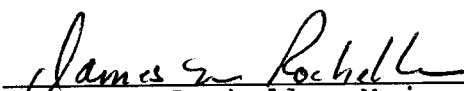
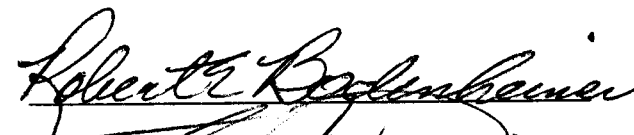
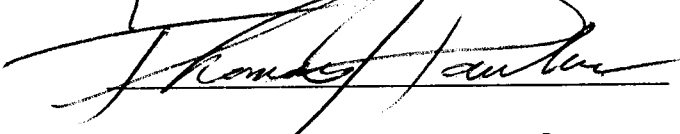


To the Graduate Council:

I am submitting herewith a dissertation written by Michael Leon Simpson entitled "Resolution Enhancement of Coaxial, High-Purity, Ge Gamma-Ray Detectors." I have examined the final copy of this dissertation for form and content and recommend that it be accepted in partial fulfillment of the requirements for the degree of Doctor of Philosophy, with a major in Electrical Engineering.


James M. Rochelle, Major Professor

We have read this dissertation
and recommend its acceptance:



Carroll R. Bingham


Accepted for the Council:


Associate Vice Chancellor
and Dean of The Graduate School

RESOLUTION ENHANCEMENT OF
COAXIAL, HIGH-PURITY, Ge
GAMMA-RAY DETECTORS

A Dissertation
Presented for the
Doctor of Philosophy
Degree
The University of Tennessee, Knoxville

Michael Leon Simpson

May 1991

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ABSTRACT

The enhancement of the energy resolution of high-purity, Ge (HPGe), gamma-ray detectors was studied. In particular, circuits which correct for ballistic deficit and charge trapping were analyzed. Also, an equation from which energy line shapes can be calculated for arbitrary charge trap distributions and spectroscopy amplifier peaking times was developed. The resolution enhancement circuits were studied over the energy range from 100keV to 10MeV and spectroscopy amplifier peaking times from 1us to 12us. The charge trapping correction circuit improved the full-width-at-half-maximum (FWHM) resolution of a radiation damaged, HPGe detector from 4.3keV to 2.0keV at the 1.33MeV energy line. And, a ballistic deficit correction circuit improved the resolution of an HPGe detector from 12.1keV to 4.95keV at the 2.61MeV energy line and spectroscopy amplifier peaking time of 2us. The data also indicated the added importance of charge trapping correction at higher energies. For example, the resolution of a detector at 10MeV, which exhibited negligible ballistic deficit and charge trapping at 1.33MeV (spectroscopy amplifier peaking time = 12us), was improved from 7.3keV (calculated) to 5.6keV (extrapolated) when charge trapping and ballistic deficit corrections were used. For energies above approximately 800keV, an instrument which combined a gated integrator with

a charge trapping correction circuit offered the best resolution.

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CHAPTER 1

INTRODUCTION

1.1 Background

The study of nuclear physics often involves the precise measurement of gamma-ray energy spectra. After many years of development, the instrumentation used to measure these spectra has matured to the point that fundamental physical processes are usually the limiting factors in the energy measurement. To illustrate this point, consider the conventional gamma-ray spectroscopy system shown in figure (1-1). This system consists of a coaxial, high-purity, Ge (HPGe) detector that is fully depleted of free carriers by a bias of 3 - 5 kV; a charge-sensitive preamplifier; a spectroscopy amplifier; an analog-to-digital converter (ADC); and a multichannel analyzer (MCA). In the ideal case, this system operates as follows. A gamma-ray interacting in the bulk material of the detector creates a number of electron-hole pairs that is directly proportional to the energy of the incident gamma-ray. The electric field generated by the bias supply drives the charge across the detector to the positive (electrons) and negative (holes) electrodes where it is collected and presented to the charge-sensitive preamplifier as an impulse of current. Since an ideal charge-sensitive preamplifier is an

