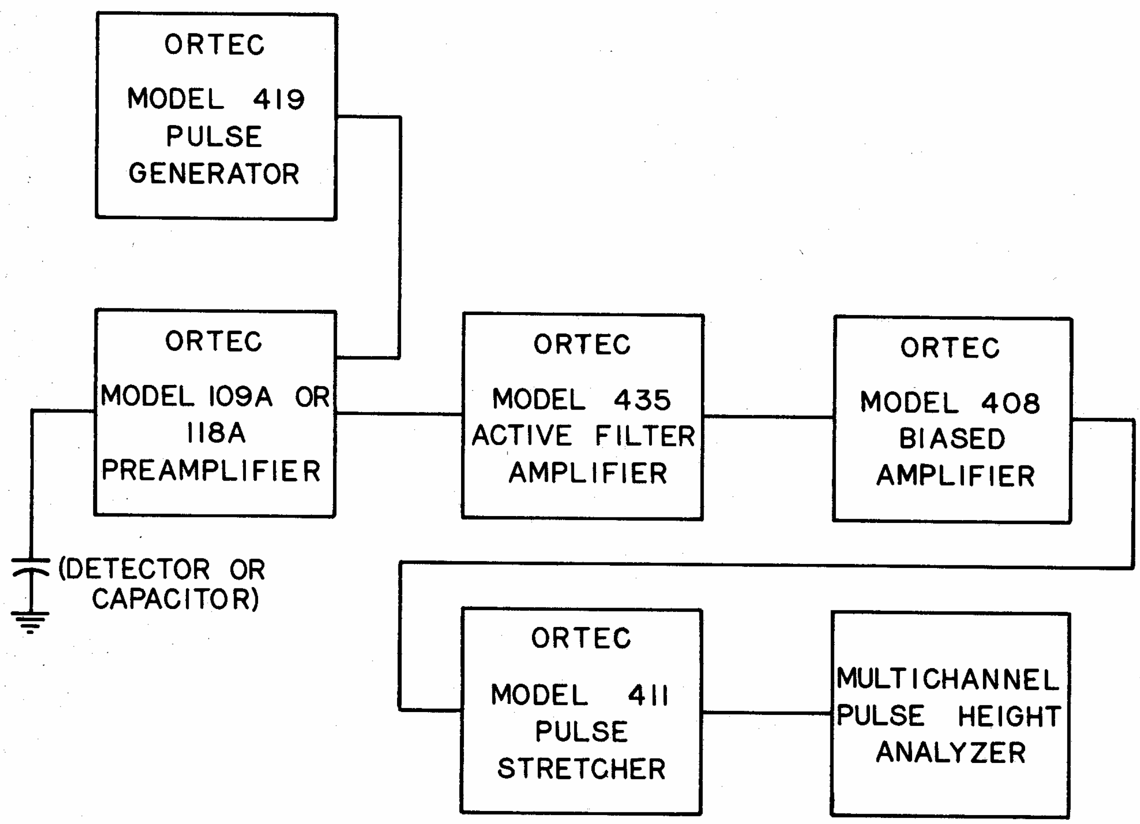


Figure 4-4 Measuring Resolution with a Pulse Height Analyzer

more rapidly than the reverse current at the onset of detector breakdown.

Figure 4-5 shows the setup required for current-voltage measurements. The ORTEC 428 Bias Supply is used as the voltage source. Bias voltage should be applied slowly and reduced when noise increases rapidly as a function of applied bias. Figure 4-6 shows several typical current-voltage curves for ORTEC silicon detectors.

When it is possible to float the microammeter at the detector bias voltage, the alternate method of detector current measurement shown by the dashed lines in Figure 4-6 is preferable. The detector is grounded as in normal operation and the microammeter is connected to the current monitoring jack on the 428 Detector Bias Supply.

4.3.6 Recommended Method for Preamp-Main Amp Gain Adjustments as a Function of Input Particle Energy

With the input energy at a constant, or maximum, known value, the total system gain of the preamp and main amplifier can be adjusted to an optimum value by utilizing the following general considerations:

(1) The primary design criterion for the preamp is best signal-to-noise ratio at the output; therefore, the preamp should be operated with the gain switch in its maximum gain position. This will result in the best signal-to-noise ratio available, and at the same time the absolute voltage amplitude of the preamp signal will be maximized.

(2) Under normal operating conditions, the signal-to-noise ratio will be determined by detector and preamplifier noise and will be constant for any gain setting. However, in some circumstances, the amplifier may contribute additional noise to the overall system. If this situation exists, the best signal-to-noise performance will be obtained at maximum gain and the worst condition will be when either the X8 or X2 or both gain switches are in the X1 position. The signal-to-noise ratio is only slightly affected by the X4 gain switch position and the Fine Gain control.

