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High-Count-Rate Spectroscopy with Ge Detectors: Quantitative Evaluation of the Performance of High-Rate Systems[1]

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The performance of a high-count-rate system depends on the performance of its components, but cannot be guaranteed by the performance of any particular one. The system must be considered as a number of interdependent components.

A method is presented that allows calculation of system performance from component specifications. Results are shown to be consistent with published manufacturer's data. Contrary to popular belief, it is often the amplifier, rather than the ADC, performance that exerts the greatest influence on maximum system throughput.

Detector size is shown to be of critical importance for high-count-rate applications.

A figure of merit is given by which the detection limit of high-count-rate systems may be compared. Experimental methods for testing system performance are provided.

I. Introduction

There is increasing interest in germanium detector gamma-ray spectroscopy performed at high and ultra-high count rates. Applications exist in fields as diverse as neutron-activation analysis, post-accident sampling, stack monitoring, and fuel-pin scanning. The publication of a recent paper claiming exceptional performance for an ultra-high-count-rate package of instruments is the catalyst for this review of the issues involved in high-count-rate spectroscopy systems and performance criteria for such systems.[2]

Commercial literature on the subject of high-rate spectroscopy is filled with terminology and specifications that are frequently ambiguous, incomplete, and confusing. A recent paper provides a useful introduction to high-count-rate Ge gamma-ray spectroscopy.[3] This current document considers the key issues further and then suggests methods for selecting the best instrumentation solution for a particular measurement problem.

II. Why Choose a High-Rate Gamma-Spectroscopy System?

Before discussing detailed system specifications, it is worth considering some of the typical application issues that lead to consideration of a high-count-rate system. Since operation at high rates can make stringent demands on system performance, a user will not normally choose to operate at high rates simply to "get the data faster;" the reasons to operate at high rates are more fundamental.

Surprising as it may seem, important similarities exist between high- and low-count-rate system requirements. At the low-count rates experienced when performing environmental measurements, the task is to reach the required detection limits in a given time. Shorter counting times and improved detection limits can be achieved by using larger germanium detectors, which by virtue of their higher efficiencies and higher peak-to-Compton ratios, result in improved peak-to-background (signal-to-noise) ratios.

For high-count-rate situations — such as may be encountered in neutron-activation analysis or primary coolant monitoring — one might suppose the situation to be entirely different: that since there is "no shortage of counts," a larger detector is unnecessary. But this is not true if the intensity of the gamma rays of interest is very low compared with the total flux incident on the detector. Just as for environmental measurements, a high signal-to-noise ratio is essential. Achieving this by choosing a large detector may mandate system operation at elevated count rates. A consequential disadvantage of employing this tactic to improve signal-to-noise ratio may be to increase an already high count rate even further, perhaps to a rate at which the system does not function acceptably. A correctly designed system can, however, operate at such count rates and can accommodate this situation.

Another measurement problem that demands a system capable of operating at high count rates is one requiring wide dynamic range counting. For this purpose, the system must operate "acceptably" over many decades of input count rate without physical or electronic adjustment. An application example is stack monitoring with the requirement of being able to handle a potential emergency release.

In each instance, be it wide dynamic range counting, or searching for trace elements in primary coolant, as much "good data" as possible must be obtained. The quality of the result can always be enhanced by improving the signal-to-noise ratio in the data.

III. The High-Rate Spectroscopy System, Component By Component

A. The Detector Element

It is commonly believed that using a small detector (small planar or small coaxial) is a good way to reduce excessively-high count rate. *This is not correct*. It is generally better to collimate a larger detector than to choose a low-efficiency detector with its relatively poor peak-to-Compton ratio. Choosing a small detector — which is not reasonably "black" to higher energies* — as a means of reducing count rate, results in the following performance disadvantages when compared to a collimated larger coaxial detector:

*The special case, as in certain Pu safeguard measurements, of quantifying a single low-energy line close to nearby low-energy lines, but in the absence of any high-energy interference, demands only exceptionally good energy resolution; in this case, a small planar detector is probably the best choice.

Above several hundred keV (say, above 500 keV) the spectrum from the smaller detector will have fewer photopeak counts than a larger collimated detector *operating at the same total count-rate*. In addition, in the low-energy region of the spectrum, the spectrum from the smaller detector will have more counts in the background due to Compton scatter from higher-energy photopeak events (i.e., those photons that have not been fully absorbed).

This may be summarized by saying that the smaller detector has smaller peaks towards higher energies and higher background towards lower energies.

The benefits of this strategy can be seen clearly in Figure 1A (data from Keyser[4]) that shows two spectra, *taken at identical count rates for the same sources*. One spectrum was taken with an uncollimated 12% coaxial detector, and the other was taken with a collimated 120% coaxial detector. The improvement in signal-to-noise ratio is clear. Above approximately 500 keV, there is a slightly higher background from the 120% detector, but this is accompanied by a dramatic increase in the height of the photopeaks with increasing energy. Figure 1B shows that the peaks at 1173 and 1332 keV from ^{60}Co are 2.5 to 3 times larger with the 120% detector. Below approximately 500 keV, the background from the larger detector is *lower*, the difference increasing with decreasing energy. This is due to the higher peak-to-Compton ratio of the larger detector. At energies below approximately 500 keV, the larger detector has approximately the same photopeak efficiency as the smaller detector, since it is “black” to radiation of that energy, but the lower background still implies a worthwhile improvement in signal-to-noise ratio.

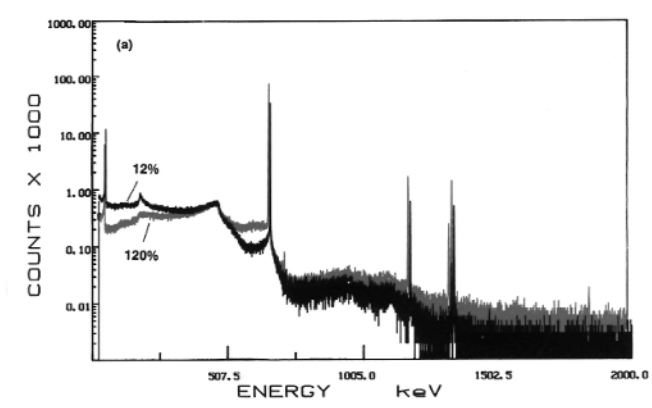

Figure 1A. Comparison of Spectra Taken under Conditions of Identical Count Rate and Count Time for a 12% Uncollimated P-Type Detector and a 120% P-type Detector Collimated to Produce the Same Count Rate. The larger detector shows larger high-energy photopeaks and generally lower low-energy background.