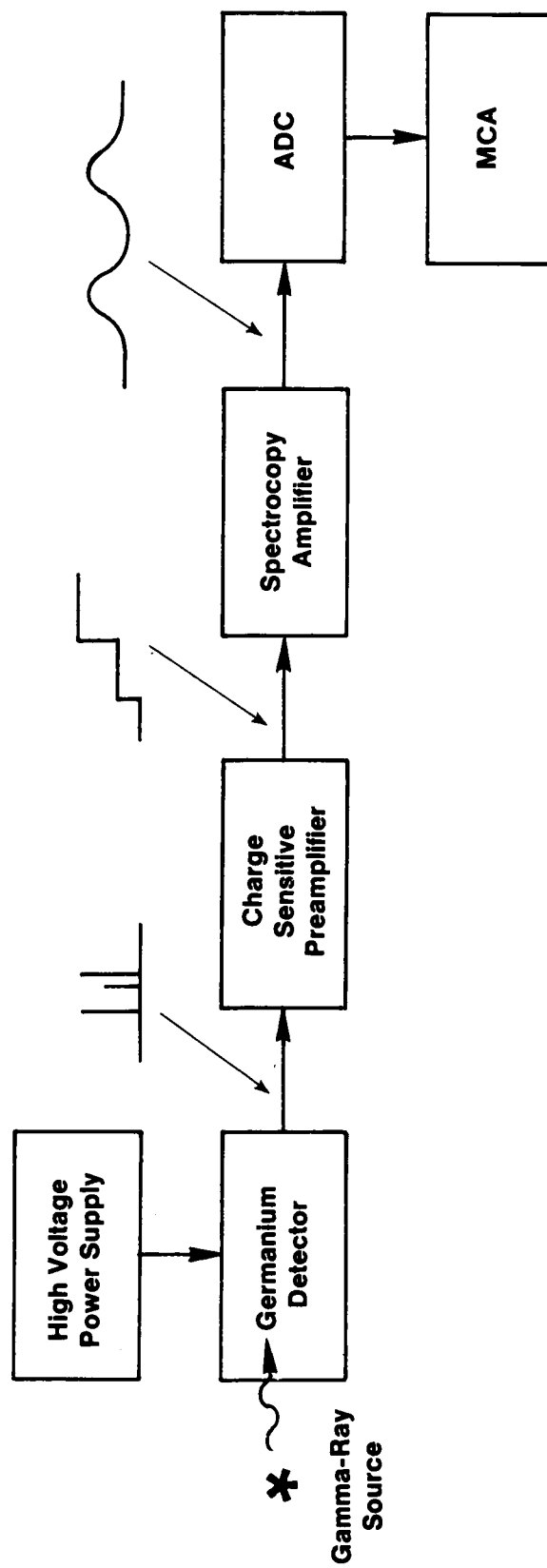


Figure (1-1). Ideal gamma-ray spectroscopy system.

integrator, its response to this impulse of current would be a step function with zero rise time. The spectroscopy amplifier shapes this step function into a pulse whose peak amplitude is detected and digitized by the ADC. The ADC transmits these digital signals to the MCA, where they are sorted according to magnitude into a histogram. This histogram is indicative of the number of gamma-rays that interacted in the detector at every energy in the range of the spectrometer. The energy spectrum that would result from such a system would be composed entirely of impulse functions (lines) with the only broadening due to the Heisenberg uncertainty principle as shown in figure (1-2).

In the spectrum of figure (1-2), the energy of each line can be accurately assigned, a small peak arbitrarily close to a large peak can be easily resolved, and the ratio of the peak height to the Compton scattered background [12] is maximized. Compton scattering is the process whereby a portion of a gamma-ray's energy is transferred to an electron in the detector [12]. This energetic electron will produce a number of electron-hole pairs that is proportional to the energy transferred to the electron. However, the scattered gamma-ray may travel out of the detector without depositing the remainder of its energy, placing a count in the spectrum at an energy less than the energy of the gamma-ray. The amount of energy transferred to the electron is randomly distributed between zero and a fraction less than

