ORTEC®

Introduction to Amplifiers

Choosing the Right Amplifier for the Application

The amplifier is one of the most important components in a pulse processing system for applications in counting, timing, or pulse-amplitude (energy) spectroscopy. Normally, it is the amplifier that provides the pulse-shaping controls needed to optimize the performance of the analog electronics. Figure 1 shows typical amplifier usage in the various categories of pulse processing.

When the best resolution is needed in energy or pulse-height spectroscopy, a linear pulse-shaping amplifier is the right solution, as illustrated in Fig. 1(a). Such systems can acquire spectra at data rates up to 7,000 counts/s with no loss of resolution, or up to 86,000 counts/s with some compromise in resolution.

The linear pulse-shaping amplifier can also be used in simple pulse-counting applications, as depicted in Fig. 1(b). Amplifier output pulse widths range from 3 to 70 µs, depending on the selected shaping time constant. This width sets the dead time for counting events when utilizing an SCA, counter, and timer. To maintain dead time losses <10%, the counting rate is typically limited to <33,000 counts/s for the 3-µs pulse widths and proportionately lower if longer pulse widths have been selected.

Some detectors, such as photomultiplier tubes, produce a large enough output signal that the system shown in Fig. 1(d) can be used to count at a much higher rate. The pulse at the output of the fast timing amplifier usually has a width less than 20 ns. Consequently, maximum counting rates in excess of a few MHz are feasible without suffering more than 10% dead time losses.

The two common schemes for deriving signals to achieve nanosecond and sub-nanosecond time resolution are outlined in Fig. 1(c) and (e). Both applications utilize a fast timing amplifier. Fig. 1(e) illustrates the preferred solution for single-photon or single-ion detection and timing with photomultiplier tubes, electron multipliers, microchannel plates, microchannel plate PMTs, and channeltrons. Although the scheme designated in Fig. 1(c) can also be used with these same types of detectors, it is more commonly employed with high-resolution semiconductor detectors, since such detectors require a low-noise, charge-sensitive preamplifier.

Whatever your application, the brief descriptions of performance characteristics on the next few pages and the selection guide charts that follow will help you to choose the best amplifier for your situation.

Fast-Timing Amplifiers

When a detector signal from the preamplifier or photomultiplier tube is of sufficient amplitude, direct coupling of that output to a timing discriminator provides the best available rise time, and minimizes the effects of noise on time resolution. When a detector signal must be amplified or shaped before deriving the time information, an amplifier specifically designed for timing should be used.

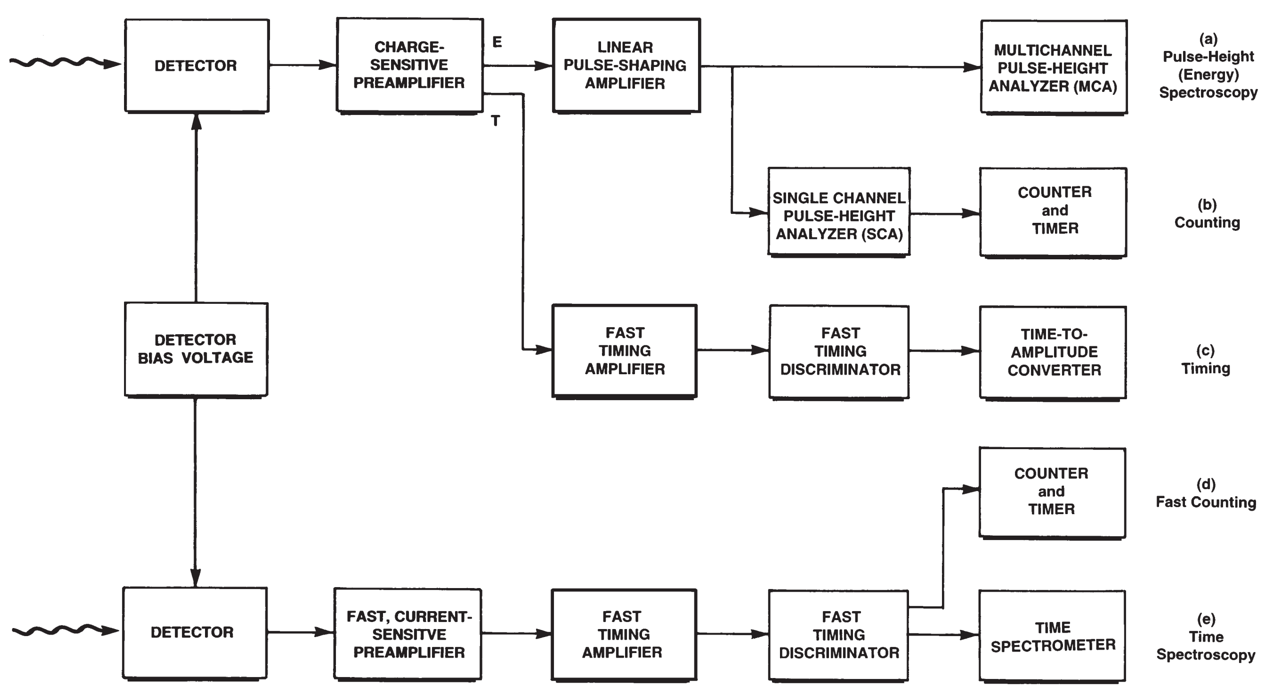


Fig. 1. Typical Amplifier Applications in Pulse Processing.

Timing amplifiers are designed to have output rise times in the low nanosecond or sub-nanosecond range. Achieving such fast rise times usually compromises linearity and temperature stability. The latter parameters are not as important as low noise and fast rise times in timing applications. The output pulse polarity is normally negative for compatibility with fast timing discriminators, which were historically designed to work directly with the negative output pulses from photomultiplier tube anodes.

Two types of fast amplifiers are available: wideband amplifiers and timing filter amplifiers. Wideband amplifiers offer no control over the rise time or the decay time of the signal. They are typically used with photomultiplier tubes [Fig. 1(e)], and silicon charged-particle detectors [Fig. 1(c)], where the fastest rise times are required for good time resolution. Wideband amplifiers rely on the preceding electronics to limit the pulse length. Timing filter amplifiers offer independent CR differentiator and RC integrator controls for adjustable pulse shaping. The timing filter amplifier is used with germanium detectors (Fig. 2), or for any other application requiring adjustment of the pulse shaping. Both types of amplifiers may be either ac- or dc-coupled. The timing filter amplifiers typically include a baseline restorer.

For timing applications with either type of amplifier, the rise time should be selected to be less than the inherent rise time of the preamplifier so that there will be no degradation of the signal rise time. Excessively fast amplifier rise times should be avoided, since they will result in more noise and no improvement in the signal rise time. If adjustment of the differentiator time constant is available, it should be set just long enough to avoid significant loss of signal amplitude.

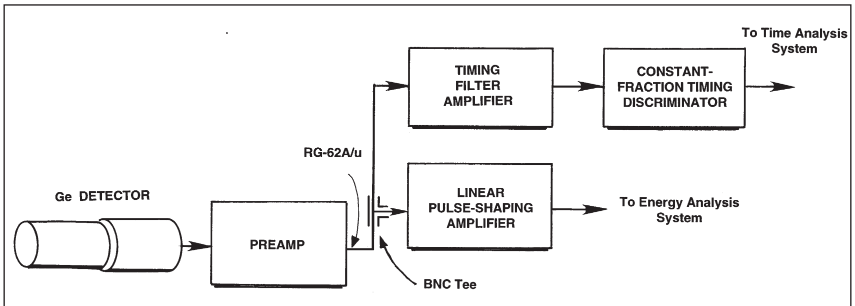


Fig. 2. Application of the Timing Filter Amplifier.

