TIME-SAMPLING AMPLIFIER

William Goldsworthy

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A time-sampling amplifier system employing a logic timing approach to input information retention, signal sampling and inhibiting, differentiation, and integration is used to provide excellent spectral resolution at low and high counting rates. The utilization of active, time-constant switching gates in the differentiation circuits of this amplifier provides low attenuation during time sampling, and very rapid base-line restoration after time sampling. The use of two of these synchronized time-sampling gate differentiators, one intermediate in the amplifier system and one at output level, provides the excellent base-line stability needed for high-rate high-resolution work without sacrifice of energy resolution.

Inclusion of a logic program for all pulses being processed provides means whereby measurement of "distressed" events can be inhibited. Measurement of distressed events resulting from after-pulse recovery or stack-up can be easily eliminated by simple logic decisions. By preventing measurement of these distressed events much of the resolution smear usually accompanying high counting rates can be eliminated.

Techniques employed in programming of the preamplifier circuits used in this time-sampling amplifier have permitted application of the spectral subtract function directly to the detector's output signal. This greatly simplifies the inclusion of the spectral subtraction and expansion functions often needed in a high-rate high-resolution-capable linear amplifier system.
RESOLUTION PROBLEMS OF HIGH-RATE COUNTING

Basically, energy resolution in high-rate nuclear spectroscopy systems is limited by the time requirements of making a measurement decision, the interaction of energy-dependent electrical pulses being processed through the system, and any average reference-level shift that might arise from the nonsymmetrical nature of the electrical pulses being processed.

Not only are there high-rate-dependent limitations; there also may be many pulses produced by the nuclear detector itself that should be rejected upon the basis of excessive collection time or erratic operation causes.

Some decision originally has to be made regarding the time to be expended in measuring each nuclear event, since there may be many of these randomly spaced in time. In general a compromise must be made in which a reasonable time period is chosen per event. This time period, for most nuclear spectroscopy systems, turns out to be about 1 µsec, where there is a reasonable compromise between event-produced statistics and system noise. It also follows that the most accurate measurement decision should be attained by maintaining high information density.

In the normal amplifier system the basic time taken per event is largely determined by the differentiation time chosen; and the information is averaged during this time period by integration. Both high- and low-frequency noise present at the detector, in the amplifier's input, and in early amplifier stages, are thus highly attenuated by the band-pass action of the filter so formed by integration and differentiation,
while frequency components comprising the signal of interest suffer only minimal attenuation because of their frequency predominance alignment with the filter. This usual process of integration and differentiation, however, leaves much to be desired with respect to good information density and good time definition.

In an amplifier employing time sampling, both the information density and time definition can be ideal.

Time expended per measured event can be very accurately predetermined, information can be accumulated during this time uniformly and at a high rate, and can be averaged over a clearly defined time period.

Figure 1 compares the time definition and information density characteristics of the conventional and time-sampling techniques.

**Interevent Interference**

A significant factor impairing high-counting-rate energy resolution is prompt interevent interference. This interference, as shown in Fig. 2, can result from prompt base-line disturbances following each pulse, as a consequence of the differentiation process, or by pulse addition (stack-up) within the decision-making time period already predetermined in the amplifier.

A number of techniques are now employed to reduce the effects of prompt base-line disturbances. These all rely upon reduction of the time or amplitude (or both) of these disturbances. Techniques employed include pole-zero cancellation of pulse undershoot, delay-line clipping, multiple delay-line clipping, multiple RC clipping, and switched-time-constant restoration. Each of these means, however, has limitations.
Pole-zero cancellation provides greater susceptibility to low-frequency disturbances originating from power source, other low-frequency interference, and microphonics; multiple clipping schemes generally produce poorer resolution, either through loss of pulse information per unit time in double RC or through addition of nonsynchronous noise without like addition of event-produced information per unit time in multiple line clipping.