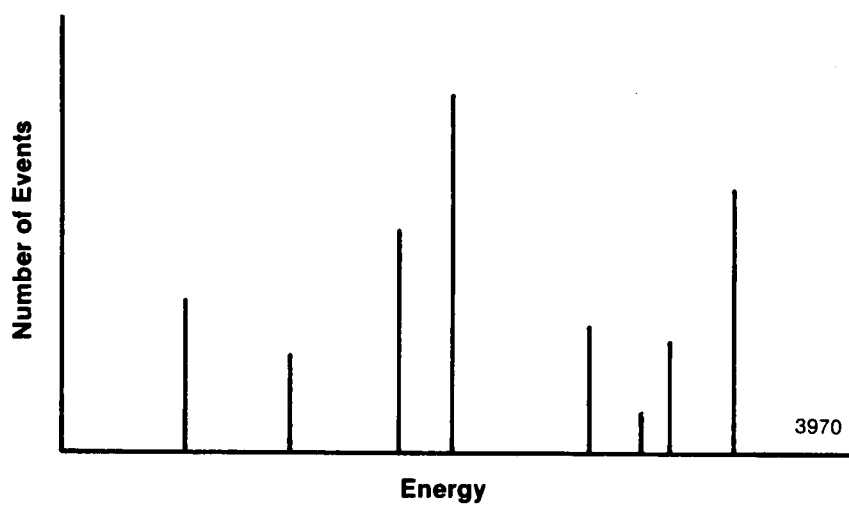


Figure (1-2). Energy spectrum from system in figure (1-1).

one, and gives rise to a background spectrum even in the presence of a mono-energetic gamma-ray source. The ratio of a peak height to the Compton scattered background is a measure of the spectroscopy system's ability to detect very weak gamma-ray sources.

Unfortunately, the ideal system described above is not realizable, and the spectrum of figure (1-2) cannot be obtained. In a spectrum measured by any actual system, the energy lines will have a non-zero width. Thus, the peak-to-Compton ratio will not be optimum, closely spaced peaks may not be resolved, and the accuracy of the energy assignments may suffer. The resolution of the energy lines is determined by several fundamental physical processes, each of which will be described below.

1.2 Resolution Limitations in Ge Spectrometers

One important resolution limiting mechanism is governed by the random nature of charge production in the detector. In the ideal case considered above, it was assumed that all the energy of incident gamma-rays was used to form electron-hole pairs. However, in actual detectors, a portion of each gamma-ray's energy is transferred to the crystal lattice of the detector in the form of phonons instead of electron-hole pair generation [2,6,12]. If the production of electron-

hole pairs were a Poisson process, the variance in the number of pairs produced, v , would be

$$v = N \quad (1-1)$$

where N is the mean number of electron-hole pairs generated [2,6,12]. In this case, the full-width-at-half-maximum (FWHM) for each energy line would be given by (all other line broadening mechanisms excluded)

$$\text{FWHM} = 2.35(E/e)^{0.5}, \quad (1-2)$$

where E is the energy of the incident gamma-ray, e is the average energy required to produce an electron-hole pair, and the units of equation (1-2) are electron-hole pairs. Since the energy deposited by the gamma-ray in the creation of N electron-hole pairs is eN , equation (1-2) can be written as

$$\text{FWHM} = 2.35(eE)^{0.5}, \quad (1-3)$$

where the units are electron volts (eV). Resolution measurements with many types of detectors have shown that the achievable FWHM are much lower than those predicted by equation (1-3) [12]. To account for this discrepancy, a factor F , which is less than unity, is included in equation

(1-3) such that

$$\text{FWHM} = 2.35(eFE)^{0.5}, \quad (1-4)$$

and F is defined as

$$F = v_0/v, \quad (1-5)$$

where v_0 is the actual variance in the number of electron-hole pairs that are generated. F is known as the Fano factor after U. Fano who first treated charge generation fluctuations in gas radiation detectors [5].

It is difficult to measure the intrinsic Fano factor for a detector material, since other effects, most notably charge trapping phenomenon, hide the charge production fluctuations in any practical measurement [6]. However, the Fano factor for Ge at 77 K is known to have an upper bound of 0.08 [18].

Before discussing other resolution limiting phenomenon, it is necessary to digress briefly and examine the signal processing characteristics of the spectroscopy amplifier. The function of the spectroscopy amplifier is to shape the signal to optimize the performance of the spectrometer [6]. This usually involves a compromise between achieving the best possible resolution while obtaining the necessary count rate performance [6]. The spectroscopy amplifier accepts

the step function output of the preamplifier as an input, and responds with a pulse that may appear like that shown in figure (1-3). The time between the occurrence of a gamma-ray interaction in the detector and the spectroscopy amplifier reaching its peak amplitude is known as the peaking time, t_p , of the amplifier. As will be described later in this section, the shape and peaking time of the spectroscopy amplifier output signal are the parameters which control the compromise between obtaining the best resolution and the necessary count rate performance.