Linear, Pulse-Shaping Amplifiers for Pulse-Height (Energy) Spectroscopy

For pulse-height or energy spectroscopy, the linear, pulse- shaping amplifier performs several key functions. Its primary purpose is to magnify the amplitude of the preamplifier output pulse from the millivolt range into the 0.1- to 10-V range. This facilitates accurate pulse amplitude measurements with analog-to-digital converters, and single-channel pulse-height analyzers. In addition, the amplifier shapes the pulses to optimize the energy resolution, and to minimize the risk of overlap between successive pulses. Most amplifiers also incorporate a baseline restorer to ensure that the baseline between pulses is held rigidly at ground potential in spite of changes in counting rate or temperature.

Frequently, the requirement to handle high counting rates is in conflict with the need for optimum energy resolution. With many detector-preamplifier combinations, achieving the optimum energy resolution requires long pulse widths. On the other hand, short pulse widths are essential for high counting rates. In such cases a compromise pulse width must be selected which optimizes the quality of information collected during the measurement.

The following sections describe the various techniques available for pulse shaping in the linear amplifier. Each method has benefits for specific applications.

Accepting Preamplifier Pulse Shapes

The linear, pulse-shaping amplifier must accept the output pulse shapes provided by the preamplifier and change them into the pulse shapes required for optimum energy spectroscopy. Two general types of charge-sensitive preamplifiers are in common use: the resistive-feedback preamplifier,* and the pulsed-reset preamplifier. Each of these places slightly different demands on the amplifier's functions.

The Resistive-Feedback Preamplifier

Figure 3(a) illustrates the typical output pulse shapes from a resistive-feedback preamplifier. The output for each pulse consists of a rapidly rising step, followed by a slow exponential decay. It is the amplitude of the step that represents the energy of the detected radiation. The exponential decay time constant is normally determined by the

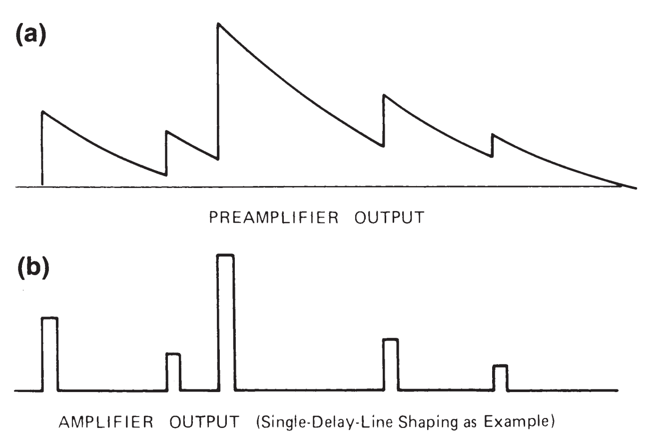


Fig. 3. Output Pulse Shapes from (a) a Resistive-Feedback Preamplifier, and (b) the Delay-Line Shaping Amplifier Connected to the Preamplifier.

* Pulse shapes from a parasitic-capacitance preamplifier are similar to those from a resistive-feedback, charge-sensitive preamplifier.

feedback resistor in parallel with the feedback capacitor. Decay time constants of 50 μs are prevalent, but longer time constants are encountered on some preamplifiers.

For detectors with very short charge collection times, the rise time of the preamplifier output pulse is controlled by the preamplifier itself, and the rise time is usually in the range from 10 to 100 ns. For detectors with long charge collection times, such as NaI(Tl) detectors, proportional counters, and coaxial germanium detectors, the output rise time of the preamplifier is controlled by the detector charge collection time. The output rise time can range up to 700 ns for large coaxial germanium detectors, and into the microsecond range for positive ion collection with proportional counters. For NaI(Tl) detectors, the scintillator decay time causes a preamplifier output rise time of approximately 500 ns.

In normal operation at ordinary counting rates, the rising step caused by each detector event rides on the exponential decay of a previous event, and the preamplifier output does not get a chance to return to the baseline. Since the amplitude of detector events is usually variable and the time of occurrence is random, the preamplifier output is usually irregular, as shown in Fig. 3(a). As the counting rate increases, the piling up of pulses on the tails of previous pulses increases, and the excursions of the preamplifier output move farther away from the baseline. The power supply voltages eventually limit the excursions, and determine the maximum counting rate that can be tolerated without distortion of the output pulses.

Before amplification, the pulse-shaping amplifier must replace the long decay time of the preamplifier output pulse with a much shorter decay time. Otherwise, the acceptable counting rate would be severely restricted. Figure 3(b) demonstrates this function using the simple example of a single-delay-line, pulse-shaping amplifier. The energy information represented by the amplitudes of the steps from the preamplifier output has been preserved, and the pulses return to baseline before the next pulse arrives. This makes it possible for an analog-to- digital converter (ADC) to determine the correct energy by measuring the pulse amplitude with respect to the baseline. With the shorter pulse widths at the amplifier output, much higher counting rates can be tolerated before pulse pile-up again causes significant distortion in the measurement of the pulse heights above baseline.

Pulsed-Reset Preamplifiers

Pulsed-reset preamplifiers were developed to eliminate the noise contributions of the preamplifier feedback resistor, and to improve the high counting rate capability of the preamplifier. There are two types: Pulsed optical feedback preamplifiers are often employed with Si(Li) detectors for x-ray spectrometry,[1] and transistor-reset preamplifiers are used to achieve high counting rates with germanium detectors.[2,3] In both cases the feedback resistor is replaced with a feedback device that is turned on only for the very short time needed to reset the preamplifier output back to the baseline. The behavior at the output of the preamplifier is illustrated in Fig. 4(a).

With no feedback resistor to remove the charge from the feedback capacitor between detector events, each event steps the preamplifier output up to a higher dc voltage. Eventually, the staircase of pulses approaches the power supply voltage, and the voltage across the feedback capacitor must be reset back to the starting value. A voltage comparator in the preamplifier senses the upper limit of the staircase, and turns on the reset device just long enough to discharge the feedback capacitor back to the starting condition. By this method, the preamplifier output is maintained within its linear operating range, even at high counting rates. The limitation on counting rate with a pulsed-reset preamplifier is the percent dead time caused by the reset. At higher counting rates the reset must happen more frequently. When the percent dead time from resetting becomes too high to tolerate, the upper limit on counting rate has been reached.

Although the preamplifier output looks different from that with resistive-feedback preamplifiers, the function of the amplifier with pulsed-reset preamplifiers is similar. The pulse-shaping amplifier

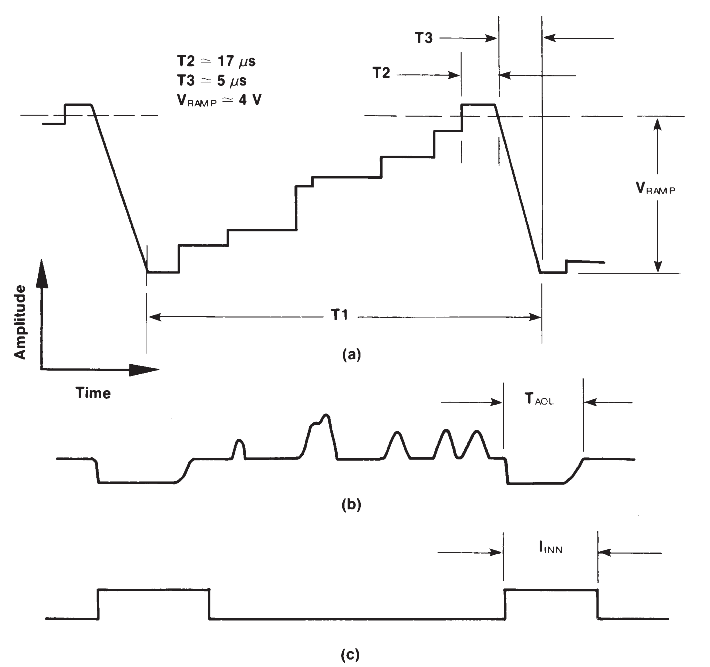


Fig. 4. (a) The Output from a Transistor-Reset Preamplifier; (b) the Same Events After Passing through a Semi-Gaussian Pulse-Shaping Amplifier; (c) the Inhibit Signal, which Prevents Data Collection During Reset and Reset Recovery.