Base-line restorers generally suffer from noise-induced offset that results from the rapid modulation of the restoration capacitor's charge by noise. This instantaneous charge modulation appears as a voltage which may either add to or subtract from the entering pulse, depending on its polarity at the time of arrival of the pulse of interest. Where two such active base-line restorers are employed as differentiators in widely different sections of an amplifier system and are accurately time-synchronized, the offset noise modulation is considerably less than the sum of both individually, since the instantaneous noise voltage applied to the second differentiator is altered in phase by the action of the first. In practice the improvement over two nonsynchronous differentiators appears to approximate $\sqrt{2}$, which now compares favorably with the performance obtained from most single-differentiating or single-clipping means. This is significant, since we now have the base-line restoration capability of double clipping while keeping low-noise characteristics of single clipping.

**Stack-Up**

Those interevent interferences which occur during the sampling time of an amplifier and add to each other we call stack-up.
With the inclusion of a logic sequence in the amplifier system, these distressed events can be easily rejected on the basis of shape or timing if the decision-making time can be made to follow the complete time-sampling period.

Spectral Shift

Additional spectral resolution loss and shift at high counting rates can result from the nonsymmetric nature of the event-produced pulses being processed. A long-time average-reference-level shift can occur at an amplifier's output proportional to the average counting rate. This is usually referred to as base-line shift. Pulses measured from this false reference of course measure in error, producing a shift in spectral position. The random nature of nuclear events does, however, lead to some resolution loss as well.

Several means are available to reduce the effects caused by these moderate to long-time variations in reference base line. These are means to improve pulse symmetry, to restore the reference base line by after-event modification of differentiation time, to subtract the error produced as a function of counting rate, or to direct-couple all post differentiation stages.

As has already been pointed out, the use of multiple clipping schemes to improve symmetry and provide more rapid base line restoration, as well as the use of base-line restorers, generally impairs basic energy resolution.

Subtraction of an average error is a quite feasible means of spectral shift correction, but is somewhat difficult to apply to the normal amplifier. Effective direct coupling of all post differentiation
stages also has limitations. It may be quite difficult to accomplish in a practical case in which postdifferentiation gain is moderate to high. Base-line instabilities resulting from dc coupling can produce serious impairment of resolution, and there is no means of suppressing low-frequency disturbances present in earlier amplifier stages.

Although the problem of erratic or spurious pulse formation is a detector and not a rate problem, it would still be desirable to not measure these distressed events. This can be accomplished by rejection on the basis of rise time or shape. Logic circuitry associated with an amplifier system can be very easily programmed also to fulfill this function.

ADVANTAGES OF TIME SAMPLING

Several advantages ensue by employing the time-sampling technique described. These are improved base line restoration, without resolution sacrifice, reduced rise time jitter modulation, simplification of the high rate spectral expansion problem, insensitivity to low frequency interferences and the ability to reject distressed events on a timing or shape basis.

Active Time-Sampling Differentiators

Employment of multiple active time-sampling differentiation appears to be an attractive means of reducing the time necessary for after-pulse base-line recovery and of providing improved stability of the reference base line without impairing basic energy resolution. In this differentiating scheme, as shown in Fig. 3, pulses entering the main time-sampling amplifier initiate a logic sequence which provides for active
operation of the two differentiation circuits used. Before arriving at the first differentiator all pulses are delayed, to permit initiation of the differentiation cycle logic prior to their arrival. After initial differentiation, pulses are further amplified and finally arrive at the second active differentiator, where they are again differentiated. Operation of the differentiators is as follows:

Previous to the initiation of the amplifier's logic sequence gates 1 and 2 are in a locked or conducting state, providing an initial differentiating time constant of $C_1R_1$. Upon initiation of the logic sequence, gates 1 and 2 are switched open by removing base drive from transistors $Q_1$ and $Q_2$. The differentiation time constant during this sampling-time period, which has been instantly altered, is now approximately equal to $C_1R_2$, which in the case shown is made to equal $100C_1R_1$. Pulses now entering from the preamplifier arrive following the rapid change in time constant because of the delay provided, and suffer only minimal loss in amplitude through both differentiators during the differentiation time period because of the long differentiating time constants present. Following the differentiation period gates 1 and 2 are again returned to their normal or conducting state. This restores the initial short time constant, and thereby quickly restores the reference base lines in both circuits. In the output level differentiator the time-constant-modifying transistor serves also as a signal gate. This eliminates base line recovery displacement from the output pulse picture. Both gates are time-synchronized and operated prior to signal arrival so as to minimize offset error. Also
To avoid this difficulty a technique is employed to produce a finite-width zero slope characteristic of all preamplifier pulses, as shown in Fig. 5c. This is accomplished by inhibiting the resistive feedback path of the input charge-sensitive amplifier during the amplifier's time-sampling period, as shown in the preamplifier block diagram of Fig. 6. Thus a charge proportional to that collected in the detector from an event occurrence is placed on the capacitor $C_{fb}$ of the charge-sensitive input stage and held temporarily as a dc voltage. After sampling, this charge must be drained off. The run-down resistor $R_{fb}$ provides this function. It is driven partially from the output of the charge-sensitive amplifier and (to a much lesser degree) from the output of the second inverting operational amplifier, normally having unity gain. This second operational amplifier can be gain-switched through gating of its feedback parameters, resistors $R_4$ and $R_5$ and $Q_3$. Upon gating command from the logic system of the main amplifier, this second amplifier switches abruptly during the time-sampling period to a gain at which equal contributions of opposite polarity run-down voltages are presented through resistors $R_4$ and $R_2$ to their common connection with resistor $R_{fb}$. Under this condition discharge of $C_{fb}$ is inhibited during sampling time, allowing for zero-slope preamplifier output pulse shape for duration of the time-sampling period. Capacitor $C_n$ is employed to neutralize undesirable charge-transfer effects from the gain-switching amplifier, through the feedback resistor $R_{fb}$'s feedthrough capacitance, into the preamplifier's input. This must be done to maintain excellent resolution when preamplifier pulse pile-up conditions exist.
Where detector-to-preamplifier coupling is capacitive, as shown, another effect can arise which can cause spectral shift and some degradation of energy resolution at high counting rates. As rates become high an average dc voltage shift can occur across this input coupling capacitor, which will in turn produce an opposite potential shift at the charge-sensitive preamplifier's output. This dc potential increase then adds to those differential potentials developed across the feedback network of the input amplifier by nuclear events. This increase in total voltage in turn produces more rapid discharge of the time-constant-determining parameters of the feedback network and consequently a slight modification of the preamplifier's output fall time. This slight increase in fall time can produce an average shift of spectral position as well as some spectral smear. This slight spectral shift can be later in the amplifier system by obtaining a correction voltage proportional to total pulse area and adding this correction to all amplifier output pulses. For best correction the time constant associated with this area integration should equal that of the shift-producing parameters $R_{fb}$ and $C_c$ of the charge-sensitive input amplifier.

This correction is quite simply added in the time-sampling amplifier by injecting the dc correcting potential developed into the first time-sampling differentiator, as shown in Fig. 7. Here it is added by applying an offsetting current through resistor $R_c$.

**Spectral Subtraction and Expansion**

Input spectral subtraction, as suggested by Gatti a number of years ago, has never to my knowledge been practically applied. Practical
application of this technique has been limited by the difficulty of
developing randomly occurring exponentially shaped subtracting pulses
on demand, and by the lack of an effective technique for injecting sub-
tracting charge directly into the detector's output circuit.

Since the preamplifier already described can develop zero-slope
characteristics during sampling time, the problem of developing suitably
shaped subtract pulses on demand is greatly simplified. Square-topped
subtracting pulses occurring only during the time-sampling period are
now all that is required. These pulses are easily generated by driving
the normally cut off transistor Q1 of Fig. 6 into conduction during the
time-sampling period. The developed output-pulse amplitude of this
circuit will be determined by the voltage amplitude of the subtracting
potential applied to the transistor's collector. As the transistor is
switched to its conductive state an abrupt potential change will occur.
This potential change will also appear across the subtract-injecting
capacitor $C_s \cdot A - \Delta Q$ associated with this potential change, equal to
$-(\Delta E) \cdot C_s$, will therefore be injected into the preamplifier's input.
This $-\Delta Q$ will subtract from the $\Delta Q_d$ developed by the nuclear event
interaction in the detector, producing this charge difference only during
the time-sampling period. Upon termination of the time-sampling
period the gate transistor's resetting removes this $-\Delta Q$ almost instantly.
The difference action at the preamplifier's input develops a pulse shape
from the preamplifier's output similar to that in Fig. 8b. It is only the
flat portion or pedestal, following initiation of the event and the logic
cycle, that is employed to later amplify and expand.
Obvious advantages of this means of spectral subtraction and expansion are that considerable simplicity ensues and that design of complex high-rate postamplification spectral expansion is now unnecessary.