B. Detector Preamplifier: Choice of Feedback Technique

A detector preamplifier converts the charge pulse from the detector crystal into a voltage pulse to be processed by the main amplifier. Each successive charge pulse into the preamplifier causes a step-function increase in the output voltage. To prevent saturation, a feedback circuit restores the preamplifier output to the baseline.

The feedback circuit operates differently in the two most common types of Ge detector charge-sensitive preamplifiers —resistive feedback and transistor-reset (TRP) — used for gamma-ray spectroscopy. In the TRP circuit, the transistor acts as a switch across the feedback capacitor, which periodically resets the preamplifier output to ground. In the resistive feedback preamplifier, an exponential “tail” is produced, as the charge decays through the feedback circuit. This exponential decay from the resistive feedback preamplifier necessitates making a critical “pole-zero” amplifier adjustment in order to obtain symmetrical spectral peak shapes. This adjustment is difficult to make sufficiently accurate for high count rates. When a transistor-reset preamplifier is employed, this adjustment is unnecessary.

1. Preamplifier Energy-Rate Product Concerns

The energy-rate product is the number of photons per second at each energy in the radiation flux striking the detector, multiplied by the energy that each photon deposits in the detector, and the result summed for all energies. The energy-rate product, which is usually stated in MeV per second, represents the rate at which energy is being absorbed by the detector. This translates to the rate at which charge is being generated in the detector.

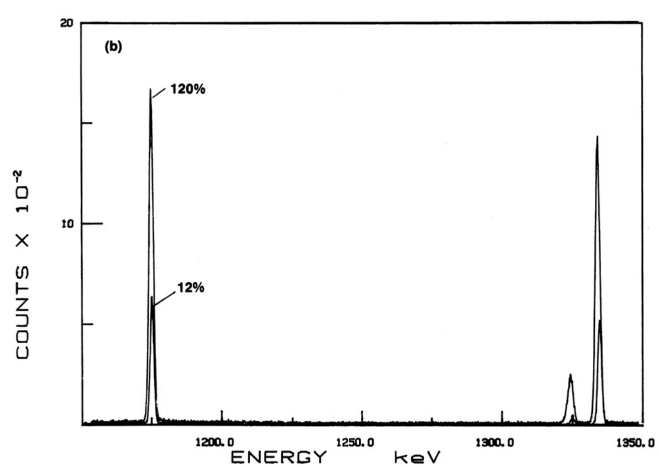

Figure 1B. 1173- and 1332-keV ^{60}Co Peaks from Figure 1A, Shown on Linear Scale. Peak net areas from the 120% detector are approximately three times as large as those from the 12% detector. (The peak at 1321 keV is a sum peak from ^{137}Cs in the sample.)

In the case of a resistive feedback preamplifier, there is a maximum energy-rate product, at which the feedback circuit is only just able to keep up with the charge being delivered. At rates above this, the preamplifier saturates, or "locks up," as the dc level of the preamplifier output increases to the maximum voltage limit of the preamplifier power supply. The preamplifier operates again when the input energy rate falls below this maximum. Typical maximum energy-rate products for commercial detectors, with this type of preamplifier, are in the range from 120,000 to 180,000 MeV per second. Reducing the value of the feedback resistor increases the maximum energy-rate product, but at the cost of increased noise and corresponding resolution degradation.

For example: one manufacturer states that for a detector feedback resistor reduced by a factor of ten to enhance the maximum energy-rate product, the ^{60}Co resolution typically degrades from 1.81 to 2.13 keV. (The capacitance and inductance of what is supposed to be a purely resistive element makes it extremely difficult for a detector manufacturer to use a low-value feedback resistor and still maintain good and constant resolution from low to high count rates.)

A ten-fold increase in maximum preamplifier energy-rate does not necessarily imply a ten-fold increase in system throughput if, as is often the case, the amplifier is the limiting component from a throughput point of view. There will be no improvement in system throughput from a misguided sacrifice of resolution and resolution stability intended to increase the preamplifier maximum output rate beyond that which the amplifier can handle.

2. The Benefits of the Transistor Reset Preamplifier

Table 1 summarizes the relative merits of the TRP versus the resistive-feedback preamplifier. The TRP does not suffer from the inductive effects associated with low-value feedback resistors, and is free from pole-zero adjustment requirements. *The TRP is non-paralyzable to input count rates more than ten times higher than for a standard resistive preamplifier*.

Table 1. Transistor-Reset Vs. Resistive Feedback Preamplifier.

Transistor-Reset Preamplifier	Resistive Feedback Preamplifier
No PZ adjustment required	PZ adjustment needed (difficult at high rates)
Widest dynamic range of count rates (maximum energy-rate >1 GeV/s)	Maximum energy-rate limited by feedback resistor
Better resolution at all count rates than with low-value resistor technique	Low-value resistor for high rates substantially degrades resolution, especially at low rates, and inductive effects cause resolution to degrade further with increasing count rates
Reset pulse adds to system dead time	No reset pulse needed

C. Amplifiers

To achieve optimum amplifier performance at maximum throughput, there are three issues that must be considered.

1. Peak-Shape Deterioration at Short Shaping Time Constants

In order to keep pulse-processing times to a minimum at very high count rates, short amplifier shaping times are desirable. However, in the case of germanium detectors, there is in practice a lower limit to the shaping time, below which catastrophic resolution degradation will occur. This is due to a phenomenon known as ballistic deficit. Ballistic deficit arises from the charge collection time variations associated with events detected in different locations in the detector. Slower rise-time charge pulses may not be fully integrated at short shaping times, leading to the characteristic appearance of a "low-energy tail" on the photopeaks with energy FWHM (full-width-at-half-maximum) resolution degraded to several keV.

The larger the detector diameter, the greater the ballistic deficit, and the shorter the amplifier shaping time, the worse the effect of ballistic deficit on the resolution. Figure 2 shows that, even for a relatively small coaxial detector (efficiency 13.3%), ballistic deficit effects can be significant.

There are two approaches to ballistic deficit compensation. A ballistic deficit corrector uses an analog computer to calculate a pulse-by-pulse correction to the pulse height based on the pulse rise time.[5] This works well for large detectors and up to moderately high count rates, but at ultra-high count rates, the method is not sufficiently accurate.[6] The gated integrator (GI) amplifier approach is superior at very short shaping times.

The operation of the gated integrator for high-count-rate spectroscopy is described in detail elsewhere (see, for example, Knoll[7]). A short summary suffices here.