4.4 Operation in Spectroscopy Systems

4.4.1 High-Resolution Alpha-Particle Spectroscopy System

The block diagram of a high resolution spectroscopy system for measuring natural alpha-particle is shown in Fig. 4-7. Since natural alpha-particle radiation only occurs above several MeV, an ORTEC 408 Biased Amplifier and ORTEC 411 Stretcher are used to suppress the unused portion of the spectrum. The

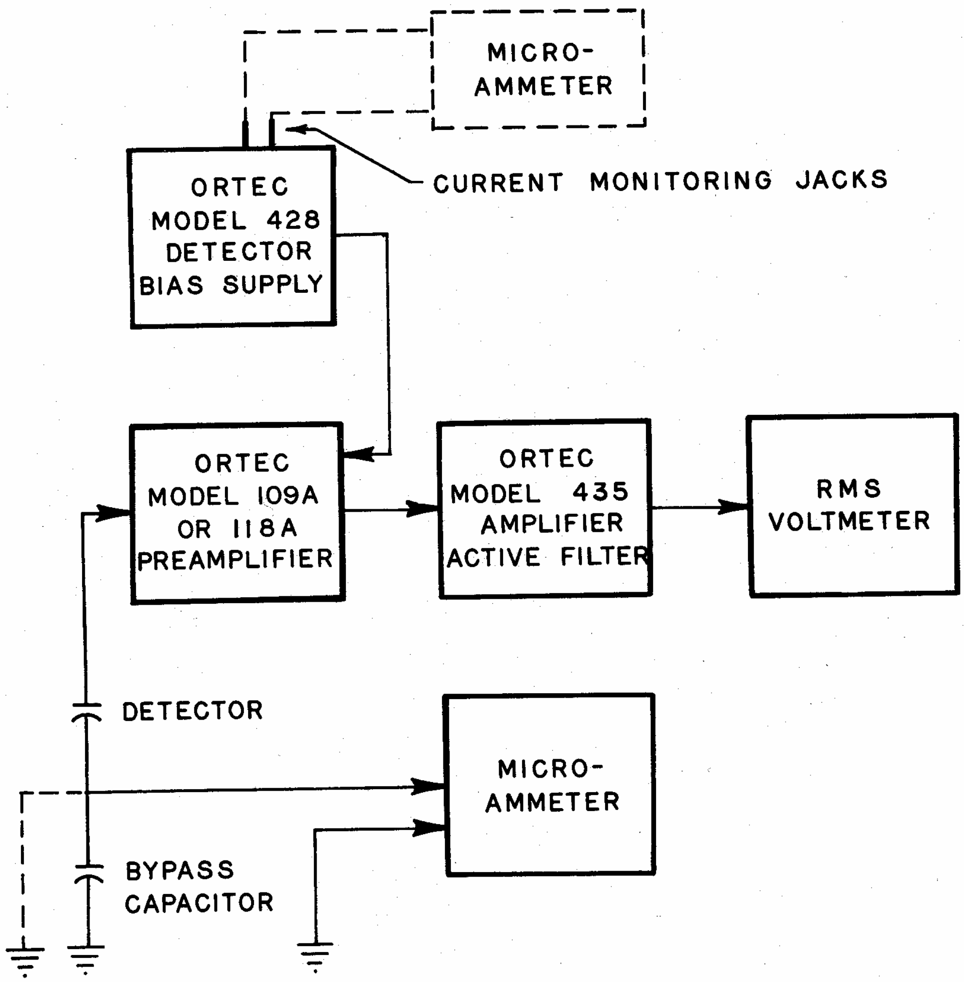


Figure 4-5 Measuring Detector Current-Voltage Characteristics

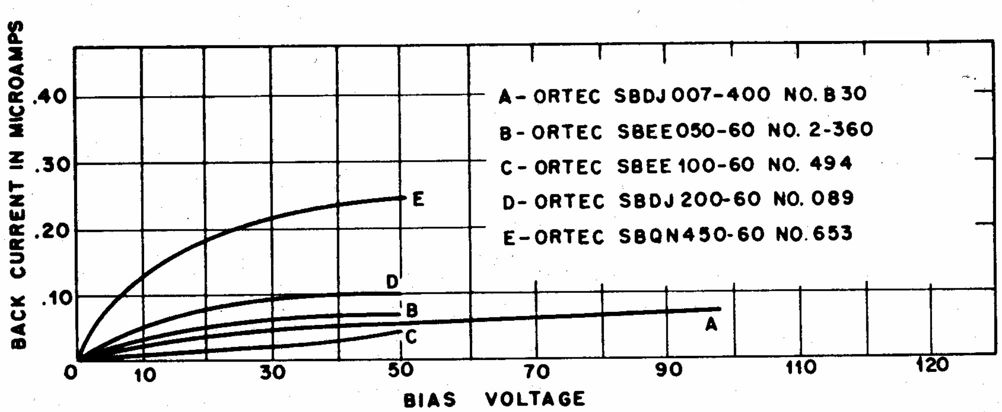


Figure 4-6 Silicon Detector Back Current Versus Bias Voltage

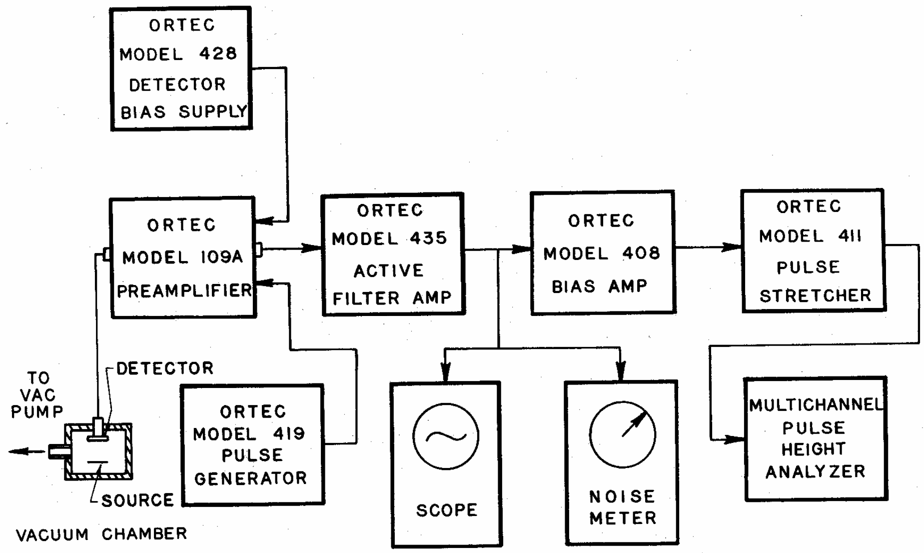


Figure 4-7 High Resolution Alpha Particle Spectroscopy System

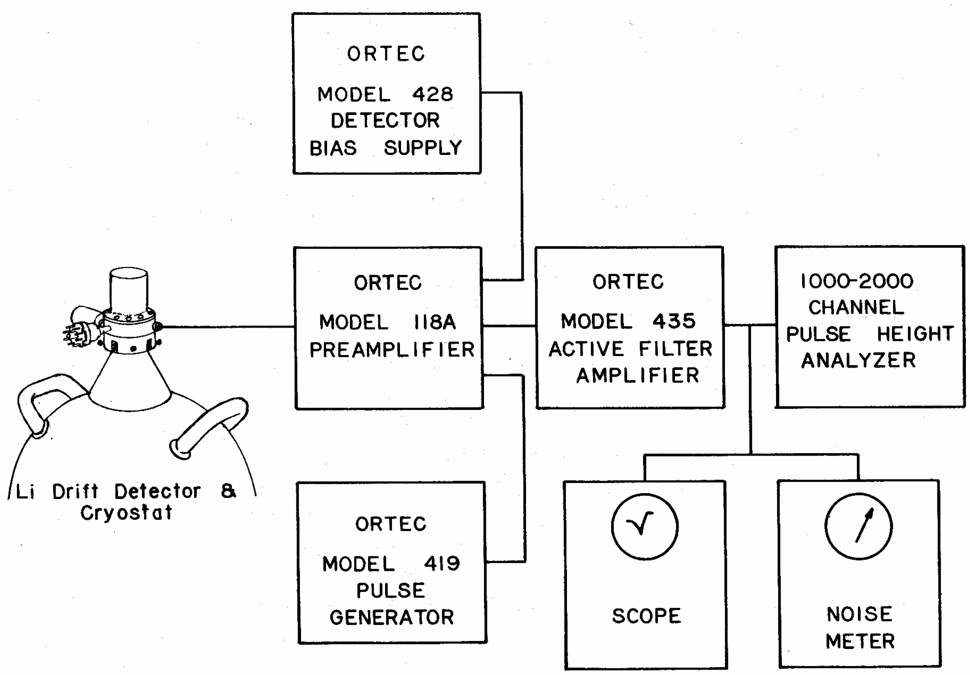


Figure 4-8 High Resolution Gamma Spectroscopy System Using a Lithium Drifted Germanium Detector

411 is used to shape the output pulses after biasing to avoid pulse height analyzer nonlinearities.

Alpha particle resolution is obtained in the following manner:

(1) Using maximum preamplifier gain, meduim amplifier gain, and minimum biased amplifier gain and bias level, accumulate the alpha peak in the multichannel analyzer.

(2) Slowly increase the bias level and biased amplifier gain until the alpha peak is spread over 5 to 10 channels and the minimum to maximum energy range desired corresponds to the first and last channels of the analyzer.

(3) Calibrate the analyzer in keV per channel using the pulser and the known energy of the alpha peak (see Section 4.3.1).

(4) The resolution can be obtained by measuring the full width at half maximum (fwhm) of the alpha peak in channels and converting to keV.

4.4.2 High Resolution Gamma Spectroscopy System

A high resolution gamma system block diagram is shown in Figure 4-8. Although a biased amplifier is not shown (a larger channel analyzer being preferred), it can be used if only a smaller channel analyzer is available and only higher energies are of interest.

When using lithium drifted germanium detectors cooled by a liquid nitrogen cryostat, it is possible to obtain resolutions from about 1 keV fwhm up (depending on the energy of the incident radiation and the size and quality of the detector). Reasonable care is required to obtain such results. Some guide lines for obtaining optimum resolution are:

(1) Keep interconnection capacities between the detector and preamplifier to an absolute minimum (no cables).

(2) Keep humidity low near the detector-preamplifier junction.