Reduction of Sensitivity to Variations in Input Capacitance

Where serious alpha spectroscopy studies are being performed, it is highly desirable to reduce preamplifier sensitivity to variations in input capacitance. Alpha detectors used at room temperature exhibit rather poor capacitance-temperature and capacitance-voltage characteristics. Temperature variations of only a degree or so can seriously impair resolution if poor charge sensitivity prevails in the preamplifier. Sensitivity to voltage variations could also become a problem as high-rate pile-up conditions produce shifts in the average of the detector bias voltage. A considerable improvement in detector voltage-temperature stability can be easily attained by applying partial positive feedback to the usual charge-sensitive input amplifier. The large increase in open loop gain thus afforded can yield improvement approaching two orders of magnitude. A technique describing means for easily accomplishing this increase has been described in a previous paper. This technique is applied to the preamplifier used, where positive feedback is applied from the charge-sensitive amplifier's output back to the junction point of the upper and lower sections of the input cascode stage.

CIRCUIT DETAILS

Preamplifier

The preamplifier arrangement of Fig. 6 is used for high-rate
alpha counting. This preamplifier employs two operational amplifiers, the first as a charge-sensitive amplifier and the second as a gain-switching amplifier. Also included are gates for run-down inhibition and for spectral subtraction.

In the charge-sensitive amplifier, feedback occurs from output to input through $C_{fb}$ and $R_{fb}$. Part of the feedback applied to $R_{fb}$ is, however, applied from the output of the second gain-switching amplifier, whose action was described earlier. Also attached to the preamplifier's input circuit are capacitors $C_c$, $C_s$, $C_p$, and $C_n$. With the exception of the coupling capacitor $C_c$, these are all fractional picofarad capacitors so as to produce minimal input capacitance shunting. Capacitor $C_s$ is used to inject subtract charge for spectral subtraction, capacitor $C_p$ is for calibration pulse-generator injection, and capacitor $C_n$ is for neutralization of unwanted capacitance feed-through across $R_{fb}$. The second or gain-switching amplifier is employed solely as a means of preventing discharge of $C_{fb}$ during the time-sampling period, as described earlier. Transistor $Q_4$ serves to generate the spectral subtract voltage, $Q_2$ acts as the gain-switching gate for the second operational amplifier, and $Q_3$ acts as an inverter to produce proper phase signals for neutralization of the feed-through capacitance of $R_{fb}$. Output circuits of both operational amplifiers employ low-impedance complementary emitter-follower drivers for greater output isolation; direct coupling is used throughout to eliminate any multiple differentiation tendency and high open-loop gain is assured by the use of dynamic constant-current loading of voltage gain stages and the inclusion of cascaded current gain stages.
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A time-sampling and logic-program approach has been applied to a nuclear-event-processing amplifier system in order to provide excellent low- and high-counting-rate energy-resolution performance. This approach has also permitted practical inclusion of low-level input spectral subtraction having great simplicity and little loss of high-rate energy resolution.
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With the inclusion of a logic sequence in the amplifier system, these distressed events can be easily rejected on the basis of shape or timing if the decision-making time can be made to follow the complete time-sampling period.

Spectral Shift

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stages also has limitations. It may be quite difficult to accomplish in a practical case in which postdifferentiation gain is moderate to high. Base-line instabilities resulting from dc coupling can produce serious impairment of resolution, and there is no means of supressing low-frequency disturbances present in earlier amplifier stages.

Although the problem of erratic or spurious pulse formation is a detector and not a rate problem, it would still be desirable to not measure these distressed events. This can be accomplished by rejection on the basis of rise time or shape. Logic circuitry associated with an amplifier system can be very easily programmed also to fulfill this function.

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Several advantages ensue by employing the time-sampling technique described. These are improved base line restoration, without resolution sacrifice, reduced rise time jitter modulation, simplification of the high rate spectral expansion problem, insensitivity to low frequency interferences and the ability to reject distressed events on a timing or shape basis.