[1]Ron Jenkins, R.W. Gould, Dale Gedcke, Quantitative X-Ray Spectroscopy, Marcel Dekker Inc, New York, 1981, pp 175–177.

[2]D.A. Landis, C.P. Cork, N.W. Madden, F.S. Goulding, *IEEE Trans. Nucl. Sci.,* **NS**-29(1), 619 (1982).

[3]C.L. Britton, T.H. Becker, T.J. Paulus, R.C. Trammell, *IEEE Trans. Nucl. Sci.,* **NS**-31(1), 455 (1984).

must preserve the amplitude of the steps from the preamplifier, and cause the pulses to return to baseline quickly between the steps. This function is demonstrated in Fig. 4(b) using a semi-Gaussian, pulse-shaping amplifier. Although slightly rounded in shape to improve the signal-to-noise ratio, the amplitudes of the amplifier output pulses are proportional to the step amplitudes from the preamplifier.

One additional characteristic shows up at the amplifier output with a pulsed-reset preamplifier. Each preamplifier reset causes a large, negative polarity, output pulse to be generated. The duration of this reset recovery pulse is determined by the pulse-shaping circuits in the amplifier, the gain of the amplifier, and the voltage swing of the reset. Typically, it lasts two to three times as long as the positive polarity pulses from detector events. During the reset-recovery pulse, data collection must be inhibited to prevent measurement of distorted pulse heights. The inhibit logic signal in Fig. 4(c) is generated by the preamplifier and/or the amplifier, and is used to inhibit data acquisition in the pulse-height analyzer during reset recovery.

With both the resistive-feedback preamplifier and the pulsed-reset preamplifier, the amplifier input must be able to accept the voltage swings of the preamplifier output without causing any distortion of the pulse amplitudes.

Delay-Line Pulse Shaping

Amplifiers employing delay-line pulse shaping are well suited to the pulse processing requirements of scintillation detectors. The propagation delay of distributed or lumped delay lines can be combined into suitable circuits to produce an essentially rectangular output pulse from each step-function input pulse. For pulse pile-up prevention, this shaping method is close to ideal because an immediate return to baseline is obtained. With scintillation detectors, the signal-to-noise ratio of the preamplifier and amplifier combination is seldom a limitation on the energy resolution. As a result of the high gain of the photomultiplier tube, the energy resolution is determined by the statistics of light production in the scintillator and the conversion to photoelectrons at the cathode. However, for detectors having no internal gain, delay-line shaping is seldom appropriate, because the signal-to-noise ratio for preamplifier noise with delay-line shaping is inferior to that obtained with simple CR-RC or semi-Gaussian shaping.

There are many circuits that can be used for delay-line shaping, and the circuit shown in Fig. 5 is typical of one that is very tolerant of delay-line imperfections. The step pulse from the preamplifier is inverted, delayed, and added back to the original step pulse. The result is a rectangular output pulse with a width equal to the delay time of the delay line. In practice, the value of the resistor labeled $2R_D$ is made adjustable over a small portion of its nominal value to allow compensation for the exponential decay of the input pulse. With proper adjustment, the output pulse will return to baseline promptly without undershoot. The main advantage of delay-line shaping is a rectangular output pulse with fast rise and fall times. In fact, the falling edge of the pulse is a delayed mirror image of the rising edge. These characteristics make delay-line pulse shaping ideal for timing and pulse-shape discrimination applications with scintillation detectors at low or high counting rates.

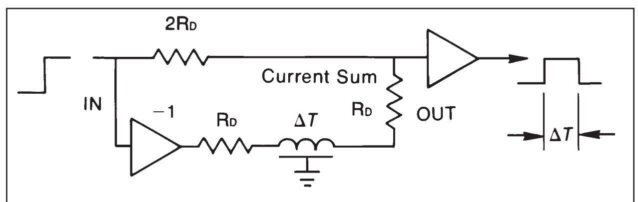


Fig. 5. Single-Delay-Line Pulse Shaping.

By following one delay-line shaper with a second, a doubly- differentiated delay-line shape is obtained, as illustrated in Fig. 6. The result is an output pulse shape that has a positive rectangular lobe followed by a negative rectangular lobe with equal amplitude and duration. The double-delay-line shaping is ideal for use with scintillation detectors in systems incorporating ac-coupling. The baseline shift caused by changing counting rates in ac-coupled systems is virtually eliminated by the two lobes having equal area above and below the baseline. This benefit is gained at the expense of doubling the pulse width. Double-delay-line shaping is often useful for simple zero-crossover timing with scintillation detectors at low or high counting rates. Double-delay-line shaping is not a good choice for detectors having a substantial preamplifier noise. Its signal-to-noise ratio is worse than single-delay-line shaping, and much worse than semi-Gaussian shaping.

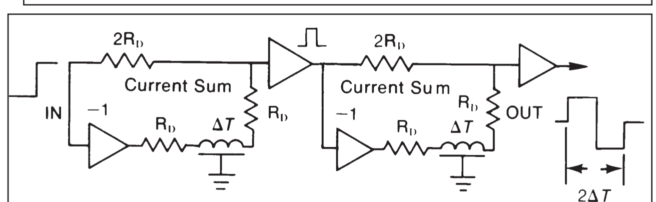


Fig. 6. Double-Delay-Line Pulse Shaping.

CR-RC Pulse Shaping

The simplest concept for pulse shaping is the use of a CR high-pass filter followed by an RC low-pass filter. Although this rudimentary filter is rarely used, it encompasses the basic concepts essential for understanding the higher-performance, active filter networks.

In the amplifier, the preamplifier signal first passes through a CR, high-pass filter (Fig. 7). This improves the signal-to-noise ratio by attenuating

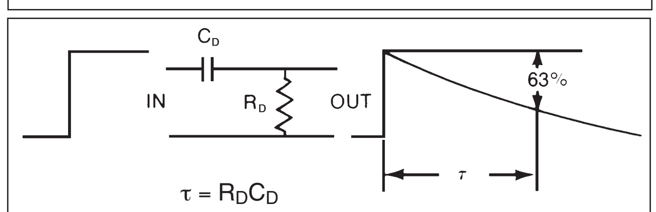


Fig. 7. CR Differentiation.

the low frequencies, which contain a lot of noise and very little signal. The decay time of the pulse is also shortened by this filter. For that reason, it is often referred to as a "CR differentiator." (Note that the differentiation function is not a true mathematical differentiation.)

Just before the pulse reaches the output of the amplifier, it passes through an RC low-pass filter (Fig. 8). This improves the signal-to-noise ratio by attenuating high frequencies, which contain excessive noise. The rise time of the pulse is lengthened by this filter. Although this filter does not perform an exact mathematical integration, it is frequently called an "RC integrator."

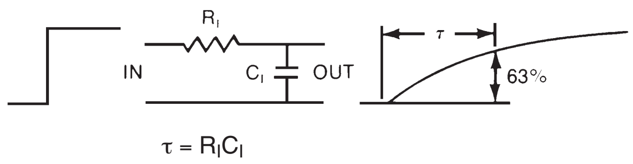


Fig. 8. RC Integration.

Figure 9 demonstrates the effect of combining the high-pass and low-pass filters in an amplifier to produce a unipolar output pulse. Typically, the differentiation time constant $\tau_D = C_DR_D$ is set equal to the integration time constant $\tau_I = R_IC_I$, i.e., $\tau_D = \tau_I = \tau$. In that case, the output pulse rises slowly and reaches its maximum amplitude at 1.2τ. The decay back to baseline is controlled primarily by the time constant of the CR differentiator. In this simple circuit there is no compensation for the long decay time of the preamplifier. Consequently, there is a small amplitude undershoot starting at about 7τ. This undershoot decays back to baseline with the long time constant provided by the preamplifier output pulse.

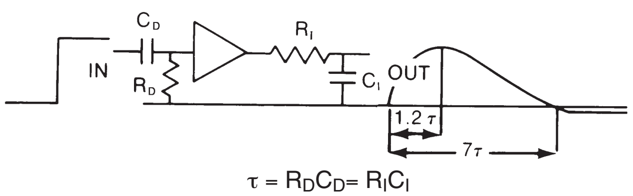


Fig. 9. CR-RC Pulse Shaping.