(3) Operate in amplifier and preamplifier gain regions which provide the best signal-to-noise ratio.

(4) Operate at the highest allowable detector bias to keep the input capacity low.

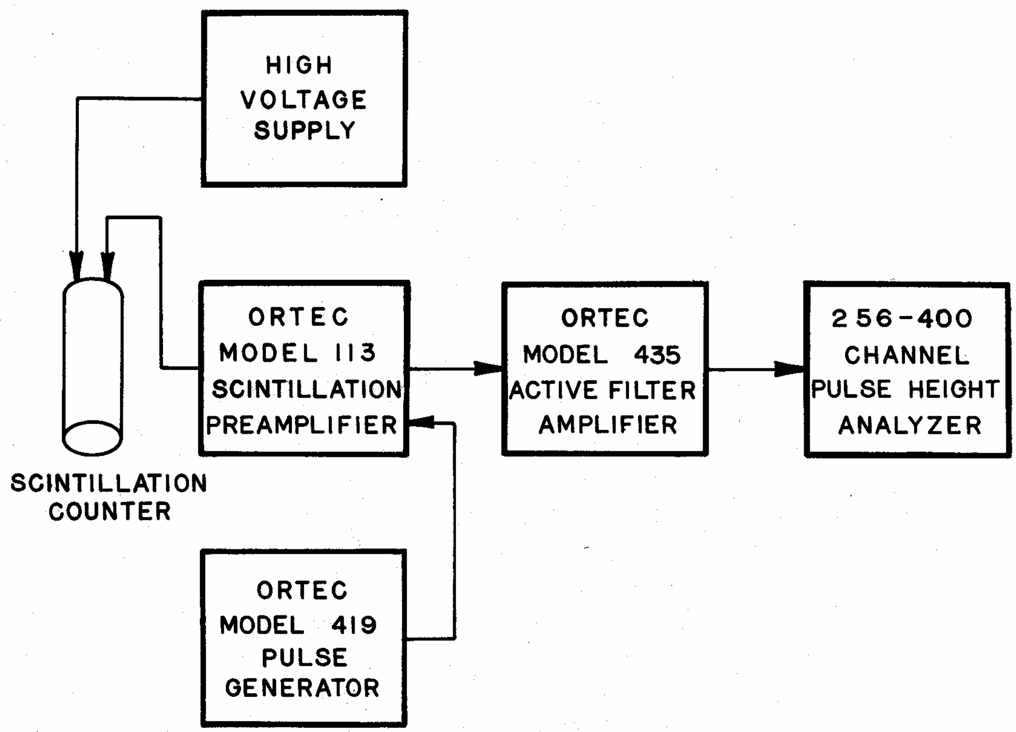


Figure 4-9 Scintillation Counter Gamma Spectroscopy System

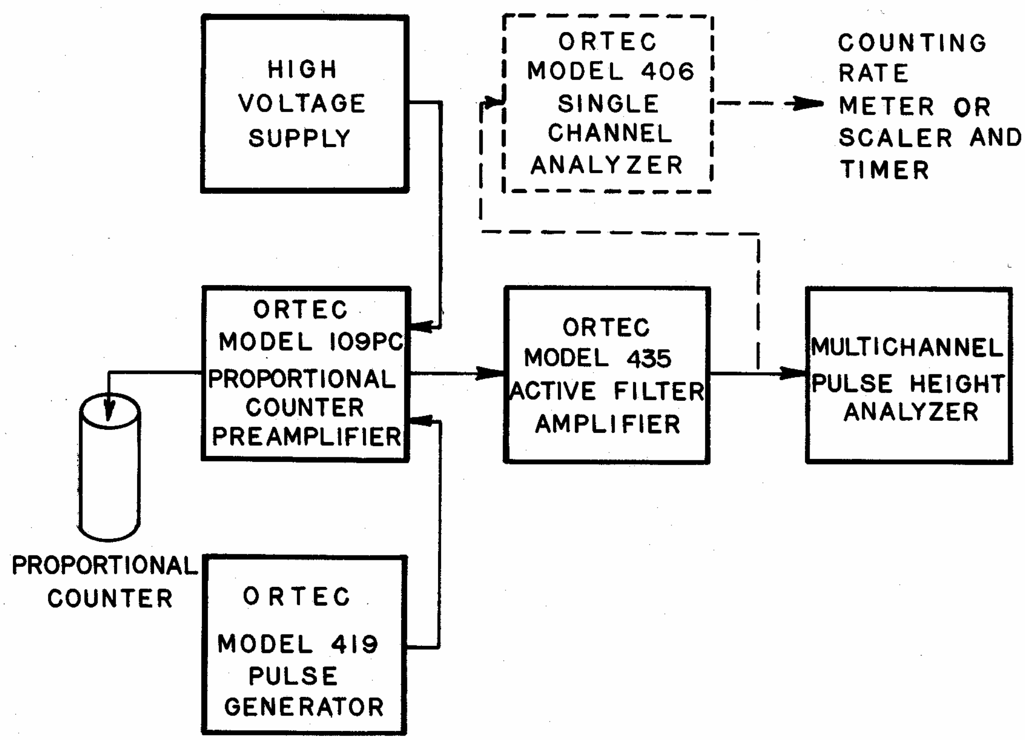


Figure 4-10 High Resolution X-Ray Spectroscopy System

4.4.3 Scintillation Counter Gamma Spectroscopy Systems

The 435 can be used in scintillation counter spectroscopy systems as shown in Fig. 4-9. The amplifier clipping time constants are proper for NaI or plastic scintillators. For scintillators having longer decay times, the time constants must be changed (see Section 5.3).