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Employment of multiple active time-sampling differentiation appears to be an attractive means of reducing the time necessary for after-pulse base-line recovery and of providing improved stability of the reference base line without impairing basic energy resolution. In this differentiating scheme, as shown in Fig. 3, pulses entering the main time-sampling amplifier initiate a logic sequence which provides for active
operation of the two differentiation circuits used. Before arriving at
the first differentiator all pulses are delayed, to permit initiation of the
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this sampling-time period, which has been instantly altered, is now
approximately equal to $C_1 R_2$, which in the case shown is made to
equal 100 $C_1 R_1$. Pulses now entering from the preamplifier arrive
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quickly restores the reference base lines in both circuits. In the out-
put level differentiator the time-constant-modifying transistor serves
also as a signal gate. This eliminates base line recovery displacement
from the output pulse picture. Both gates are time-synchronized and
operated prior to signal arrival so as to minimize offset error. Also
associated with the second differentiator are provisions for separate time period output sampling and for signal integration.

**Improved Information Density and Rise Time Jitter Modulation Suppression**

Active gate differentiation provides almost constant pulse output amplitude for the duration of the sampling period. This gives an almost ideal information-density characteristic, except for the rise-time portions of pulses being processed.

Since information averaging over this time period could be affected by variations in pulse rise time, actual signal sampling occurs only during the latter part of the differentiation time period as shown in Fig. 1. This signal-sampling interval is developed by the logic generator, which drives a direct coupled output gate that is initiated at a predetermined time after the onset of an entering pulse and that terminates prior to termination of the differentiation period.

**Logical Rejection of Distressed Events**

Development of a logic sequence in a nuclear pulse amplifier greatly eases the problems of rejecting distressed events. Additional logic circuitry can be very easily provided to inhibit readout on the basis of pulse interaction or shape.

**Integration Function**

Although an output level integrator of the conventional finite time-constant variety is used, it is obviously desirable to use an integrator that can totally sum pulse information occurring during the time sampling period and hold this information until logical decisions can be made with respect to accepting for read out.
Present plans call for the use of an infinite time-constant programmed integrator similar to the one described in an earlier paper. An integrator of this type can be activated and reset rapidly. It can be logic-programmed to produce an output dc voltage proportional to the area of the input pulse. This temporarily stored output voltage can then be sampled and read later by means of a read gate actuated by the amplifier's logic cycle. Since readout would follow the complete prior time-sampling period in the amplifier, any "inhibit" decision based upon pulse interaction or shape could already have been made. Also, amplifier output would be based on total information accumulated during the time-sampling period and not upon a portion, as would be the case if the information-averaging time constant were finite. A more accurate measuring decision should be possible from this total available information. A function diagram of the programmed integrator is shown in Fig. 4 where gate transistor Q4 provides switching of the integrator's time constant by lockout of the operational amplifier's resistive feedback path.

Special Preamplifier Requirements

Aside from the usually desired preamplifier characteristics of excellent differential linearity, freedom from multiple differentiation, and excellent stability, there are those of providing pulse shape compatibility with the time-sampling technique.