This pulse-shaping technique can be used with scintillation detectors. For that application, the shaping time constant τ should be chosen to be at least three times the decay time constant of the scintillator to ensure complete integration of the scintillator signal. The disadvantage in using CR-RC shaping with scintillation detectors is the much longer pulse duration compared with that of single-delay-line shaping.

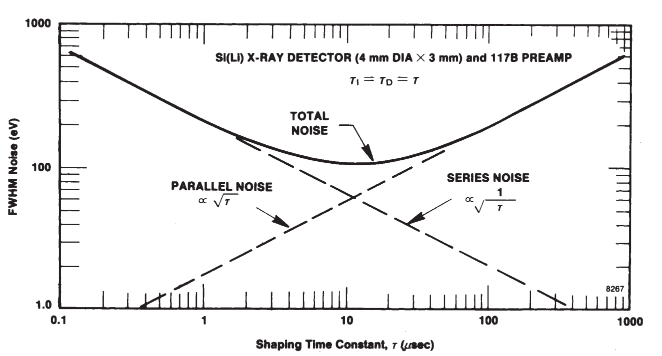


Fig. 10. The Dependence of the Preamplifier Noise Contribution on the Amplifier Shaping Time Constant.

On silicon and germanium detectors, the electronic noise at the preamplifier input makes a noticeable contribution to the energy resolution of the detector. This noise contribution can be minimized by choosing the appropriate amplifier shaping time constant. Figure 10 shows the effect. At short shaping time constants, the series noise component of the preamplifier is dominant. This noise is typically caused by thermal noise in the channel of the field-effect transistor, which is the first amplifying stage in the preamplifier. At long shaping time constants the parallel noise component at the preamplifier input dominates. This component arises from noise sources that are effectively in parallel with the detector at the preamplifier input (e.g., detector leakage current, gate leakage current in the field-effect transistor, and thermal noise in the preamplifier feedback resistor). The total noise at any shaping time constant is the square root of the sum of the squares of the series and parallel noise contributions. Consequently, the total noise has a minimum value at the shaping time constant where the series noise is equal to the parallel noise. This time constant is called the "noise corner time constant." The time constant for minimum noise will depend on the characteristics of the detector, the preamplifier, and the amplifier pulse shaping network. For silicon charged-particle detectors, the minimum noise usually occurs at time constants in the range from 0.5 to 1 µs. Generally, minimum noise for germanium and Si(Li) detectors is achieved at much longer time constants (in the range from 6 to 20 µs). Such long time constants impose a severe restriction on the counting rate capability. Conse-quently, energy resolution is often compromised by using shorter shaping time constants, in order to accommodate higher counting rates.

Figure 11 demonstrates the bipolar output pulse obtained when a second differentiator is inserted just before the amplifier output. Double differentiation produces a bipolar pulse with equal area in its positive and negative lobes. It is useful in minimizing baseline shift with varying counting rates when the electronic circuits following the amplifier are ac-coupled. It is also convenient for zero-crossover timing applications. The drawbacks of double differentiation relative to single CR differentiation are a longer pulse duration and a worse signal-to-noise ratio.

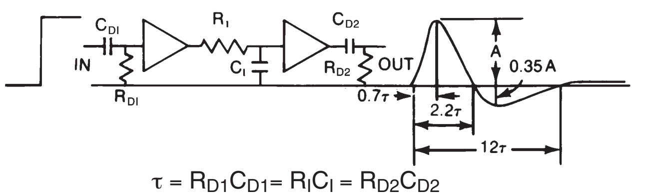


Fig. 11. Doubly-Differentiated CR-RC-CR Shaping.

Pole-Zero Cancellation

In the simple CR-RC circuit described in Fig. 9, there is a noticeable undershoot as the amplifier pulse attempts to return to the baseline. This is a result of the long exponential decay on the preamplifier output pulse. At medium to high counting rates, a substantial fraction of the amplifier output pulses will ride on the undershoot from a previous pulse. The apparent pulse amplitudes measured for these pulses will be too low, which leads to a broadening of the peaks recorded in the energy spectrum. Most spectroscopy amplifiers incorporate a pole-zero cancellation circuit to eliminate this undershoot. The benefit of pole-zero cancellation is improved peak shapes and resolution in the energy spectrum at high counting rates.

Figure 12 illustrates the pole-zero cancellation network, and its effect. In Fig. 12(a), the preamplifier signal on the left is applied to the input of the normal CR differentiator circuit in the amplifier. The output pulse from the differentiator exhibits the undesirable undershoot. The following equation applies:

$$\frac{\text{Undershoot Amplitude}}{\text{Pulse Amplitude}} = \frac{\text{Differentiator Time Constant}}{\text{Decay Time Constant of Preamp Pulse}}$$

For a given preamplifier decay time constant, longer amplifier shaping time constants yield larger undershoots.

In Fig. 12(b), the resistor R_{pz} is added in parallel with capacitor C_D, and adjusted to cancel the undershoot. The result is an output pulse exhibiting a simple exponential decay to baseline with the desired differentiator time constant. This circuit is termed a "pole-zero cancellation network" because it uses a zero to cancel a pole in the mathematical representation by complex variables. Virtually all spectroscopy amplifiers incorporate this feature, with the pole-zero cancellation adjustment accessible through the front panel. Exact adjustment is critical for good spectrum fidelity at high counting rates. Some of the more sophisticated amplifiers simplify this task with an automatic PZ-adjusting circuit.

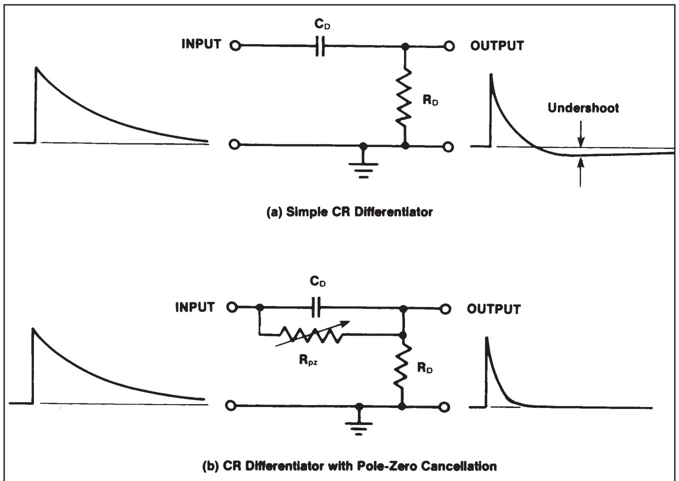


Fig. 12. The Benefit of Pole-Zero Cancellation.

Semi-Gaussian Pulse Shaping

By replacing the simple RC integrator with a more complicated active integrator network (Fig. 13), the signal-to-noise ratio of the pulse-shaping amplifier can be improved by 17% to 19% at the noise corner time constant. This is important for semiconductor detectors, whose energy resolution at low energies and short shaping time constants is limited by the signal-to-noise ratio. Amplifiers incorporating the more complicated filters are typically called "semi-Gaussian shaping amplifiers" because their output pulse shapes crudely approximate the shape of a Gaussian curve [Fig. 14(a)]. A further advantage of the semi-Gaussian pulse shaping is a reduction of the output pulse width at 0.1% of the pulse amplitude. At the noise corner time constant, semi-Gaussian shaping can yield a 22% to 52% reduction in output pulse width compared with the CR-RC filter. This leads to better baseline restorer performance at high counting rates. The reduction in pulse width corresponds to a 9% to 13% reduction in the amplifier dead time per pulse.