4.4.4 X-Ray Spectroscopy Using Proportional Counters

Space charge effects in proportional counters operated at high gas amplification tend to drastically degrade the resolution capabilities at x-ray energies, even at relatively low counting rates. By using a high gain, low noise amplifying system and lower gas amplification, these effects can be reduced and a considerable improvement in resolution can be obtained. The block diagram in Figure 4-10 shows a system of this type. Analysis can be accomplished by simultaneous acquisition of all data on a multichannel analyzer or counting a region of interest in a single channel analyzer window with a scaler and timer or counting rate meter.

4.5 Typical System Block Diagrams

This section contains block diagrams illustrating how the 435 and other ORTEC 400 Series modules can be used in experimental set-ups.

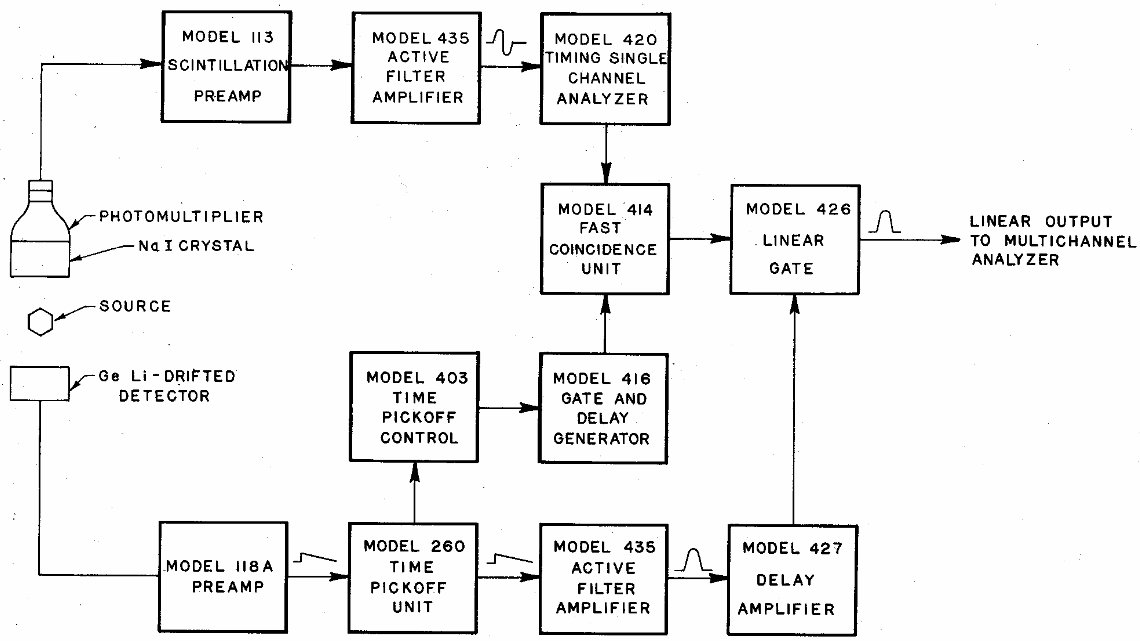


Figure 4-11 Gamma Ray-Gamma Ray Coincidence Experiment - Block Diagram

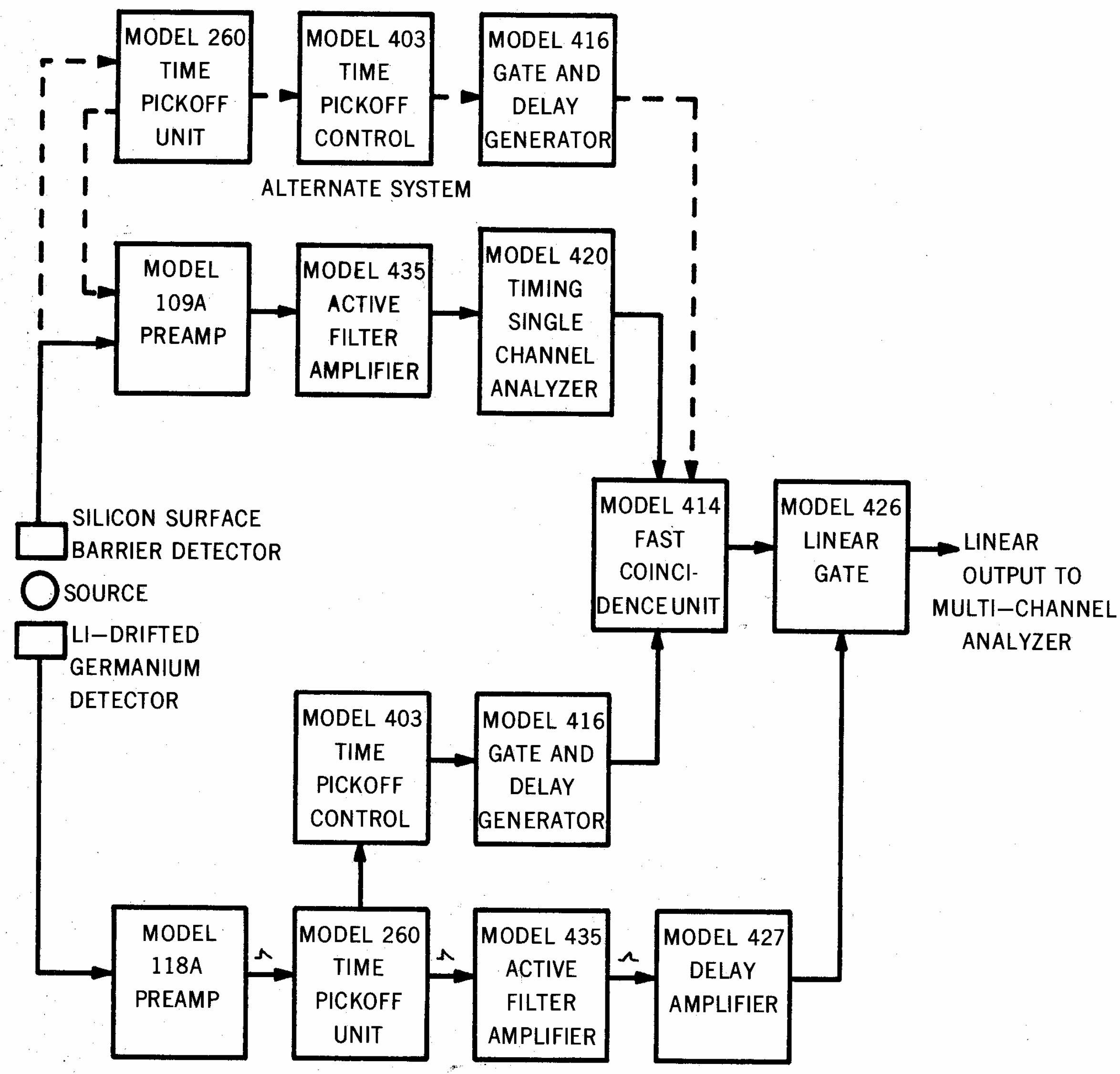


FIGURE 4–12 GAMMA RAY–CHARGED PARTICLE COINCIDENCE EXPERIMENT–BLOCK DIAGRAM

5. CIRCUIT DESCRIPTION

5.1 General Block Diagram

The 435 Active Filter Amplifier contains four basic gain stages as shown in the block diagram in Fig. 5-1. The differential input stage provides the functions of polarity inversion and common mode noise rejection as well as additional amplification before the first clip to improve the noise characteristics of the amplifier.