Normal exponentially decaying pulses are not compatible with the time-sampling technique, as postdifferentiation skewing of pulse tops results from preamplifier pile-up condition, as shown in Fig. 5a and b. Pulse-top skewing arises from pile-up produced compounding of the preamplifier's output fall time.
To avoid this difficulty a technique is employed to produce a finite-width zero slope characteristic of all preamplifier pulses, as shown in Fig. 5c. This is accomplished by inhibiting the resistive feedback path of the input charge-sensitive amplifier during the amplifier's time-sampling period, as shown in the preamplifier block diagram of Fig. 6. Thus a charge proportional to that collected in the detector from an event occurrence is placed on the capacitor \( C_{fb} \) of the charge-sensitive input stage and held temporarily as a dc voltage. After sampling, this charge must be drained off. The run-down resistor \( R_{fb} \) provides this function. It is driven partially from the output of the charge-sensitive amplifier and (to a much lesser degree) from the output of the second inverting operational amplifier, normally having unity gain. This second operational amplifier can be gain-switched through gating of its feedback parameters, resistors \( R_4 \) and \( R_5 \) and \( Q_3 \). Upon gating command from the logic system of the main amplifier, this second amplifier switches abruptly during the time-sampling period to a gain at which equal contributions of opposite polarity run-down voltages are presented through resistors \( R_4 \) and \( R_2 \) to their common connection with resistor \( R_{fb} \). Under this condition discharge of \( C_{fb} \) is inhibited during sampling time, allowing for zero-slope preamplifier output pulse shape for duration of the time-sampling period. Capacitor \( C_n \) is employed to neutralize undesirable charge-transfer effects from the gain-switching amplifier, through the feedback resistor \( R_{fb} \)'s feedthrough capacitance, into the preamplifier's input. This must be done to maintain excellent resolution when preamplifier pulse pile-up conditions exist.
Where detector-to-preamplifier coupling is capacitive, as shown, another effect can arise which can cause spectral shift and some degradation of energy resolution at high counting rates. As rates become high an average dc voltage shift can occur across this input coupling capacitor, which will in turn produce an opposite potential shift at the charge-sensitive preamplifier's output. This dc potential increase then adds to those differential potentials developed across the feedback network of the input amplifier by nuclear events. This increase in total voltage in turn produces more rapid discharge of the time-constant-determining parameters of the feedback network and consequently a slight modification of the preamplifier's output fall time. This slight increase in fall time can produce an average shift of spectral position as well as some spectral smear. This slight spectral shift can be later in the amplifier system by obtaining a correction voltage proportional to total pulse area and adding this correction to all amplifier output pulses. For best correction the time constant associated with this area integration should equal that of the shift-producing parameters $R_{fb}$ and $C_c$ of the charge-sensitive input amplifier.

This correction is quite simply added in the time-sampling amplifier by injecting the dc correcting potential developed into the first time-sampling differentiator, as shown in Fig. 7. Here it is added by applying an offsetting current through resistor $R_c$.

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Input spectral subtraction, as suggested by Gatti, has never to my knowledge been practically applied. Practical
application of this technique has been limited by the difficulty of developing randomly occurring exponentially shaped subtracting pulses on demand, and by the lack of an effective technique for injecting subtracting charge directly into the detector's output circuit.

Since the preamplifier already described can develop zero-slope characteristics during sampling time, the problem of developing suitably shaped subtract pulses on demand is greatly simplified. Square-topped subtracting pulses occurring only during the time-sampling period are now all that is required. These pulses are easily generated by driving the normally cut off transistor Q1 of Fig. 6 into conduction during the time-sampling period. The developed output-pulse amplitude of this circuit will be determined by the voltage amplitude of the subtracting potential applied to the transistor's collector. As the transistor is switched to its conductive state an abrupt potential change will occur. This potential change will also appear across the subtract-injecting capacitor $C_s \cdot A - \Delta Q$ associated with this potential change, equal to $-(\Delta E) \cdot C_s$, will therefore be injected into the preamplifier's input. This $-\Delta Q$ will subtract from the $\Delta Q_d$ developed by the nuclear event interaction in the detector, producing this charge difference only during the time-sampling period. Upon termination of the time-sampling period the gate transistor's resetting removes this $-\Delta Q$ almost instantly. The difference action at the preamplifier's input develops a pulse shape from the preamplifier's output similar to that in Fig. 8b. It is only the flat portion or pedestal, following initiation of the event and the logic cycle, that is employed to later amplify and expand.
Obvious advantages of this means of spectral subtraction and expansion are that considerable simplicity ensues and that design of complex high-rate postamplification spectral expansion is now unnecessary.

Reduction of Sensitivity to Variations in Input Capacitance

Where serious alpha spectroscopy studies are being performed, it is highly desirable to reduce preamplifier sensitivity to variations in input capacitance. Alpha detectors used at room temperature exhibit rather poor capacitance-temperature and capacitance-voltage characteristics. Temperature variations of only a degree or so can seriously impair resolution if poor charge sensitivity prevails in the preamplifier. Sensitivity to voltage variations could also become a problem as high-rate pile-up conditions produce shifts in the average of the detector bias voltage. A considerable improvement in detector voltage-temperature stability can be easily attained by applying partial positive feedback to the usual charge-sensitive input amplifier. The large increase in open loop gain thus afforded can yield improvement approaching two orders of magnitude. A technique describing means for easily accomplishing this increase has been described in a previous paper. This technique is applied to the preamplifier used, where positive feedback is applied from the charge-sensitive amplifier's output back to the junction point of the upper and lower sections of the input cascode stage.