Although the unipolar output pulse from a semi-Gaussian shaping amplifier is normally the better choice for energy spectroscopy [Fig. 14(a)], a bipolar output is typically also

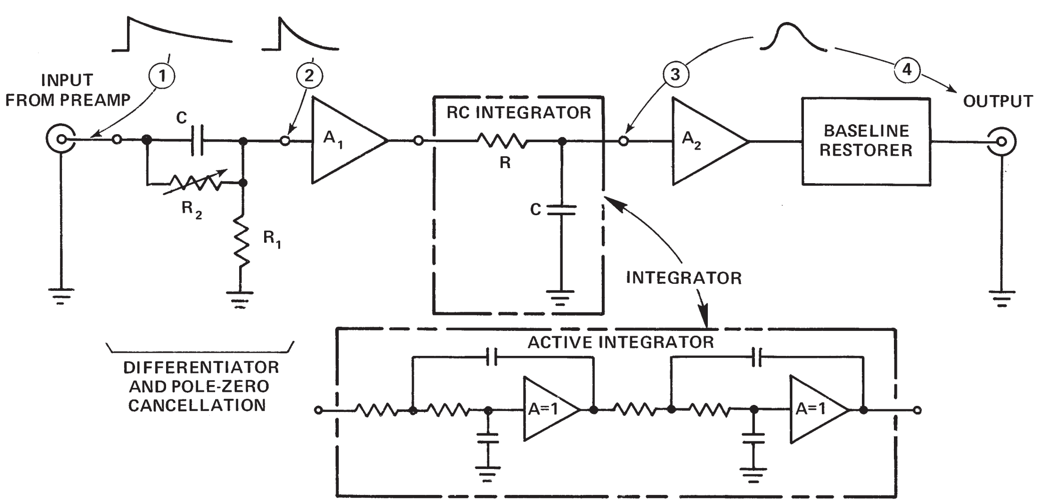
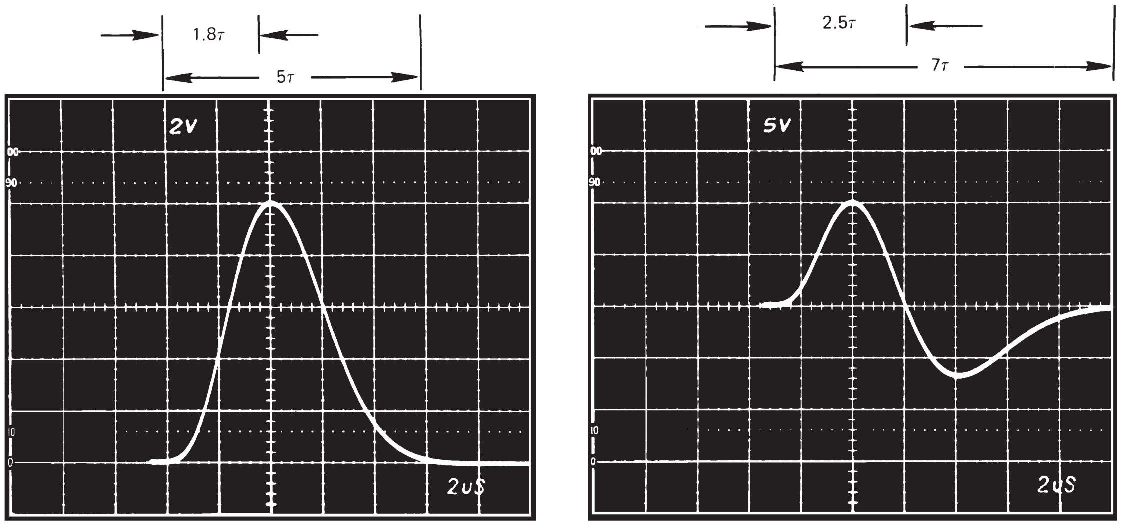


Fig. 13. Pulse Shaping in the Semi-Gaussian Shaping Amplifier.



a. Unipolar Output. b. Bipolar Outputs.

Fig. 14. Typical (a) Unipolar, and (b) Bipolar Output Pulse Shapes from a Semi-Gaussian Shaping Amplifier.

available [Fig. 14(b)]. The bipolar output is useful in minimizing baseline shift with varying counting rates when the electronic circuits following the amplifier are ac-coupled. It is also convenient for zero-crossover timing applications. The drawbacks inherent in the bipolar output relative to the unipolar output are a longer pulse duration and a worse signal-to-noise ratio.

Quasi-Triangular Pulse Shaping

By summing contributions from the various filter stages in a semi-Gaussian amplifier, a unipolar output pulse with a much more linear rise can be generated [Fig. 15(b)]. This pulse shape is referred to as quasi-triangular because it is a crude approximation to a true triangular pulse shape. The quasi- triangular pulse shaping is advantageous at shaping time constants shorter than the noise corner time constant. Under these conditions, the series noise component is dominant. Consequently, the quasi-triangular pulse shape yields approximately 8% lower noise than the semi-Gaussian pulse shape, with virtually identical dead time per amplifier pulse.

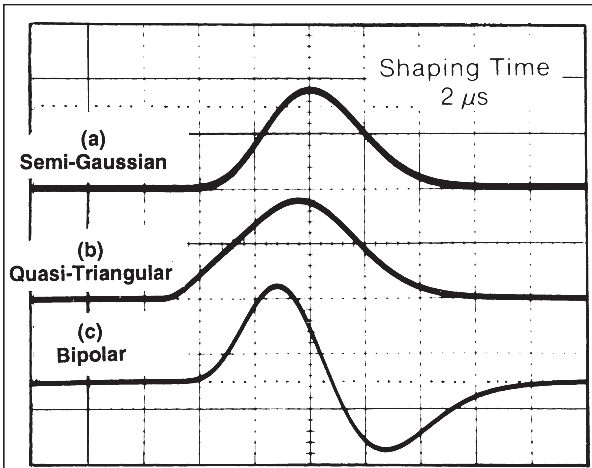


Fig. 15. A Comparison of (a) Semi-Gaussian, (b) Quasi-Triangular, and (c) Bipolar Pulse Shapes at a 2-µs Shaping Time Constant. Vertical scale, 5 V per division; horizontal scale, 2 µs per division.

Gated-Integrator Pulse Shaping

With germanium detectors, the time required to collect all of the charge from a gamma-ray interaction in the detector depends on the location of the interaction in the detector. The charge collection time can vary from 100 to 200 ns in a small detector, and by as much as 200 to 700 ns in a large germanium detector. As a result, the preamplifier output pulses have rise times varying over that same time range. In conventional pulse-shaping amplifiers (e.g., semi-Gaussian pulse shaping), these variations in rise time can affect the amplitude of the amplifier output pulse and cause degradation of the energy resolution. The longer rise times on the preamplifier output pulse cause a lower amplitude on the amplifier output pulse. This effect is called the "ballistic deficit." For shaping time constants in the range from 6 to 10 µs, the effect is negligible, because the peaking time of the amplifier output pulse is very long compared with the longest charge collection time in the germanium detector. However, when high counting rates are anticipated, much shorter shaping time constants must be used. The contribution of ballistic deficit to resolution degradation increases rapidly as the shaping time constant is reduced below 2 µs. Consequently, ballistic deficit becomes the dominant limitation on energy resolution at high counting rates using conventional, semi-Gaussian, or triangular pulse-shaping amplifiers.

The gated-integrator amplifier solves the ballistic deficit problem by integrating the signal until all the charge is collected from the detector. Figures 16 and 17 illustrate the principle. For simplicity, the prefilter has been depicted as a delay-line shaping amplifier. The width of the prefilter pulse determines the shaping time for the entire gated-integrator amplifier. For illustration purposes, two extreme rise timecases are drawn for the preamplifier pulse: zero rise time (solid line) and a long rise time (dashed line). At the output of the prefilter, the zero rise time pulse produces a rectangular pulse shape, while the longer rise time pulse generates a trapezoid. The duration of the trapezoid is longer than the rectangular pulse by an amount equal to the preamplifier pulse rise time.