The shaping characteristics of the spectroscopy amplifier also influence the noise performance of the spectrometer. In the ideal spectroscopy system described earlier, it was assumed that the signal was unperturbed by electronic noise. However, the detector will have a small but finite leakage current that will give rise to a shot noise component in parallel to the signal, and the preamplifier will have an equivalent input noise generator in series with the signal as shown in figure (1-4). Several authors have discussed the spectroscopy amplifier pulse shape for the optimum signal-to-noise ratio (SNR) [3,4,30]. The infinite cusp, shown in figure (1-5), is the pulse shape which allows for the highest possible SNR [4]. However, the cusp is non-realizable since it requires an output before an input has occurred. Wilson [30] described a circuit which approximated the cusp response and was causal. However, the

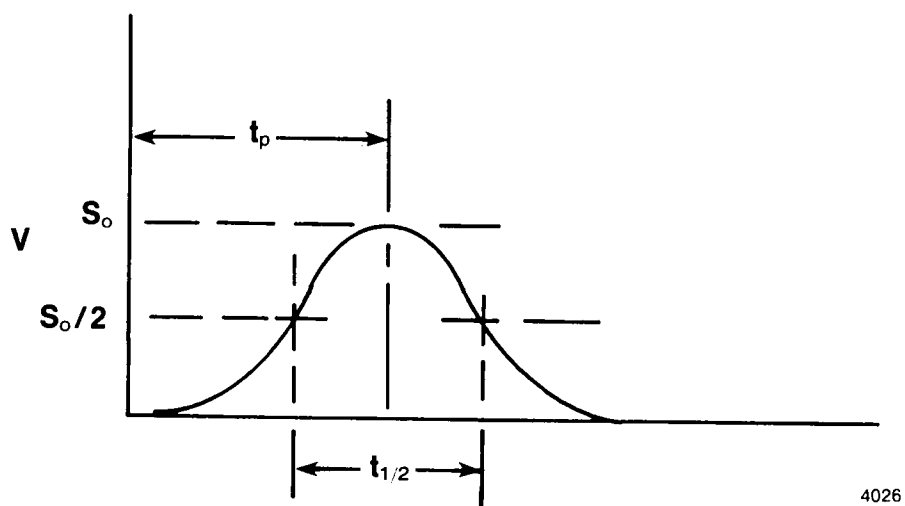
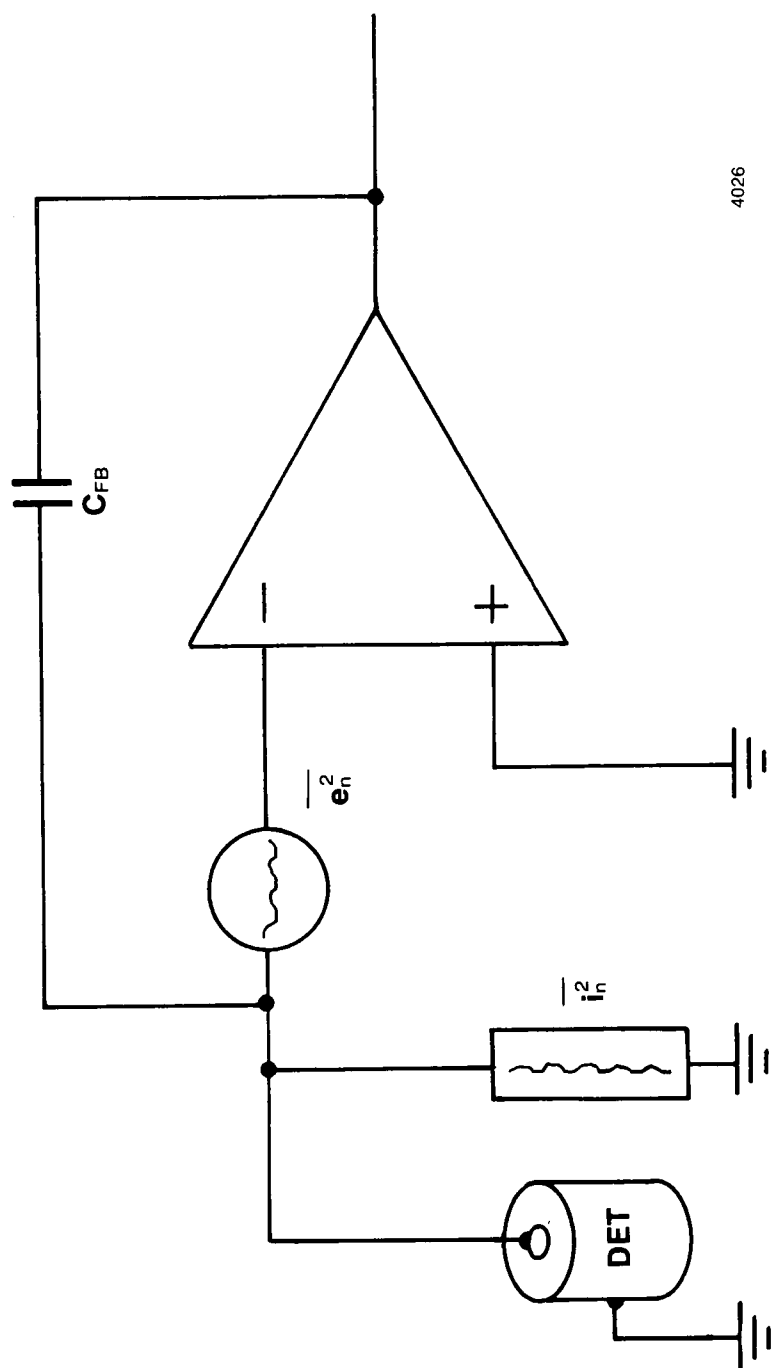