CIRCUIT DETAILS

Preamplifier

The preamplifier arrangement of Fig. 6 is used for high-rate...
alpha counting. This preamplifier employs two operational amplifiers, the first as a charge-sensitive amplifier and the second as a gain-switching amplifier. Also included are gates for run-down inhibition and for spectral subtraction.

In the charge-sensitive amplifier, feedback occurs from output to input through $C_{fb}$ and $R_{fb}$. Part of the feedback applied to $R_{fb}$ is, however, applied from the output of the second gain-switching amplifier, whose action was described earlier. Also attached to the preamplifier's input circuit are capacitors $C_c$, $C_s$, $C_p$, and $C_n$. With the exception of the coupling capacitor $C_c$, these are all fractional picofarad capacitors so as to produce minimal input capacitance shunting. Capacitor $C_s$ is used to inject subtract charge for spectral subtraction, capacitor $C_p$ is for calibration pulse-generator injection, and capacitor $C_n$ is for neutralization of unwanted capacitance feed-through across $R_{fb}$. The second or gain-switching amplifier is employed solely as a means of preventing discharge of $C_{fb}$ during the time-sampling period, as described earlier. Transistor $Q_4$ serves to generate the spectral subtract voltage, $Q_2$ acts as the gain-switching gate for the second operational amplifier, and $Q_3$ acts as an inverter to produce proper phase signals for neutralization of the feed-through capacitance of $R_{fb}$. Output circuits of both operational amplifiers employ low-impedance complementary emitter-follower drivers for greater output isolation; direct coupling is used throughout to eliminate any multiple differentiation tendency and high open-loop gain is assured by the use of dynamic constant-current loading of voltage gain stages and the inclusion of cascaded current gain stages.
Output signals from this preamplifier are taken directly from the output of the charge-sensitive operational amplifier and fed to the input of the main time-sampling amplifier.

Main Time-Sampling Amplifier

This unit whose block diagram is shown in Fig. 3 comprises the main amplifier and the logic sequence generator. The main amplifier contains gain stages, two time-sampling differentiators, a time-sampling gate, an integrator, and necessary gain, integrate, and subtract controls.

The logic sequence generator provides all the logic commands and decisions necessary for proper operation of differentiation integration, signal sampling, and distressed-event inhibiting.

Amplifier Circuits

Pulses entering the main amplifier of Fig. 3 are applied simultaneously without delay to the input of the logic sequence generator and with a 100-nanosecond delay to the input of the amplifier section.

The input amplifier is quite similar in design to those employed in the preamplifier, with particular emphasis upon low noise generation. From the extremely low-impedance output of this input amplifier, the first time-sampling differentiation circuit is driven. The capacitor used in this differentiator and that used in the output differentiator form the only ac coupling circuits used in the entire amplifier. Outputs from the first time-sampling differentiator pass into the input of a differential input low-noise operational amplifier. This differential input arrangement is employed here to reduce dc drift and to provide proper
direct-coupled voltage matching to the time-differentiating gates. Following the differential input amplifier are two other cascaded operational amplifiers, which provide the required gain, and the control of the gain.

The low-impedance output of the last of these operational amplifiers in turn drives the second or output level time-sampling differentiator. Output from this second differentiator passes through the direct-coupled signal-sampling gate, through an integrating circuit, and finally to the amplifier's output through a cascaded complementary output follower.