The gated-integrator portion of the amplifier serves two functions. It reduces the high-frequency noise contribution, and it eliminates the ballistic deficit. Before the prefilter pulse arrives, switch S1 is open and switch S2 is closed, causing the gated-integrator output to be at ground potential. At the instant the prefilter pulse arrives, switch S1 closes and switch S2 opens, and the prefilter signal is integrated on capacitor C_I. The integration period is set to last as long as the longest prefilter pulse duration. Consequently, all pulses generate the same output pulse amplitude from the gated integrator, independent of their rise time at the preamplifier output. At the end of the integration period, S1 opens and S2 closes to return the output pulse to baseline quickly. Because the filter characteristics are switched at certain points in time, the gated integrator is called a time-variant filter. In contrast, the amplifiers previously discussed have time-invariant filters.

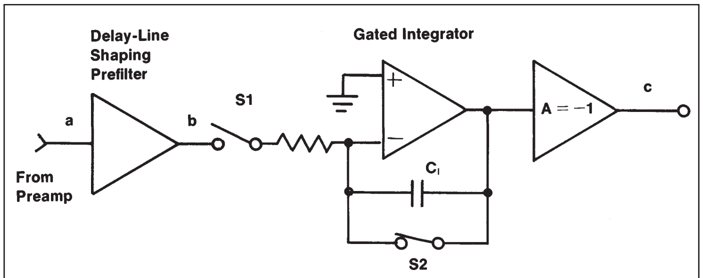


Fig. 16. A Simplified Representation of the Gated-Integrator Amplifier.

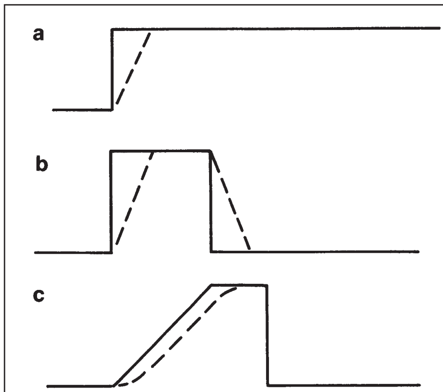


Fig. 17. Pulse Shapes in the Simplified Gated-Integrator Amplifier: (a) at the Preamplifier Output, (b) at the Prefilter Output, and (c) at the Gated-Integrator Output. See the corresponding points in Fig. 16.

The signal-to-noise ratio of the gated integrator approaches the performance of a time-invariant filter with a true triangular pulse shape. This makes it virtually the ideal filter for the short shaping times required for high counting rates.

Because it is difficult to implement a delay-line prefilter with a quality that is adequate for germanium detectors, practical gated integrator amplifiers typically utilize active RC networks in the prefilter. This results in the pulse shapes shown in Fig. 18. The deviation from a rectangular prefilter shape and the extra integration time required to accommodate the longest charge collection times causes a minor loss of signal-to-noise ratio compared with an ideal triangular pulse shape. However, the signal-to-noise ratio is less important than elimination of ballistic deficit for optimum energy resolution at the short shaping times required for high counting rates.

Gated-integrator amplifiers permit operation at ultra-high counting rates with germanium detectors without a substantial sacrifice of energy resolution (Fig. 19).

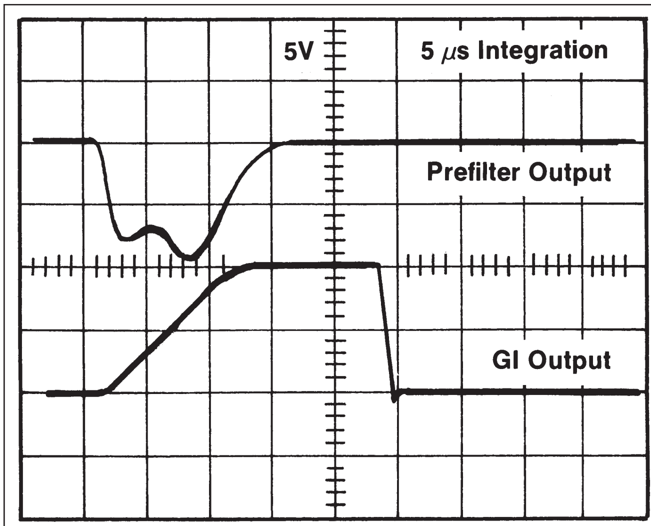


Fig. 18. Pulse Shapes in the Model 973 Gated-Integrator Amplifier for a 5-μs Integration Time.

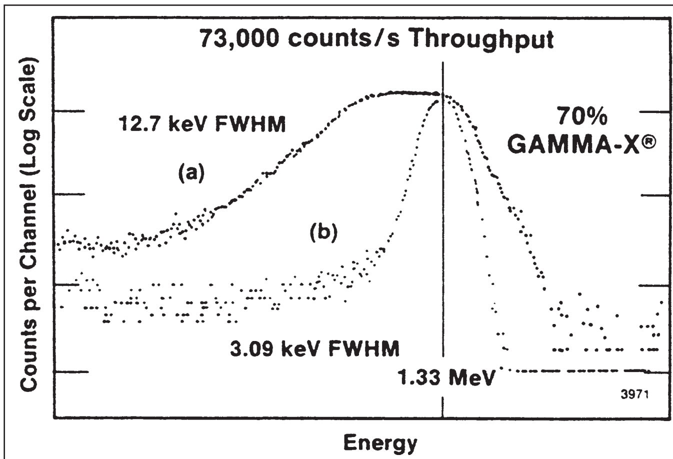


Fig. 19. The 1.33-MeV Gamma-Ray Peak from a ^{60}Co Source, Acquired with (a) a Model 672 Amplifier with a Triangular Pulse Shape and 0.5-μs Time Constant, and (b) the Model 973 Amplifier with a 2.5-μs Integration Time. Maximum amplifier throughput is 73,000 counts/s for both cases. (Peak heights normalized for comparison.)

The Baseline Restorer

To ensure good energy resolution and peak position stability at high counting rates, the higher-performance spectroscopy amplifiers are entirely dc-coupled (except for the CR differentiator network located close to the amplifier input). As a consequence, the dc offsets of the earliest stages of the amplifier are magnified by the amplifier gain to cause a large and unstable dc offset at the amplifier output. A baseline restorer is required to remove this dc offset, and to ensure that the amplifier output pulse rides on a baseline that is securely tied to ground potential.

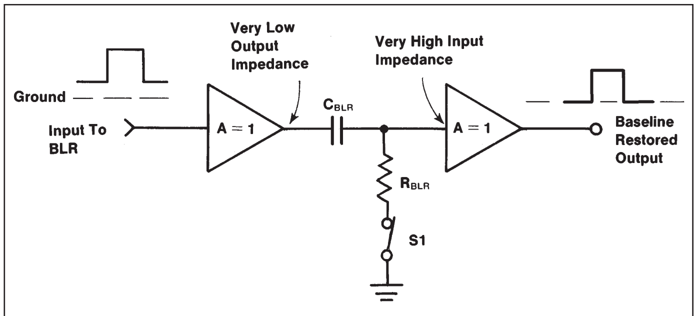


Fig. 20. A Simplified Diagram of a Baseline Restorer.

Figure 20 illustrates the basic principle of a baseline restorer. In the case of the simpler, time-invariant baseline restorers, switch S1 is always closed. The time-invariant baseline restorer behaves just like a CR differentiator. The baseline between pulses is returned to ground potential by resistor R_{BLR}. In order not to degrade the signal-to-noise ratio of the pulse-shaping amplifier, the C_{BLR} R_{BLR} time constant must be at least 50 times the shaping time constant employed in the amplifier.

The simple, time-invariant baseline restorer does not adequately maintain the baseline at ground potential at high counting rates. Since the time-invariant baseline restorer is really a CR differentiator, the average signal area above ground must equal the average signal area below ground at the baseline restorer output. At low counting rates, the spacing between pulses is extremely long compared with the pulse width. Consequently, the baseline between pulses remains very close to ground potential. As the counting rate increases, the baseline must shift down, so that the area of the signal remaining above ground potential is equal to the area between ground potential and the shifted baseline. The amount of baseline shift increases as the counting rate increases. Diode networks are typically incorporated to reduce this shift, but such solutions are unable to make the shift negligible.