Figure 3 (upper) shows a typical semi-Gaussian amplifier output pulse. T_p is the time to peak, and T_w is the total pulse width. At the 0.1% amplitude points, T_p is typically equal to 2.0 times the amplifier shaping time, and T_w equals 6.3 times the shaping time. The period $T_w - T_p$, beyond the peak of the

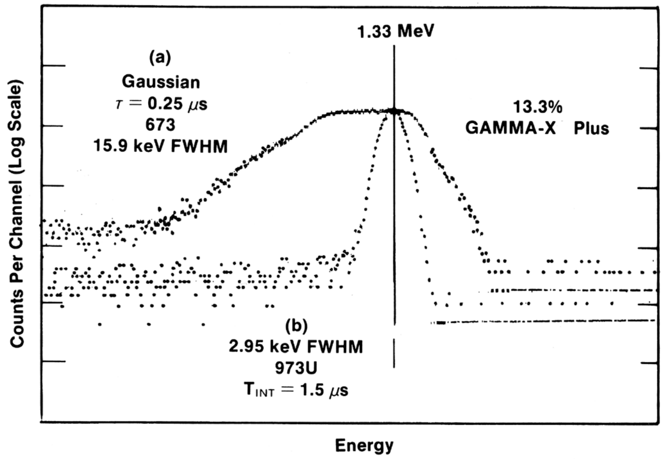

Figure 2. The 1.33-MeV Gamma-Ray Peak from a ^{60}Co Source, acquired from a germanium detector using (a) an amplifier with a Gaussian pulse shape and a 0.25-µs shaping time constant, and (b) the Model 973U gated integrator amplifier with a 1.5-µs integration time. Vertical scales normalized for peak shape comparison.

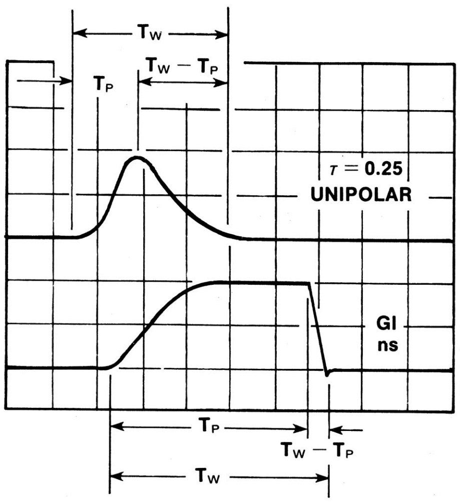

Figure 3. Output Pulse Shapes from (Upper) Semi-Gaussian, and (Lower) Gated Integrator Amplifier.

amplifier output pulse constitutes part of the total dead time for that pulse, but occurs after the peak of the pulse itself, and therefore any charge collected in the detector in that period is lost. Thus, in the semi-Gaussian amplifier, only 2.0/6.3 = 32% of the pulse duration is available for integration of the charge from the detector. Sixty-eight percent is "wasted time" since the next pulse may not start until the amplifier has returned to the baseline. In the case of a Gaussian or triangular filter amplifier, if the ADC conversion time is less than this "wasted time" during which the amplifier output falls back to the baseline, no extra dead time is added due to the ADC, since its conversion time is "masked" by the amplifier dead time beyond peak detect (equal to $T_W - T_P$).

The GI design, originally proposed by Radeka, greatly reduces the effects of ballistic deficit in the following way:

The GI is often added as an extra state to a Gaussian amplifier, which is then said to provide a "prefilter" to the GI. It integrates the prefilter output for a predetermined interval beyond the peak of the prefilter pulse, long enough to gather most of the charge from the detector, thereby reducing the effect of ballistic deficit. Once the predetermined interval is over, the gate is abruptly closed. The resulting output pulse [Figure 3 (lower)] peaks very late, meaning the wasted period $T_W - T_P$ is very short.

Thus in the case of the gated integrator, much more of the pulse duration is used to integrate charge from the detector. This allows good resolution to be obtained at reduced shaping times, minimizing the system dead-time-per-pulse.

In theory, the optimum prefilter pulse to employ in front of the gated integrator would be rectangular in shape, and could be produced by delay-line shaping. Delay lines do not, however, have the required stability for high-rate applications.

The ORTEC Models 973U and 973 gated integrators used in the MERCURY™ high-rate systems employ a quasi-rectangular prefilter design (also called the "Camel filter").[2] This results in 16% less noise than an equivalent Gaussian prefilter, but with none of the peak stability problems inherent in delay-line shaping.

Figure 2 illustrates the extent to which the problems caused by ballistic deficit can be alleviated by the use of a gated integrator.

2. Pile-Up and Throughput

Pile-up is the name given to the summing process that occurs when two or more pulses from the detector occur so close in time that they are treated by the amplifier-ADC circuitry as if they were one pulse. This "composite pulse" is then added to memory in a channel corresponding to the sum of the pulse heights that produced it.

The phenomenon of pile-up degrades the signal-to-noise ratio in a spectrum in two ways: by removing counts from the photopeaks and by adding counts into the background. As the total count-rate increases, pile-up increases, and it is the pile-up process that usually limits ultimate system throughput.

An ideal pile-up rejecter improves the signal-to-noise ratio by detecting all pile-up events and removing them from the spectrum, but it cannot restore the throughput lost as a result of discarding piled-up events (Figure 4). A *real* pile-up rejecter cannot reject *all* piled-up events.

Because piled-up events are undesired background, the only meaningful way to measure throughput is by the amount of "pile-up-free" data collected, that is, with the pile-up rejecter connected!

To get the best possible counting statistics in the shortest possible acquisition time, the system should be operated at the

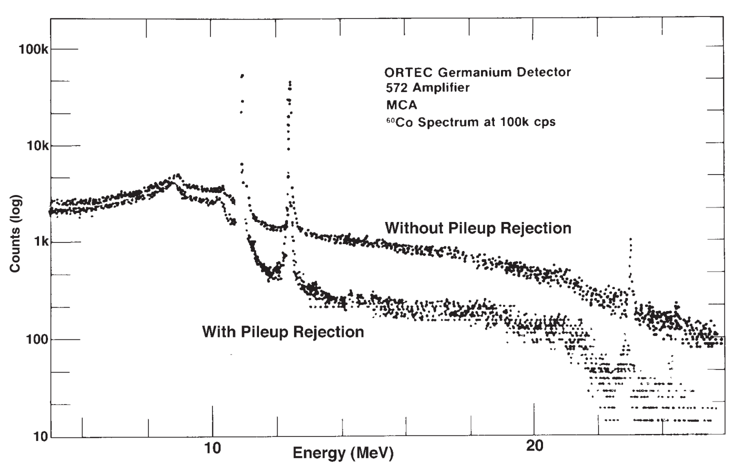

Figure 4. Reduction of High-Energy Background Due to Pile-Up by the Use of a Pile-Up Rejecter.