Figure (1-3). Impulse response of semi-Gaussian shaping amplifier.



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Figure (1-4). Preamplifier noise sources.

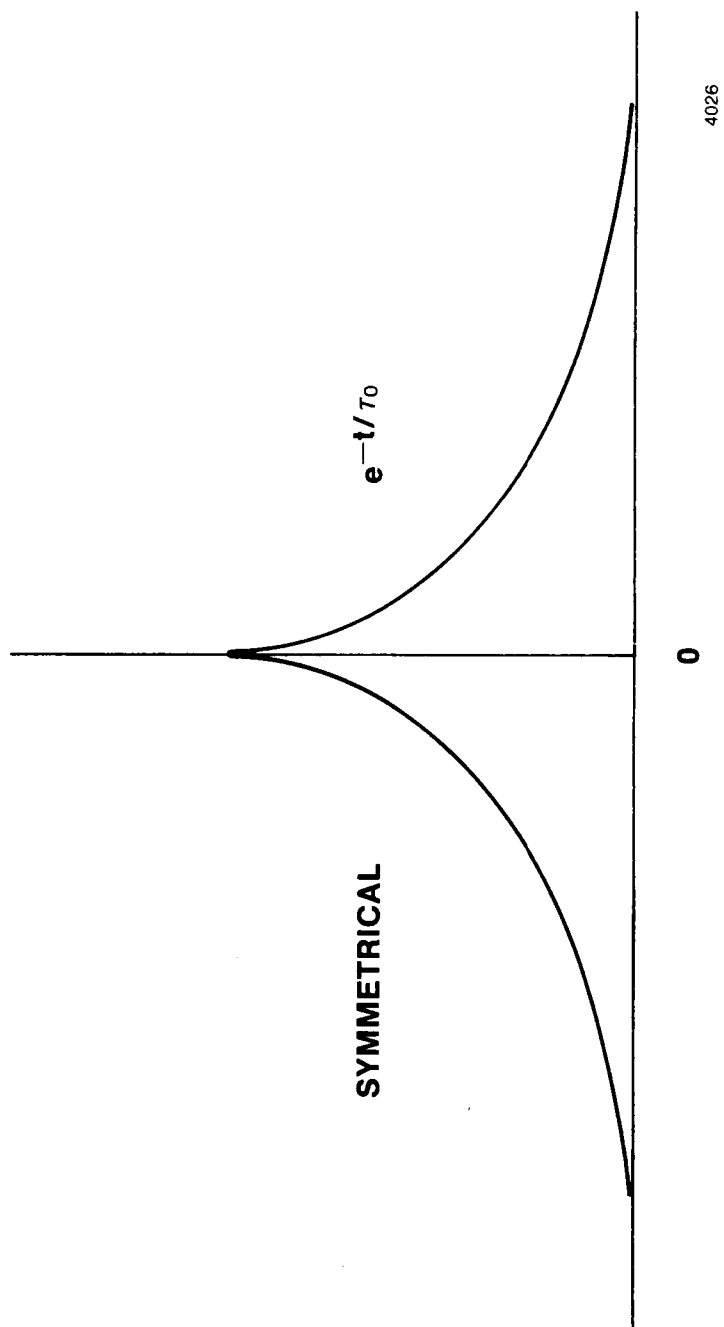


Figure (1-5). Infinite Cusp.

narrow and pointed top of the cusp is not conducive to the operation of the ADC [3,4], and the Wilson circuit has not become popular [4].

The pulse shape resulting from a single RC differentiation and x RC integrations of the preamplifier signal approaches Gaussian as x approaches infinity [4]. The noise performance of such a pulse shape is only 12% worse than that of the cusp [4], and the smoothly varying top of the Gaussian pulse is compatible with ADCs. It is not possible, of course, to create a circuit with an infinite number of RC integrators. However, a circuit with a single RC differentiator and four RC integrators has a noise performance only 16% worse than the cusp [4]. Spectroscopy amplifiers with pulse shapes very similar to this semi-Gaussian shape are now widely used for nuclear spectroscopy. Such a pulse shape is shown in figure (1-3) for the case of $x=7$. The width of a spectroscopy amplifier pulse at half of its maximum value (refer to figure (1-3)) is known as $t_{1/2}$.