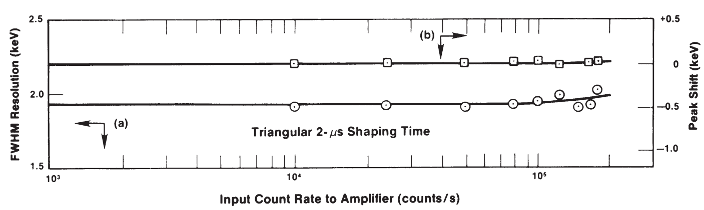


Fig. 21. (a) Resolution, and (b) Peak Position Stability as a Function of Counting Rate with a High-Performance, Gated Baseline Restorer. Measured on the 1.33-MeV gamma-ray line from a ^{60}Co radioactive source, using a 10% efficiency GAMMA-X PLUS detector.

The gated baseline restorer virtually eliminates the baseline shift caused by changing counting rates. In Fig. 20, switch S1 is opened for the duration of the amplifier pulse, and closed otherwise. Therefore, the CR differentiator function is active only on the baseline between pulses. The effect of the signal pulse is essentially eliminated. The gated baseline restorer perceives that it is operating at zero counting rate, and maintains the baseline firmly at ground potential, independent of the actual counting rate.

The stability of baseline restoration at very high counting rates with the gated baseline restorer depends on the ability of the gating control circuits to distinguish between the pulses and the baseline. In the simpler circuits, this is accomplished with a discriminator whose threshold is manually adjusted to sit just above the noise that surrounds the baseline. The more sophisticated amplifiers include automatic noise discriminators and more complicated pulse detection methods to perform this task more effectively. Figure 21 is an example of the results obtained on a high-performance baseline restorer. Peak shift and resolution broadening are both negligible over a very wide range of counting rates. At some upper limit on counting rate, there is inadequate baseline between pulses for the baseline restorer to control. Above that counting rate, the baseline will shift strongly with increasing counting rate. If counting rates must be processed above this limit, then a shorter amplifier shaping time constant must be selected.

Pile-Up Rejection

When two gamma rays arrive at the detector within the width of the spectroscopy amplifier output pulse, their respective amplifier pulses pile up to form an output pulse of distorted amplitude [Fig. 22(a)]. For detectors whose charge collection time is very short compared to the peaking time T_P of the amplifier output pulse, a pile-up rejector can be used to prevent analysis of these distorted pulses.

The pile-up rejector is implemented by adding a "fast" pulse shaping amplifier with a very short shaping time constant [Fig. 22(b)] in parallel with the "slow" spectroscopy amplifier. In the fast amplifier, the signal-to-noise ratio is compromised in favor of improved pulse-pair resolving time. A fast discriminator is set above the much higher noise level at the fast amplifier output to convert the analog pulses into digital logic pulses [Fig. 22(c)]. The trailing edge of the fast discriminator output triggers an inspection interval T_{INS} [Fig. 22(d)] that covers the width T_W of the slow amplifier pulse.

If a second fast discriminator pulse from a pile-up pulse arrives during the inspection interval, an inhibit pulse is generated [Fig. 22(e)]. The inhibit pulse is used in the associated ADC or multichannel analyzer to prevent analysis of the piled-up event.

As demonstrated in Figure 23, the pile-up rejector can deliver a substantial reduction in the pile-up background at high counting rates with germanium and Si(Li) detectors.

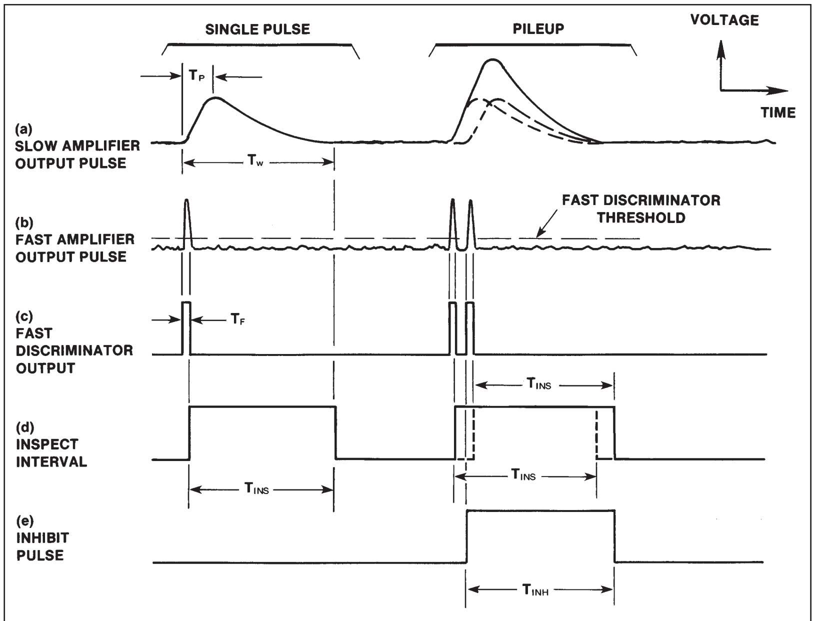


Fig. 22. Basic Waveforms in the Pile-Up Rejector.

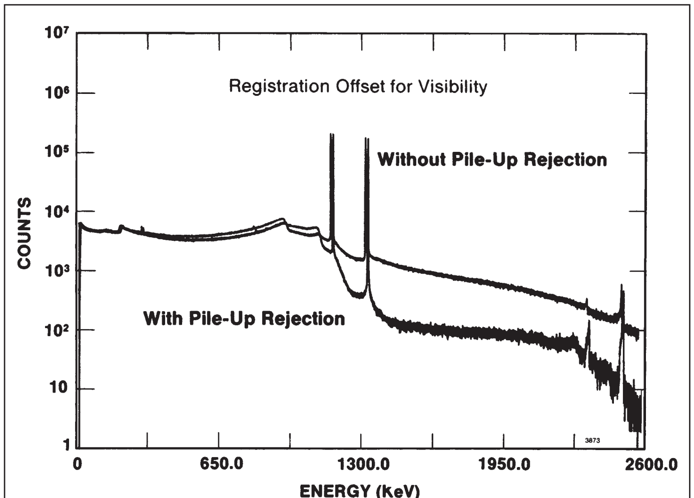


Fig. 23. Demonstration of the Effectiveness of the Pile-Up Rejector in Suppressing the Pile-Up Spectrum with a Germanium Detector and a ^{60}Co Spectrum at 50,000 Counts/s.

Amplifier Throughput

The pulse shape from the spectroscopy amplifier contributes to the dead time of the spectrometry system. The dead time attributable to the amplifier pulse shape is

$$T_D = T_P + T_W$$

where T_W is the width of the pulse above the noise level, and T_P is the time from the start of the pulse until the point at which the subsequent ADC detects peak amplitude and closes its linear gate (Fig. 22). Note that the period T_P receives double weighting because a second pulse that arrives during this period also causes the first pulse to be rejected due to pile-up. The dead time is an extending dead time, and the unpiled-up output rate r_o for the amplifier is related to the input counting rate r_i from the detector by the throughput equation

$$r_o = r_i \exp[-r_i (T_P + T_W)] \ .$$

Figure 24 illustrates this equation for amplifier shaping time constants ranging from 0.5 to 10 µs. The amplifier output counting rate reaches its maximum when $r_i = 1/T_D$. It is clear from Fig. 24 that higher counting rates require shorter shaping time constants.

When the ADC is part of the spectroscopy system, the dead times of the amplifier and the ADC are in series. The combination of the amplifier extending dead time followed by ADC non-extending dead time T_M yields a throughput described by

$$r_o = \frac{r_i}{\exp[r_i(T_W + T_P)] + r_i\,[T_M - (T_W - T_P)]\ U\,[\,T_M - (T_W - T_P)]}$$

where $U\,[T_M - (T_W - T_P)]$ is a unit step function that changes value from 0 to 1 when T_M is greater than $(T_W - T_P)$.