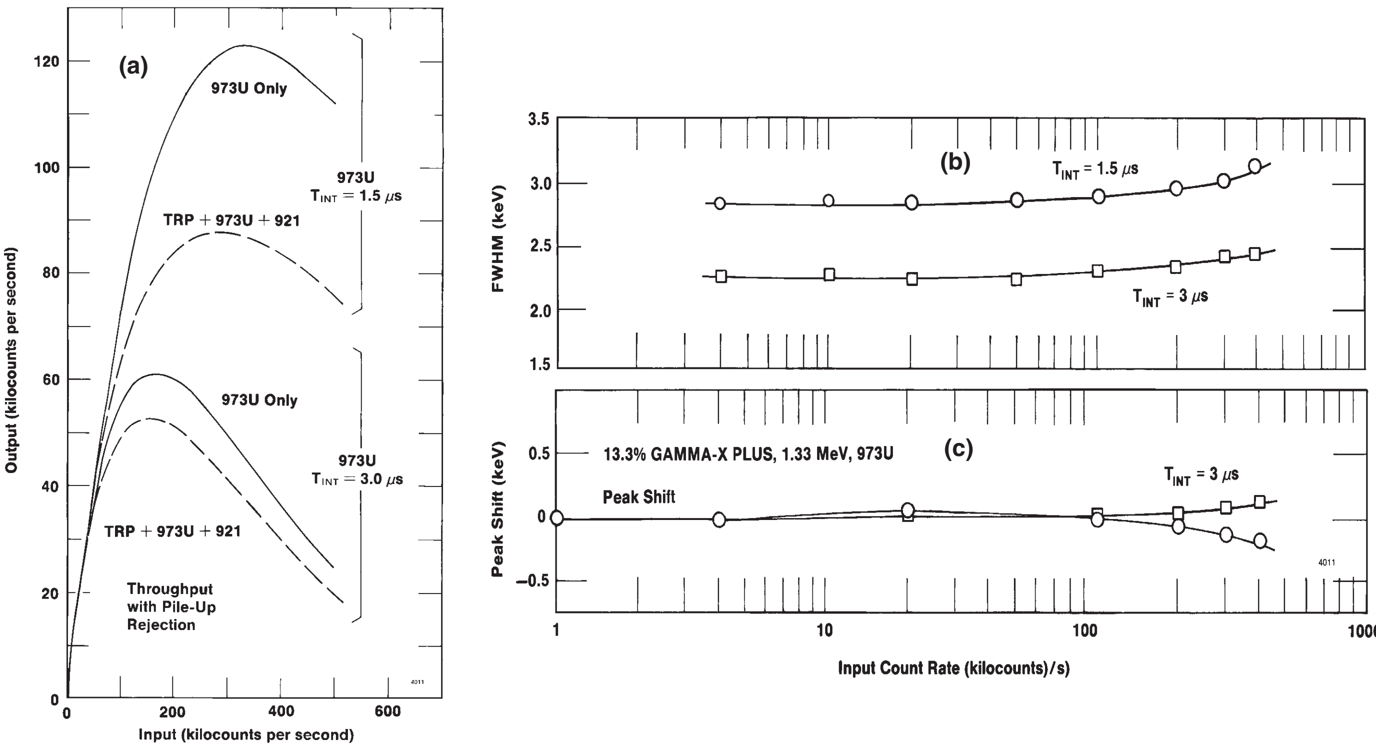

Figure 5. The Throughput-Resolution Trade-Off. (a) Throughput attainable with ORTEC Model 973U gated integrator amplifier at 1.5-μs and 3-μs integration times; (b) the corresponding resolution curves; and (c) the corresponding stability curves.

input count rate at which the throughput maximum occurs.

The extent of pile-up can be reduced by selection of narrower amplifier pulse-widths (shorter shaping times). This action implies a trade-off between system energy resolution and throughput of pile-up-free data (Figure 5). Relaxing the demands on resolution will generally allow shorter shaping times to be used, with consequent increase in pile-up-free throughput. This will of itself improve the detection limit, but the associated resolution deterioration may result in certain peaks becoming constituents of multiplets, which process tends to degrade the detection limit. For many applications, an acceptable compromise can be reached.

3. Amplifier Stability at Elevated Count Rates

Stability of peak shape and position is as important as good energy resolution. If a system is to provide good high-count-rate results, it is vital that the peaks in the spectrum do not significantly change shape (width) or position as the count rate changes. In the context, “significantly” must be measured in terms of peak width (FWHM). Ultimately, nuclide identification and quantification depends on the accuracy to which the calibrations for energy versus channel, FWHM versus energy, and efficiency versus energy are known for the sample spectrum. For example, with the commonly employed nuclide “identification window” of 0.5 • FWHM, the system must be stable enough that peaks do not drift by greater than 0.5 • FWHM over the range of count-rates to be encountered. If this cannot be achieved, then more complex techniques may be needed, such as sample dilution to reduce total count rates or recalibration at different count rates. This additional complexity is undesirable from many standpoints, including operational cost.

Generally, shorter shaping time constants preserve peak shape and position stability to higher counting rates, although this depends to a large extent on the sophistication of the amplifier and baseline restorer design. The preamplifier and ADC can also contribute to peak shape and position degradation at high counting rates.

D. ADCs

1. Speed and Conversion Time

Two types of ADCs, Wilkinson and successive-approximation (called fixed-conversion time, FCT, in Reference 3), are commonly used in nuclear spectroscopy systems. In a Wilkinson ADC, the incoming pulse amplitude is stored on a capacitor. An internal oscillator is used to measure the time it takes for the capacitor to discharge to zero volts. The speed of this type of ADC is usually quoted in terms of the oscillator speed in MHz. The actual conversion time in microseconds depends on the pulse-height being measured, and the number of channels in the spectrum (i.e., the “conversion gain”).

A successive-approximation ADC uses a system of multiple comparators for determining pulse amplitude to the desired precision. In this case, the conversion time is independent of pulse height. The speed of the successive-approximation type of ADC is usually quoted in microseconds.