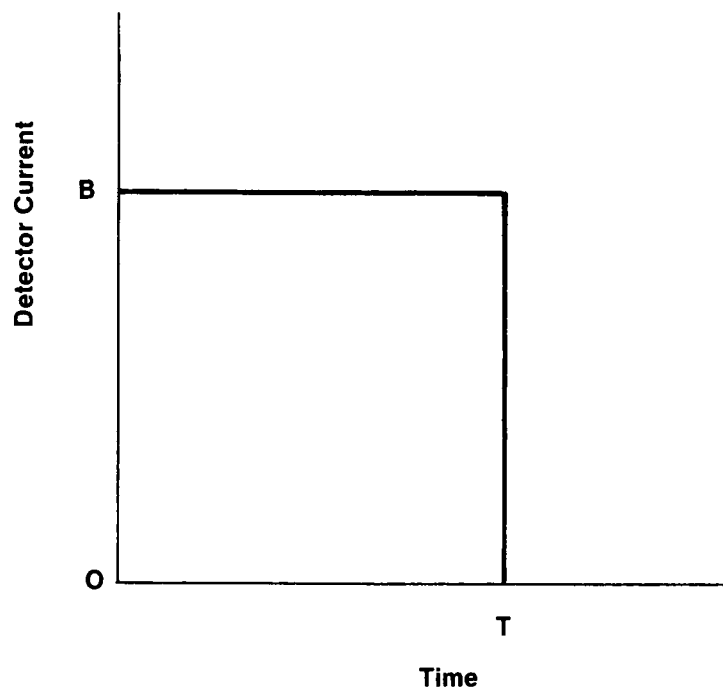
The final two energy resolution limiting phenomena to be considered in this work arise from charge transport mechanisms. In the ideal case considered earlier, it was assumed that all the charge created by the gamma-ray was collected instantaneously, producing an impulse of current. However, in actual detectors a finite amount of time will be required for the charge carriers to reach their respective

electrodes. This charge collection time will result in a current waveform which may appear like that shown in figure (1-6). In figure (1-6) it is assumed that the current very quickly reaches the value B and remains at this level until all the charge is collected. In this case, the amount of charge collected, Q_0 , would be

$$Q_0 = BT, \quad (1-6)$$

where T is the charge collection time. The response of the charge sensitive preamplifier and a semi-Gaussian spectroscopy amplifier to this current waveform can be found by the convolution of the waveform of figure (1-6) with the impulse response of figure (1-3). This procedure is graphically represented in figure (1-7).