The gain stages are integrated circuit amplifiers which provide wideband gain. Gain changing is accomplished by constant impedance T attenuators and a constant impedance potentiometer. The variable pole-zero cancelled first clipping network is between the input stage and the first attenuator. The capacitive roll-off preceding each gain stage reduces the high frequency response and increases the dynamic range of the gain stages for fast rise input pulses.

The active filter was described in general in Section 1.3. The filter is followed by the unipolar-bipolar switch which allows a choice of either unipolar or bipolar clipped pulses. A resistor in the unipolar path compensates for the added loss in the bipolar path and results in an equal amplitude output for either shaping mode.

The output driver stage provides the additional gain necessary to raise the maximum linear output to 10V. The stage also has sufficiently low output impedance to drive terminated or unterminated connecting cables.

5.2 Circuit Description

5.2.1 Differential Input Stage

Referring to the circuit diagram (Drawing 435-0101-S1), the input stage consists of a long-tail differential amplifier Q1 and Q2 driving a common-emitter output stage, Q3. The output is fed back to the input through R17 and C5. Transistor Q4 acts as a zener diode with about a 6.5-volt drop across the base to emitter junction.

When the NEG INPUT is used, the base voltage at Q2 follows the input voltage at the base of Q1. The POS INPUT must be terminated in 100 ohms under this condition and the gain is then R17/(R12 + 100) ohms.