A typical Wilkinson ADC speed is 100 MHz, whereas for the successive-approximation type, 25 µs is comparable. A 100-MHz Wilkinson ADC will convert “small” pulses in less than 25 µs and “large” pulses in more than 25 µs. For example, in a commercial Wilkinson ADC with a 100-MHz clock speed, channel 1024 is converted in ~10 µs and channel 8192 in ~80 µs. It is therefore possible to operate a Wilkinson ADC at low resolution (e.g., 1024 channels, as opposed to the usual 8192), and thereby gain higher conversion speed.

In high-rate spectroscopy, 450 MHz and 2 µs, respectively, would generally be considered “fast” for these two types of ADCs. A 450-MHz Wilkinson ADC will convert channel 1024 in ~3 µs and channel 4096 in 10 µs. A 2-µs successive-approximation type will convert any channel in its range in 2 µs.

At first glance, it might seem reasonable that an ADC with conversion time of, say 5 µs, could process data at 200,000 counts per second. This would be true, however, only if the input pulses were equally spaced in time (never the case with the random decays coming from a radioactive source).

Consideration of ADC speed alone can be misleading; there are always additional dead times during ADC readout or during addition of a count to memory. Even if the ADC conversion and MCA store-to-memory were infinitely fast, the system throughput performance would still be subject to a fundamental limit; this is due to the detector charge-collection time which places a limit on the amplifier throughput.

2. Dead-Time Correction

The object of any spectroscopic measurement is the accurate determination of net count rates in the full-energy peaks of interest. This implies an accurate knowledge of, and compensation for, all sources of dead time within the system. There are many sources of dead time to consider in a given system: preamplifier reset, amplifier busy, pile-up rejection, ADC busy, MCA store-to-memory. These combine to produce an overall system dead time, which must be carefully taken into consideration if accurate net count rates are to be determined.

In the industry, there are different methods employed for dead-time correction. No standards exist for the timing, polarity, and duration of intermodule signals, such as ADC Busy or Pile-Up Reject. Because different manufacturers use different methods, a system constructed of components from different suppliers is likely to produce unsatisfactory results. The Gedcke-Hale method (see Appendix B3) is employed by ORTEC in its high-performance systems to provide live times accurate to within 3%, up to input count rates of 400k cps.

The subject of dead-time correction has often generated animated debate. The relative advantages of several methods, discussed at some length in Reference 9, pp. 262–276, are being further investigated and will be reported on in a future communication.

IV. System Considerations in Detail

Introduction: How Preamplifier, Amplifier, and ADC Dead Times Combine with MCA Store-to-Memory Times to Limit System Maximum Throughput

Jenkins et al.[9] have derived a general formula that gives the relationship between pile-up-free counts per second stored into memory, for an amplifier-ADC combination as a function of input count-rate:

$$r_o = \frac{r_i}{\exp\left[r_i(T_w+T_p)+r_i[T_m-(T_w-T_p)]\bullet U[T_m-(T_w-T_p)]\right]} \qquad (1)$$

where

r_o = pile-up-free throughput rate stored to memory,

r_i = input count rate from the preamplifier

T_w = width of the output pulse from the amplifier,

T_p = time taken for the amplifier output pulse to reach its peak,

T_m = digitizing time for the ADC and memory, and includes not only the conversion time of the ADC, but also the time taken to add one to memory,

$U[T_m-(T_w-T_p)]$ = a step function that changes from zero to 1 when $T_m>(T_w-T_p)$.

An extra term that must be added to this formula to take account of preamplifier effects is discussed in part B of this section.

A. Amplifier Throughput Limits

Case 1: Gaussian and Triangular Filter Amplifiers

Examination of Equation (1) shows that for an amplifier with a Gaussian or triangular output pulse, there can be cases where the step function term goes to zero, that is when the ADC conversion time T_m is less than the amplifier output pulse fall time T_w-T_p. For such systems, the ADC conversion and transfer to memory always occurs before the amplifier pulse has returned to the baseline and the amplifier is ready to process another pulse. The ADC therefore adds no more dead time. *In this situation, using an ADC that can convert and store a pulse faster than the interval $T_w–T_p$ produces no improvement in throughput whatsoever.*

When the ADC is adding no extra dead time, Equation (1) reduces to:

$$r_o = \frac{r_i}{\exp\,[r_i\,(T_w+T_p)]} \tag{2}$$

The maximum throughput $r_{o_{max}}$ occurs when r_i equals $1/(T_w+T_p)$, and is equal to $1/2.7(T_w+T_p)$.

Table 2 lists, for different amplifier shaping times, the largest value of T_w-T_p (shown as "ADC Maximum Conversion Time") for which Equation 2 still holds. The corresponding maximum throughput obtained when the ADC conversion time is less than this value is also given. Reducing the ADC conversion time below the given value yields no extra throughput benefit for these settings. Since the amplifier shaping time chosen is usually dictated by the resolution requirement, these values represent fundamental limits on throughput and on the requirement for ADC speed with Gaussian amplifiers. Similar numbers apply in the case of triangular filters.

Table 2. Fundamental Limits on Throughput and on the Requirement for ADC Speed with Gaussian Amplifiers.

Gaussian Shaping Time (μs)	ADC Max Conversion Time (μs)	Max Throughput (Pile-Up-Free) (k cps)
10	43	4.4
6	26	7.4
4	17	11.1
3	13	15
2	9	22
1	4	44
0.5	2	88

For almost all applications using coaxial germanium detectors, shaping times of less than 2 μs will produce unacceptable resolution due to ballistic deficit; *this implies that 22 k cps is the maximum practically achievable throughput with Gaussian amplifiers and coaxial germanium detectors — however fast the ADC.*

Figure 6 shows maximum pile-up-free throughput, for a variety of ADC dead times (conversion plus transfer-to-memory), versus Gaussian shaping time and vs. amplifier dead time per pulse.

Case 2: Gated Integrator Amplifiers

Equation (1) behaves differently in the gated integrator case: the amplifier pulse-width is essentially equal to the time to peak, so the step function $\{U[T_m\,(T_w-T_p)]\}$ is never zero. This means that a faster ADC will always increase throughput slightly, but with diminishing returns.