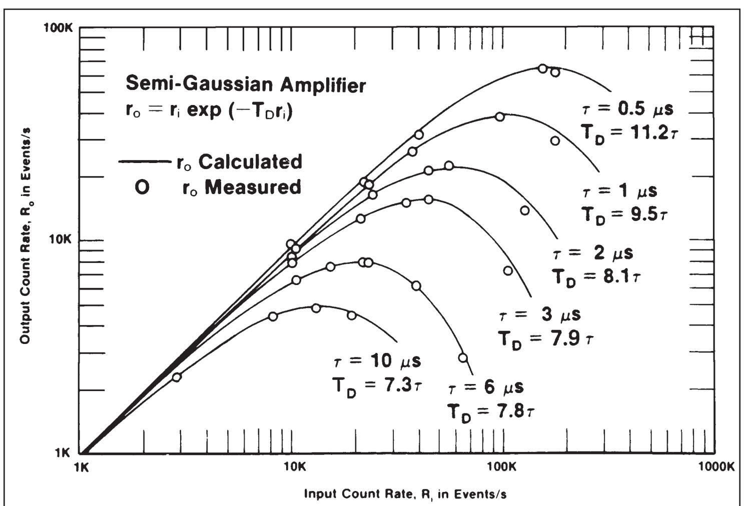


Fig. 24. Plot of the Unpiled-Up Amplifier Output Rate as a Function of Input Rate for Six Values of Shaping Time Constants.

Digital Signal Processing (DSP)

In the previous few pages the functions incorporated in linear pulse-shaping amplifiers have been described in terms of analog signal processing components. Alternatively, most of these functions can be implemented by means of Digital Signal Processing (DSP). Basically, the DSP method converts the continuous analog signal at the output of the preamplifier to a stream of digital numbers representing the history of the preamplifier output voltage. The technique is implemented by using a flash ADC to repeatedly sample and digitize the preamplifier signal. The constant interval between samples is typically small so that the digital numbers represent the pulse profiles with reasonable accuracy. For every analog pulse processing function in the continuous time domain, one can construct an equivalent function in the discrete time domain of the digital representation. Thus, the equivalent signal processing can be implemented in a computer. Because software computation would be too slow to keep up with the data rates, the processing is done in a hardware circuit known as a DSP (Digital Signal Processor).

Figure 25(A) shows the block diagram of a typical DSP MCA, which is a complete digital signal processing system for gamma-ray spectrometry. The digital signal processing in this system incorporates the low- and high-pass filters, automatic pole-zero adjustment, the baseline restorer, fine gain adjustment, a spectrum stabilizer, and means for measuring and histogramming the amplitudes of the digital pulses. This latter function replaces the multichannel analyzer normally used with analog signal processing.

Figure 25(B) illustrates the typical digital filter response in the DSPEC. The flat top is employed to eliminate the degradation of energy resolution normally caused by the variations in charge collection time in HPGe detectors (ballistic deficit). For very wide pulse widths, the flat top becomes negligible, and the pulse shape approaches a cusp. The cusp is the ideal filter for achieving the optimum signal-to-noise ratio at the noise-corner time constant. A reasonable approximation to the cusp can be readily implemented in digital signal processing, whereas it is virtually impossible to achieve using analog signal processing. The cusp shape can be easily changed to a trapezoid, which yields optimum energy resolution for shaping-time constants that are small compared to the noise-corner time constant (for higher counting rates).

The benefits of digital signal processing are:

- greater flexibility in realizing the optimum pulse-shaping filter over the entire range of shaping time constants,
- improved temperature stability,
- ballistic deficit correction at short shaping time constants and optimum energy resolution at long shaping time constants, and
- computer automated optimization of the pulse-shaping filter to suit the detector and data acquisition conditions.

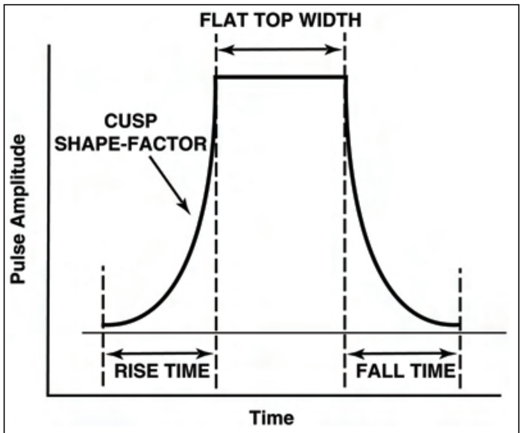


Fig. 25(B). Digital Filter Response.

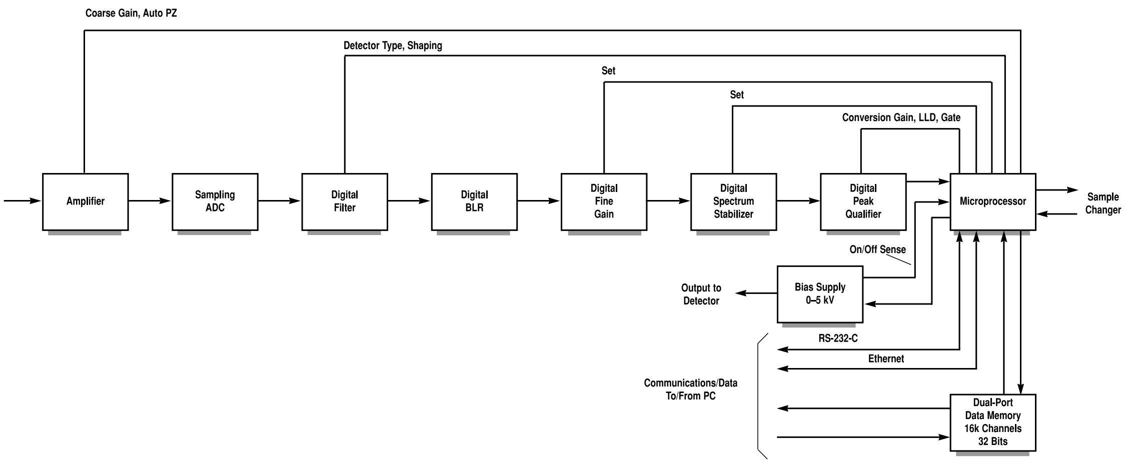


Fig. 25(A). DSPEC Block Diagram.

Delay Amplifiers

Frequently, it is necessary to delay an analog signal to align its arrival with the arrival time of a gating logic signal. This is the function of a delay amplifier. It provides an adjustable delay of the analog signal while preserving the shape and amplitude of the analog pulse. Figure 26 is a typical example involving a coincidence measurement between two detectors. Coincidence timing information is derived from the bipolar zero crossing on the two amplifier outputs using timing single- channel analyzers. The coincidence gating signal would normally arrive at the multichannel analyzer gate input too late to straddle the peak amplitude of the unipolar amplifier output pulse from detector 1. The delay amplifier is used to delay the unipolar output pulse until its peak amplitude is synchronized with the logic pulse at the gate input of the multichannel analyzer.

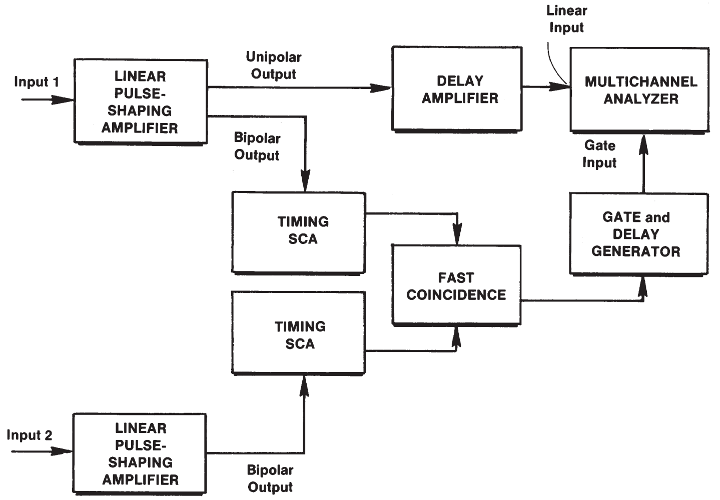


Fig. 26. Use of a Delay Amplifier to Align Analog and Gating Logic Signals.

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