As a careful study of figure (1-7) indicates, the duration of the current signal from the detector modulates both the peak amplitude and the peaking time of the spectroscopy amplifier output signal. For example, an impulse input containing charge Q would produce the impulse response of figure (1-3) as an output with the peaking time t_p and amplitude S_0 . However, an input current waveform like that shown in figure (1-6) also containing charge Q would produce an output signal of S ($< S_0$) with a peaking time of [13]



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Figure (1-6). Possible detector current waveform.

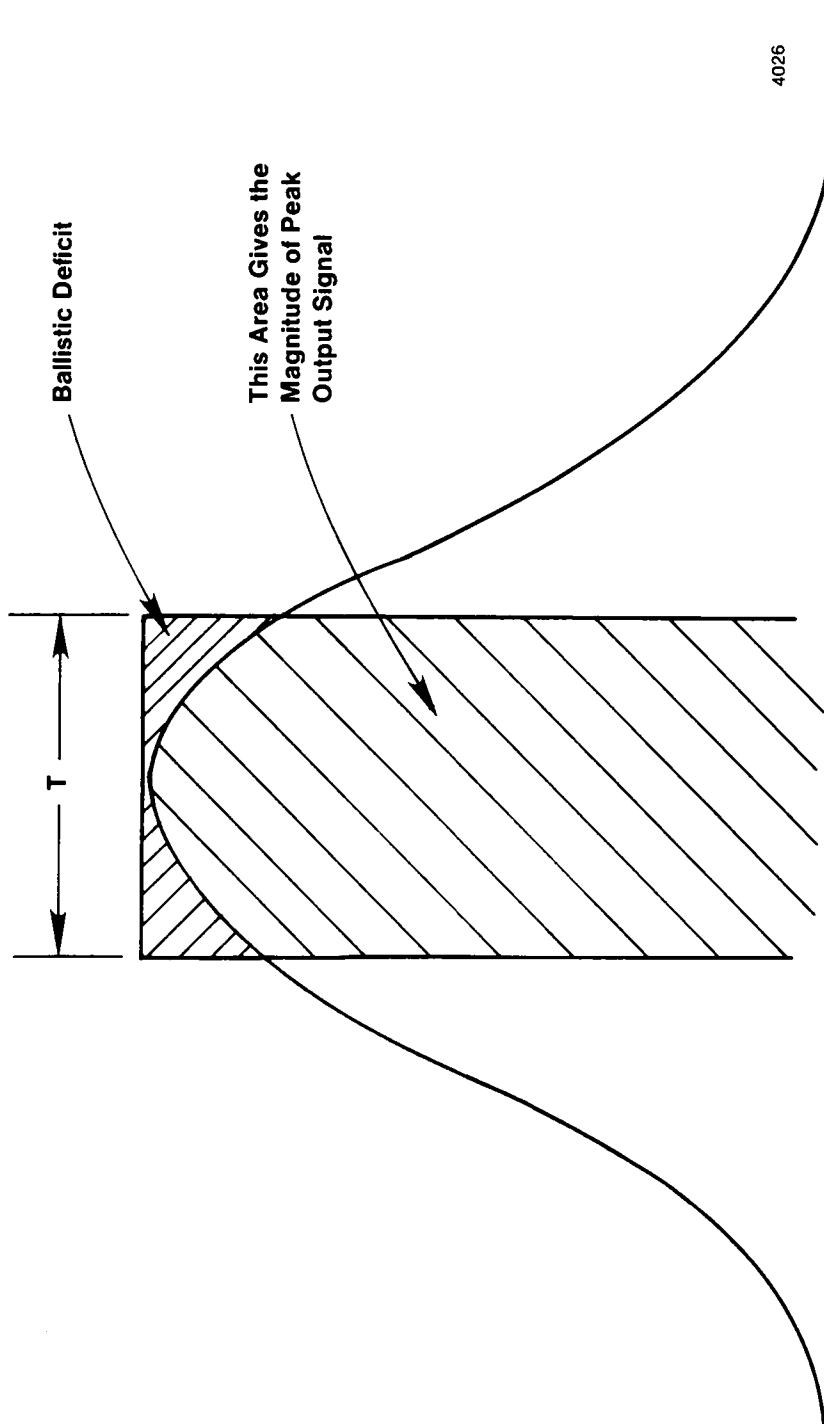


Figure (1-7). Graphical convolution of detector output current with impulse response of spectroscopy amplifier.