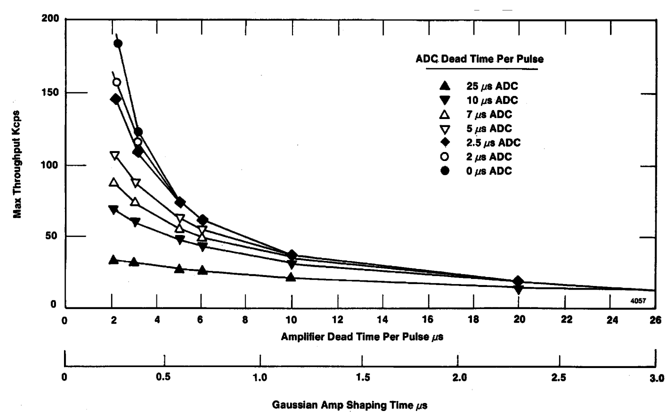

Figure 6. Maximum Throughput [Calculated, Ignoring TRP Losses, from Equation (1)] for Different ADC Speeds vs. Gaussian Shaping Time.

For the gated integrator, the integration time $T_i = T_p \cong T_w$. The throughput rate r_o has its maximum value $r_{o_{max}}$ when the input rate equals $1/(T_w+T_p)$. Substituting into Equation (1) for T_i and for r_i at $r_{o_{max}}$ gives:

$$r_{o\,max} = \frac{1}{5.4T_i+T_m} \tag{3}$$

This formula sheds some interesting light on the performance of practical, gated-integrator systems.

(a) The Gated Integrator with an Infinitely Fast ADC

Let us consider the case of the infinitely fast ADC, that is, T_m = zero. For GIs with Gaussian prefilters, a 0.25-μs shaping time means an integration time of around 10 times the Gaussian prefilter integration time. So for this case, $r_{o\,max}$ is approximately 74k cps. This is the so-called gated-integrator limit, and agrees well with published commercial data on gated integrators with Gaussian prefilters.[10,11,12]

The prefilter design of the ORTEC 973 and 973U delivers excellent resolution, even at shaping times shorter than those available on amplifiers with Gaussian prefilters (see Figure 7).[2]

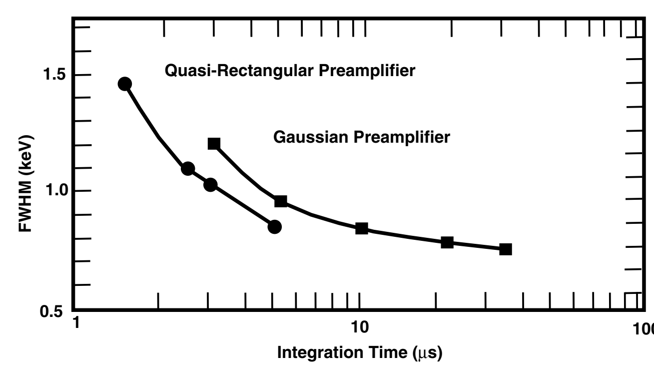

Figure 7. Noise vs. Integration Time for a Gated Integrator with a Gaussian Prefilter and a Quasi-Rectangular Prefilter.

The Model 973U 1.5-μs pulse "camel" prefilter integration time results in a GI limit of 123k cps, which again is in good agreement with published specifications, and is substantially higher than the maximum throughput typically obtained by using a gated integrator with a Gaussian prefilter.

(b) Gated Integrator Coupled to a Real ADC

When a real, rather than an infinitely fast, ADC is added, the following facts become apparent from Equation (3): once the total ADC dead time (T_m), including the MCA store-to-memory time, is much less than the contribution from the GI ($5.4T_i$), there is little to be gained in throughput from using an even faster ADC.

The ORTEC 973U, with its 1.5-μs minimum integration time, is the fastest currently available GI. The amplifier contribution to the denominator of Equation (3) in this case is 5.4 • 1.5 = 8.1 μs. Therefore, so long as the ADC conversion time, including transfer-to-memory, is much less than this, the ADC's effect on maximum throughput is small.

In the ORTEC 921 multichannel buffer, the ADC dead-time (including transfer-to-memory) is, with a conversion gain of 16k channels, less than 2 μs*. Therefore, for the combination of the Model 921 and the Model 973U at 1.5-μs integration time (the 973U's minimum allowable setting), Equation (3) gives:

$$r_{o\,max} = \frac{1}{(8.1+2) \bullet 10^{-6}} = 99\text{k cps} \qquad (4)$$

Improving the ADC speed by a factor of 2 to 1 μs, would only increase the overall throughput by an additional 11% (10k cps).

In the case of the Canberra 2024, coupled to the ND582 ADC,[10] Equation (3) gives, for the maximum throughput:**

$$r_{o\,max} = \frac{1}{(13.5+2.5)} \bullet 10^{-6} s = 62.5\text{k cps} \qquad (5)$$

The 99k and 62.5k cps values are theoretical maximum throughputs for the particular ADC attached to the particular gated integrator. No allowance is made for losses in throughput due to the presence of a TRP, and it is assumed that T_m includes the transfer-to-memory time. The 62.5k calculated is in reasonable agreement with the 66k measured throughput reported in Reference 3.

Figure 8 shows maximum pile-up-free throughput, for a variety of ADC dead times per pulse (ADC conversion plus transfer-to-memory times) versus GI integration times. For a GI with Gaussian prefilter, the appropriate integration time is given by 10 times the selected shaping time of the prefilter. The "0 μs ADC" curve shows the GI limit. At 1.5-μs integration time, the increase in throughput gained by reducing the ADC time from 2 μs to zero is approximately the same percentage as gained in going from 5 to 2 μs.

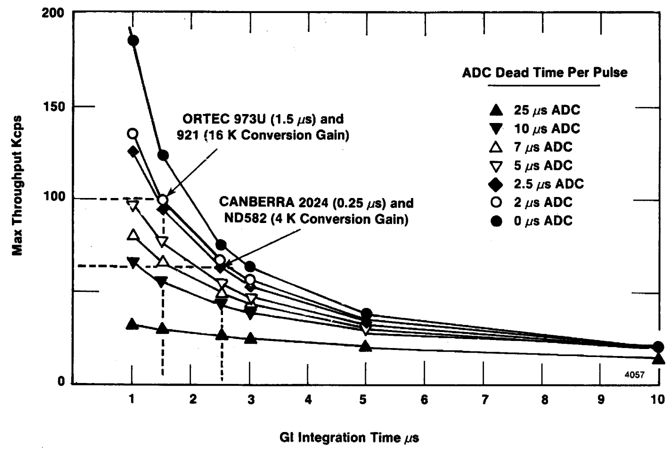

Figure. 8. Maximum Pile-Up-Free Throughput [Calculated, Ignoring TRP Losses, from Equation (1)] for Different ADC Speeds vs. Gated Integrator Time.