Logic Sequence Generator

Because of the amplifier's input delay line, by the time input pulses enter the main time-sampling amplifier, they have already initiated the logic sequence cycle. Pulses entering the logic sequence generator of Fig. 9 are amplified by an operational amplifier employing a Fairchild 710 micrologic. Output from this wide-band amplifier drives into another 710 micrologic employed as a Schmitt discriminator. Output from this Schmitt discriminator passes through a transistor emitter-follower employed in a peak charging circuit, whose output is connected to a second Schmitt discriminator. The leading edge of the output pulse of this second Schmitt discriminator is used to trigger both timing circuits used for differentiation, spectral subtraction, and signal-sampling control. Employment of the peak charging circuit and the second Schmitt trigger circuit permits inhibition of events occurring too close to the termination of previous events. If distressed events occur during this timing interval, determined by the time constant
of the peak charging circuit no output can be produced from the second Schmitt circuit to initiate the amplifier's differentiation and read-out functions. Thus no amplifier output is produced when an event falls within this predetermined recovery period of the multiple differentiators. Under undistressed-event conditions (i.e., adequate interval between events), the differentiation one-shot and the read one-shot univibrators trigger normally, providing a normal amplifier output.

Timing accuracy in the differentiating and particularly in the read one-shot circuits must be exceptional so as not to introduce reading errors. Small variations in timing can produce output pulse-area variations and consequently reading errors. The Fairchild 710 micrologic circuits employed for the univibrator functions have proved exceptionally stable.

Suitable gate drivers are provided to couple the univibrator outputs to their respective driving functions in the main amplifier and preamplifier.

RESOLUTION AND RATE PERFORMANCE

So far the time-sampling amplifier has been employed only for alpha spectroscopy. Low-rate alpha resolution approaching 14 keV with $^{244}$Cm has been possible and approximately 18 keV can be obtained at rates approaching 15 000 c/sec. Spectral shifts of $<0.05\%$ were maintained under the same conditions. In comparison with gamma-counting systems, the approximately 15 000 counts/sec alpha rate produces several times as much interpulse interference in an amplifier as does a like gamma-counting rate. This difference in the interpulse interference ratio between gammas and alphas is due to their
differences in spectral distribution. Practically all of a given alpha spectrum appears above 90% of maximum, and the greater share of a given gamma spectrum falls much below this figure.

Possible Improvements

As with any new approach, numerous improvements could be made, some of which have already been anticipated and many more which have not. Further improvement in performance should be attained by improving integrator operation, and by providing inhibition for pulses distressed by stack-up. Although like differentiation is applied before and after each event, it would probably be beneficial to provide a shorter time-constant differentiation following each event. Also the possibility of directly resetting the charge-sensitive input amplifier following each event should likewise be examined in detail. Single time-sampling differentiation may prove to be beneficial to low-counting-rate energy resolution when the second differentiator is not needed. Many such possibilities exist and can be easily realized by modification of the logic cycle sequence functions.
FOOTNOTE AND REFERENCES

*This work was performed under the auspices of the U. S. Atomic
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FIGURE LEGENDS

Fig. 1. Comparison between conventional and time-sampling techniques.

Fig. 2. Distressed-event occurrence.

Fig. 3. Active double time-sampling differentiator and signal sampler.

Fig. 4. Programmed integrator.

Fig. 5. Pile-up shape compatibility for time-sampling technique.

Fig. 6. Spectral subtracting time-sampling preamplifier.

Fig. 7. Spectral shift correction.

Fig. 8. Preamplifier output without and with spectral subtraction.

Fig. 9. Block diagram of logic sequence generator.
Conventional with differentiation and integration

Preamplify

Differentiation

Sample

Integration

Output

Time-sampling technique

Differentiate
Sample
Read
Reset

Fig. 1
Prompt base-line disturbance

Addition or stack-up

Fig. 2
Fig. 3
Sampled signal

Linear gate

Input

From amp

Logic sequence generator

Output

$C_{fb}$

$R_{fb1}$

$R_{fb2}$

$Q_1$ Gate

Output

Sampling time

XBL67II-5580-A

Fig. 4
<table>
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<th>Differentiating output</th>
<th>Preamplifier output</th>
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Fig. 5
Fig. 6
Fig. 7

-27- UCRL-17904

Input

Delay

\[ R_{\text{int}}C_{\text{int}} = R_{\text{fb}}C_o \text{ of Preamp} \]

Differentiating gate

DC Amplifier

Integrating

From amplifier output

XBL67II-5583-A
Without

a

With

b

Time sampling
Differentiation time

Fig. 8
Fig. 9
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