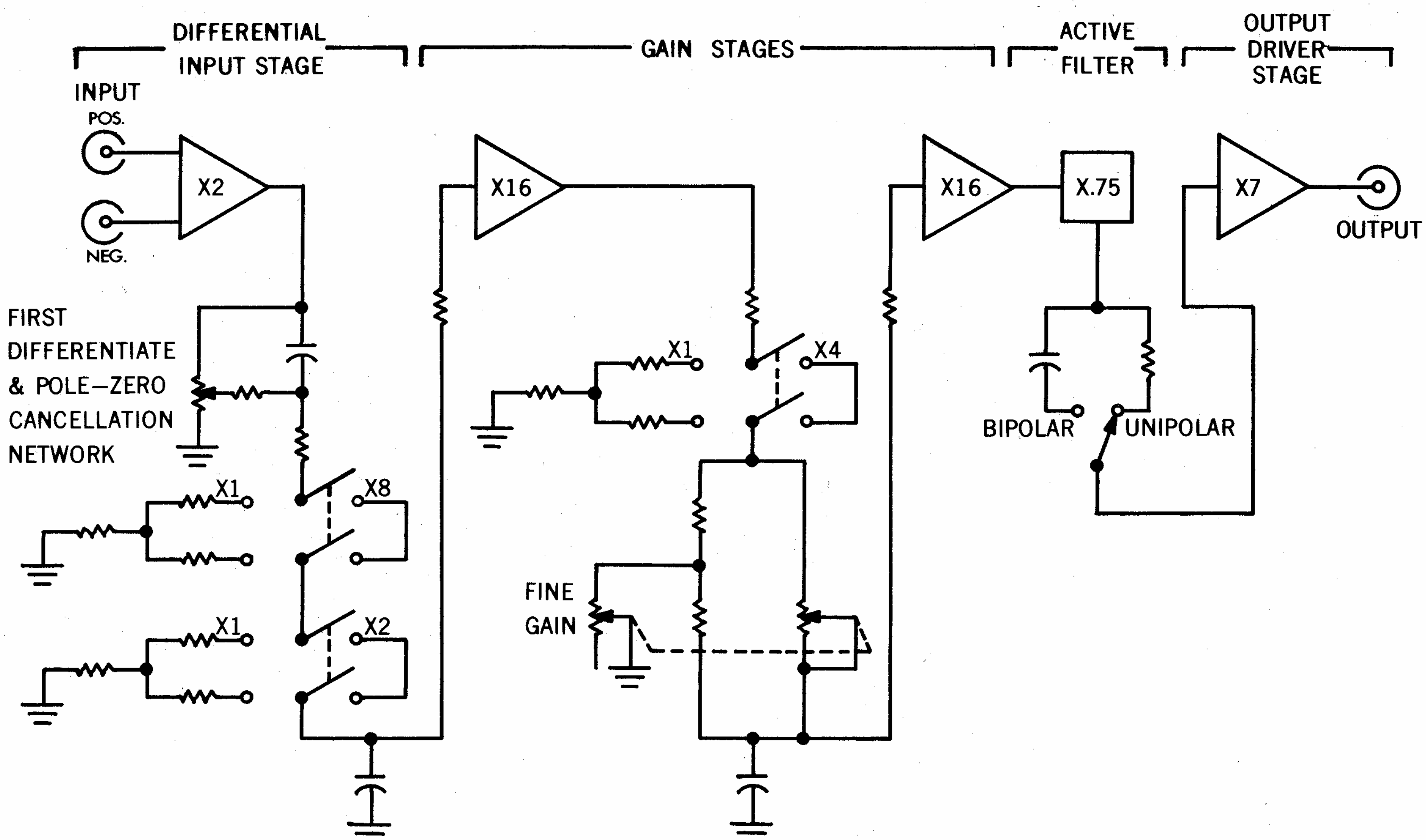


FIGURE 5–1 ACTIVE FILTER AMPLIFIER–BLOCK DIAGRAM

The network R4-C3 is a phase-lag compensation network to keep the stage from oscillating. The network C1-R5 is used to compensate for the different signal paths in the common-mode operating mode (positive signal at both inputs) and increase the common mode rejection at high frequencies (fast input rise times).

5.2.2 Pole-Zero Cancelled First Clip

The pole-zero cancelled first differentiate consists of C6, R19 and R20. For pole-zero cancellation:

$$\frac{R_{19} C_6}{K} = T_{preamp} \qquad \frac{R_{19}\ (R_{20} + R_{21})}{R_{19} + R_{20} + R_{21}} C_6 = T_{diff}$$

where T_{preamp} is the input decay time constant (50μsec) and T_{diff} is the differentiation time constant (1 μsec) and K is the fractional trim potentiometer resistance ratio ($0 \leqq K \leqq 1$). The derivation of these equations was discussed in Section 1.2.

5.2.3 Constant Impedance Attenuators

The attenuators employed are constant impedance T attenuators and the FINE GAIN control is a constant impedance Bridged T attenuator. The formulas for these attenuators are given in Fig. 5-2.