B. Additional Dead Time Due to Transistor Reset Preamplifier

Although use of the TRP allows more than an order of magnitude higher count rate and also avoids the resolution problem inherent in the choice of low-value feedback resistors, it has the disadvantage that the TRP reset period contributes to system dead time. A formula for the additional dead time due to TRP resets was originally derived by Britton et al.[13] The TRP dead time adds a multiplicative term to Equations (2) and (3) to give for the maximum throughput when a TRP is employed:

$$r^{TRP}_{o\,max} = r_{o\,max} \quad [1-\%DT(reset)/100] \qquad (6)$$

Where

$$\%DT(reset) = \frac{T_{inh} \bullet 100\%}{[V_{ramp}/(ER \bullet CG)]+T_2+T_3}$$

and

- T_{inh} = time from the initiation of a reset pulse until the spectroscopy amplifier is ready to process another pulse,
- V_{ramp} = preamplifier output voltage that triggers the reset cycle,
- ER = energy-rate product at the input,
- T_2 = delay time between the preamplifier output exceeding V_{ramp} and starting the reset process,
- T_3 = time required for the TRP output to return to baseline after the reset,
- CG = the preamplifier conversion gain.

* This performance (16k-channel resolution, 2-μs ADC dead timer per pulse) may be compared to that reported in Reference 3. (The 1.5-μs FCT referred to is actually 2.5 μs dead time per pulse.)

** ND 582 dead time per pulse = 2.5 μs (Model 582 data sheet). Assumes CI 2024 is set to the minimum allowable setting of 0.25 μs integration time. No allowance has been made for any additional transfer-to-memory time.

For the system reported by Simpson et al.,[2] using a ^{60}Co source as input, the TRP losses were about 12%. This reduces the value of 99k cps given by Equation (3) for the MERCURY system to 87k cps, in good agreement with the experimentally reported value of 86k cps.

V. Detection Limits at the Point of Maximum Throughput

It is generally the case that whatever the value of the maximum throughput, one should attempt to operate the system at the maximum throughput point, because that is the point that data are stored to memory at the fastest possible rate, leading to the shortest acquisition times for a given detection limit (MDA).

If the counting rate is too high, a means of reducing it is needed to achieve maximum throughput. As stated earlier, collimation of a large detector is generally a better strategy than selection of an uncollimated small detector as a means of reducing count rate to the maximum throughput point.

Equations (3) and (6) give the maximum throughput in counts per second into memory for a TRP-GI high-count-rate system. Keyser et al.[14] employ the concept of "Relative MDA" for comparing detectors of different efficiencies in low-count-rate applications. The relative MDA, MDA_R, is given in Reference 14 as:

$$MDA_R = \frac{[R(E_1)B(E_1)]^{1/2}}{\varepsilon(E_1)} \tag{7}$$

where

$R(E_1)$ = detector resolution at energy E_1,

$B(E_1)$ = detector background at energy E_1,

$\varepsilon(E_1)$ = detector photopeak efficiency at energy E_1.

In the high-rate domain, it is detector resolution, photopeak count rate, and background count rate which determine MDA. At low count-rate, photopeak count rate is simply proportional to the detector efficiency at the energy in question.

At the point of maximum throughput, the number of counts in the photopeak is determined by the total system throughput and also by the signal-to-noise ratio in the detector. The efficiency of the detector is irrelevant since the system is processing counts at its maximum rate. It is said the be "throughput limited."

Consider a single gamma ray of energy E_1. It is the fraction of events in the photopeak $f(E_1)$ compared to the total number of events in the spectrum (also called the photofraction), which determine how many events will be present in the photopeak in the throughput-limited case. At the point of maximum throughput, we may write:

$$MDA_R = \frac{[R(E_1)B(E_1)]^{1/2}}{f(E_1)\, r_{o\,\max}^{TRP}} \tag{8}$$

However, the photofraction $f(E_1)$ may be shown to be proportional to the product $R(E_1)P(E_1)$, where $P(E_1)$ is the peak-to-Compton ratio of the detector at energy E_1. We may therefore define a "throughput-limited MDA$_R$", MDA_R, as :

$$MDA_R^T = \frac{[R(E_1)B(E_1)]^{1/2}}{R(E_1)P(E_1)\, r_{o\,\max}^{TRP}} = \left[\frac{B(E_1)}{R(E_1)}\right]^{1/2} \bullet \frac{1}{P(E_1)\, r_{o\,\max}^{TRP}} \tag{9}$$

Equation 9 shows that for two systems with identical detectors, but differing in other respects, *the MDA at the point of maximum throughput is reduced (improved) in proportion to the value of the maximum throughput.*

Correspondingly, if two otherwise identical systems differ only in the detector used, the MDA will depend only on the signal-to-noise (peak-to-background) ratios of the detectors. The measurements made in Reference 4 provide an example. A collimated 120% detector was compared to an uncollimated 12% detector on a system where all other parameters were identical, and the system was operated at the same count rate, approximately equal to the point of maximum throughput. These data demonstrated that (for example) at 661 and 1332 keV, the larger detector gave an MDA nearly a factor of two lower than that from the small detector. At 60 keV, the MDA from the two detectors was about the same.

If follows, therefore, that if two systems are compared, one having a maximum throughput ~50% higher than the other (due to superior circuit design) and also containing a large, high peak-to-Compton ratio, collimated detector as opposed to a small uncollimated detector, *the MDA of the high throughput/large detector system can be as much as a factor of 3 lower.*

VI. Conclusions

The suitability of a high-rate spectroscopy system depends on the constraints of the application. The system performance is not guaranteed simply by the performance parameters of one component alone, but rather is a function of the combined performance of all of its components: detector element and preamplifier, amplifiers, ADC and memory. All components must be carefully matched. The important system performance parameters are:

- High signal-to-noise (peak-to-background) ratio from the detector, achieved by choosing a large detector with a high peak-to-Compton ratio, and collimating as necessary;
- High system throughput-to-memory of pile-up-free events, and appropriate resolution/throughput trade-off; this means using a TRP preamplifier, a gated integrator with a "camel" prefilter, and an ADC dead time (conversion pulse transfer-to-memory) of less than 2 μs per pulse;
- System stability (peak position and shape) with varying count rate;
- System dead-time correction accuracy.

Knowledge of the dead times associated with the preamplifier, amplifier, ADC, and transfer-to-memory allows calculation of the maximum system throughput. Knowledge of the detector resolution and peak-to-Compton ratio then allows the quantitative comparison of the MDA at maximum throughput for competing systems.

VII. References

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Appendix B1.

How to Measure System Throughput — A Logical Approach

At present, there is no IEEE (or IEC) standard for performing this measurement. Comparison of results obtained by different methods can be misleading. For example, *disconnection of the pile-up rejecter gives an apparently higher throughput. A more efficient pile-up rejecter shows a lower throughput than an inefficient pile-up rejecter, although the piled-up events themselves are undesirable*. The choice of sources also affects results. Use of a low-energy source means that a TRP resets less often, enhancing the result. A low-energy source results in better performance from a system employing a Wilkinson ADC, whose conversion time is lower for low-energy pulses. Reducing the conversion gain on a Wilkinson ADC increases the measured throughput.

The throughput specifications and curves (Figure 5) relating to the MERCURY system were obtained using a ^{60}Co source. The conversion time for the successive-approximation ADC used in the Model 921 is the same at 16k-channel resolution as it is at 1k-channel resolution!

The method given here is simple and logical, and if followed, allows two systems to be compared in a meaningful way from the standpoint of their maximum ability to store un-piled-up events into memory.

Method

1. Connect the system components according to the manufacturer's cabling diagram. It is extremely important that this step be carried out correctly, since the dead-time correction algorithm employed by the system relies on signals such as Amplifier Busy and Pile-Up Reject to calculate the dead-time correction. Caution: the interconnection of modules from different manufacturers may result in incompatibility problems, due, for example, to different signal polarity or timing.

   Make sure the pileup rejection circuit is turned on and working!
2. Adjust the amplifier according to the manufacturer's instructions.
3. Use a suitable ratemeter (for example the ORTEC Model 661) to measure the input count rate to the amplifier. An input-count-rate-monitor connector is usually provided by the amplifier manufacturer. If a suitable ratemeter is not available, or if no CRM output is provided, the input count rate can be estimated from the total counts in the spectrum divided by the live time. Record the input count rate, the total number of integrated counts in the spectrum, and the real time for various count rates across the count-rate range of interest.
4. Plot integrated counts divided by real time vs. input count rate to obtain the throughput curve.

Appendix B2.

A Simple Method to Determine the Accuracy of Your Dead-Time Correction Circuitry

1. Connect the various components according to the manufacturer's cabling diagram. It is extremely important that this step be carried out correctly, since the dead-time correction algorithm employed by the system relies on signals such as Amplifier Busy and Pile-Up Reject to calculate the dead-time correction. This should be taken as a note of caution in the interconnecting of modules from different manufacturers, since it is possible that expected compatibility may not exist, due to different signal polarity or timing. *Make sure that the pileup rejection circuit is turned on and functioning!*
2. After correct adjustment of the amplifier according to the manufacturer's instructions, set up the system to acquire a ^{60}Co spectrum with the 1.332-MeV peak in approximately channel 7000 (on an 8k-channel system). Employ a ^{60}Co source that produces a dead time of just a few percent. (If necessary, increase the source-to-detector distance.)
3. Accumulate a spectrum until the 1.173-MeV peak has a 10,000-count net area (above background). Divide the net counts by the live time to determine the net count rate. Accept this number as the "true" value. (Do not make the region of interest used to integrate the peak more than a couple of channels wider than the peak itself. Inclusion of too much background can lead to error at high count rates.)
4. Use a lower energy source, such as ^{137}Cs, to repeatedly increase the total count rate. (If a source other than ^{137}Cs is used, choose one that does not produce a sum peak near the reference peak at 1.173 MeV. (That is the reason the 1.332-keV peak of ^{60}Co is not used as the reference, since 2 x 0.662 Mev = 1.324 MeV.)

   Do not move the ^{60}Co source. For each measurement, attempt to acquire approximately the same number of counts in the ^{60}Co peak. Record the total count rate, measured with a ratemeter at the count-rate monitor output of the amplifier, and the apparent net count rate (net counts ÷ the live time) of the ^{60}Co peak.
5. Plot the ratio of (a) the apparent ^{60}Co net count rate divided by the true value versus (b) total count rate. The deviation from a value of unity is the error in the live-time clock.

Appendix B3.

The Gedcke-Hale Live-Time Clock (Dead-Time Correction Method)

An "early" description of the Gedcke-Hale Live-Time Clock can be found on pp. 266–267 of Reference 9.

The Gedcke-Hale Live-Time Clock provides correction for the dead-time losses that occur in the spectroscopy amplifier and the ADC (or MCA). It is intended for use with the unmodified, unipolar, analog output pulses from a spectroscopy amplifier. This includes the unipolar pulse shapes from gated integrators, Gaussian shaping amplifiers, triangular pulse shaping amplifiers, and single-delay-line shaping amplifiers. Although it will work with amplifiers that do not provide Busy or pile-up rejecter (PUR) signals, best accuracy is obtained when the Busy and PUR signals from the amplifier are connected to their respective inputs to the ADC.

The unique benefit of the Gedcke-Hale Live-Time Clock is that it compensates for the dead-time losses caused by pile-up in the shaping amplifier in addition to the dead-time losses caused by the ADC and memory processing times. When the counts in full-energy peak in the energy spectrum are divided by the live time, the resulting counting rate is an accurate estimate of the true counting rate for that gamma-ray energy at the detector output.

The Gedcke-Hale Live-Time Clock (dead-time correction) works as follows. Either the leading edge of the Busy signal from the amplifier, or the amplifier analog output pulse rising above the level of the ADC lower-level discriminator will cause the live-time clock to start counting backwards. The live-time clock turns off when the stretcher in the ADC detects peak amplitude on the amplifier analog pulse, or when a PUR input occurs. The live-time clock resumes counting forward after the following signal conditions are all satisfied: the ADC conversion and readout have been completed, the amplifier has fallen below the ADC lower-level discriminator, and both the Busy and PUR signals from the amplifier have returned to their inactive states. Turning off the live-time clock provides a simple measure of the dead-time interval, while counting backward with the live-time clock multiplies the measurement of the corresponding dead-time interval by a factor of two. The reason for this procedure is as follows: after peak amplitude detection, the ADC closes its linear gate and subsequent pulses cannot be analyzed during the ensuing ADC Busy time. Turning off the live-time clock compensates for the probability of losing subsequent pulses during this ADC dead time. During the interval from the start of the amplifier pulse to peak amplitude detection, the pile-up distortion caused by the arrival of a second pulse causes both the first pulse and the second pulse to be rejected. Subtracting live time during this interval provides the double weighting necessary to compensate for the probability of losing two pulses.