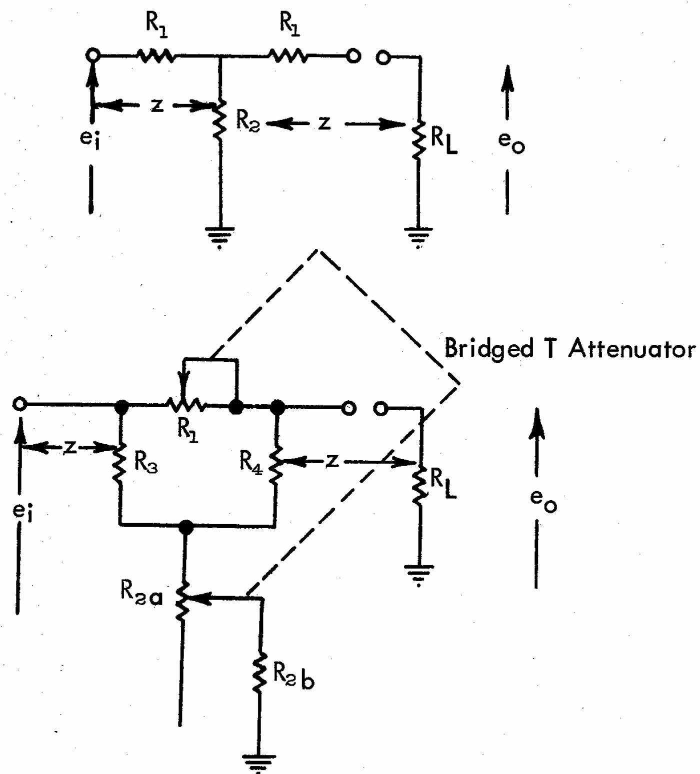
5.2.4 Gain Stages

Both gain stages utilize integrated circuit differential amplifiers in operational amplifier feedback circuits. The circuit has an open loop gain of greater than 1000 and a gain-bandwidth product greater than 1000 mHz. The integrated circuit contains a differential amplifier input driving a grounded emitter amplifier with an emitter follower output. The dc input and output levels of the first stage (IC-1) are at approximate ground; whereas the second stage (IC-2) has its input at ground and its output at about + 2 volts. This offset is necessary to provide dynamic range for the negative pulses at the output of IC-2. Phase lag roll-off networks C7-R22 and C12-R33 are necessary to prevent oscillation.

5.2.5 Active Filter

The basic active filter circuit was described in detail in Section 1.3. The filter network requires a unity gain amplifier which consists of emitter-follower Q6 and the constant current source Q5.

T Attenuator



$R_L = Z = 1000\ \Omega$

$R_1 = Z \dfrac{K - 1}{K + 1}$

$R_2 = \dfrac{2ZK}{K^2 - 1}$

$K = \dfrac{e_i}{e_o}$

$R_L = R_3 = R_4 = Z = 1000\Omega$

$R_1 = Z\,(K - 1)$

$R_2 = R_2a + R_2b = \dfrac{Z}{K - 1}$

$R_2b = 500\Omega$

Figure 5-2 ORTEC 435 Attenuator Networks

5.2.6 Output Driver Stage

The output driver stage consists of a grounded emitter amplifier Q7, with a high impedance, constant current load Q8, and a dc-offset-cancelling emitter-follower-quad consisting of Q9, Q10, Q11, and Q12. The output is fed back to the input by R46. The dc level potentiometer R48 is used to adjust the dc output to zero volts.

5.3 Circuit Modifications for Special Applications

5.3.1 Clipping and Integration Time Changes

In order to obtain optimum noise performance, both the clipping and integration times should be changed if a narrower or wider pulse is desired. To determine the proper values (for a new resolving time T_r) use the following formulas:

FIRST DIFFERENTIATE

1. Keep R20 the same (1000 ohms).
2. Change C6 and R19 to:

   C6 R19 = 40μsec (for 40μsec minimum pole-zero adjustment)

$$C6 \cdot \left(\frac{R_{19}\,(R_{20} + R_{21})}{R_{19} + R_{20} + R_{21}} \right) = T_r$$

SECOND DIFFERENTIATE

1. Keep R44 and R45 constant.
2. Change C21 to: $C21 \cdot R45 = T_r$

Active Filter

1. Keep R39 and R40 constant.
2. Change C17 and C18 to:

$$C17 \cdot R39 = T_r \qquad C18 = \frac{C17}{4}$$

3. Change C8 to: $C_8 = T_r/10^{-8}$

The two rolloff capacitors C8 and C13, necessary to obtain wide dynamic range and avoid nonlinearities, limit the minimum resolving time available from the amplifier. Operation with resolving times less than $T_r = 0.2 \mu sec$ are not recommended.

5.3.2 Changing Amplifier Attenuator Networks

By using the formulas in Fig. 5-2, the attenuator networks can be changed while maintaining constant overall impedance. However, the gain and dynamic range for each of the stages have been optimized and any changes could result in nonlinearities at some gain settings. The overall gain can be increased without difficulty by increasing the values of R25 and R36 with the added possibility of changing R35 to keep the proper dynamic range of the second gain stage.

5.3.3 Removing the Pole-Zero Cancellation Network

The pole-zero cancellation network at the first clip can be removed by simply adjusting the wiper to the ground side of the PZ Trimpot.

5.3.4 Removing the Output Isolation Capacitors

When the dc level trim potentiometer is properly adjusted, the Output Driver Stage output is at zero dc voltage. Therefore, the capacitors (C36 and C37) isolating the Output Driver Stage from the OUTPUT can be replaced by a wire (short across the capacitors) resulting in a dc coupled output.

With this dc coupled condition, the output will still sustain a direct short without catastrophic damage. The maximum counting rate at which the amplifier will sustain a short will be limited to 1000 cps for long duration shorts and 10^4 cps for one-minute shorts.