$$t_{\text{peak}} \sim t_p + T/2. \quad (1-7)$$

Equation (1-7) assumes the top of the impulse response is approximately symmetric about the maximum value and $T < t_p$. The delay in the peaking time implied by equation (1-7) is known as the unipolar peaking time delay, t_d , and is given by [13]

$$t_d \sim T/2. \quad (1-8)$$

Ballistic deficit is a resolution degrading phenomenon that is defined as

$$BD = S_0 - S, \quad (1-9)$$

and the relative ballistic deficit is defined as

$$RBD = (S_0 - S)/S_0. \quad (1-10)$$

An intuitive understanding of ballistic deficit can be obtained from careful consideration of figure (1-7). The magnitude of the ballistic deficit is directly proportional to the energy of the gamma-ray and to some function of the duration, T , of the current signal from the detector. Since equation (1-7) shows t_d to be linearly related to T , the

ballistic deficit is also proportional to a function, G , of t_d . Thus,

$$BD = S_0 G(t_d). \quad (1-11)$$

Ballistic deficit broadens the energy line in the following manner. Since interactions may occur at any point in the detector, there will be a range of collection times such that

$$T_{min} < T < T_{max}, \quad (1-12)$$

where T_{min} and T_{max} are the minimum and maximum charge collection times respectively. Thus, for a single gamma-ray energy, E_0 , there will be a corresponding spread in the measured energy, E , given by

$$(E_0 - E_0 G(T_{max}/2)) < E < (E_0 - E_0 G(T_{min}/2)), \quad (1-13)$$

producing a broadened energy line as shown in figure (1-8).

One obvious method to reduce the ballistic deficit is to increase the $t_{1/2}$ of the spectroscopy amplifier. As shown in figure (1-9), as the $t_{1/2}$ of the semi-Gaussian spectroscopy amplifier is increased, the portion of the impulse response near the peak value becomes flatter, thus

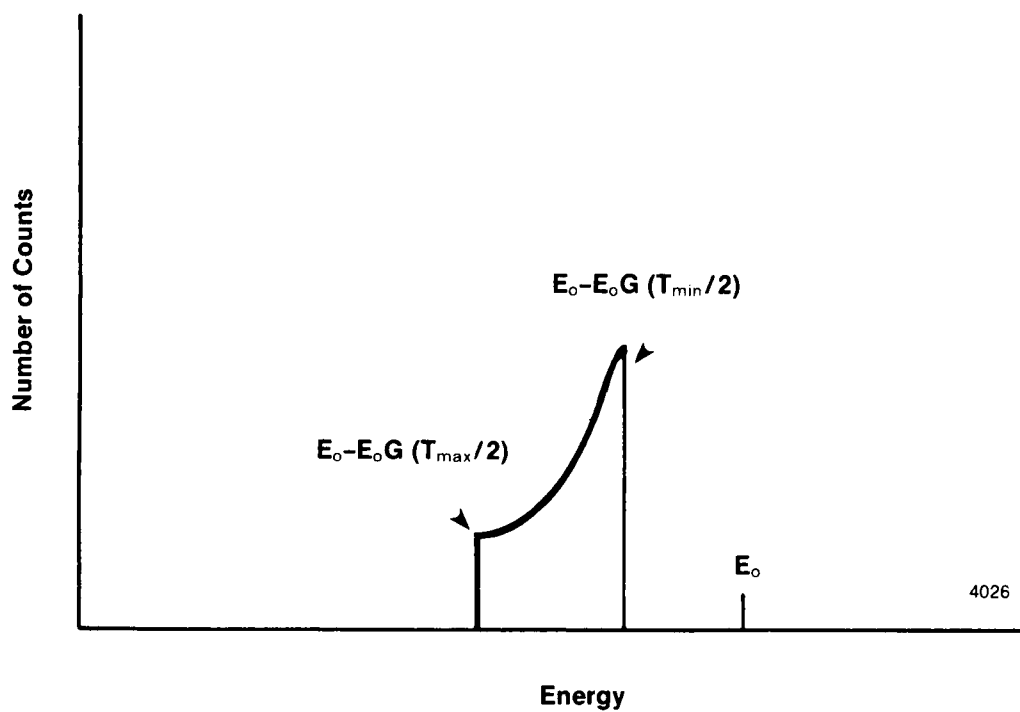
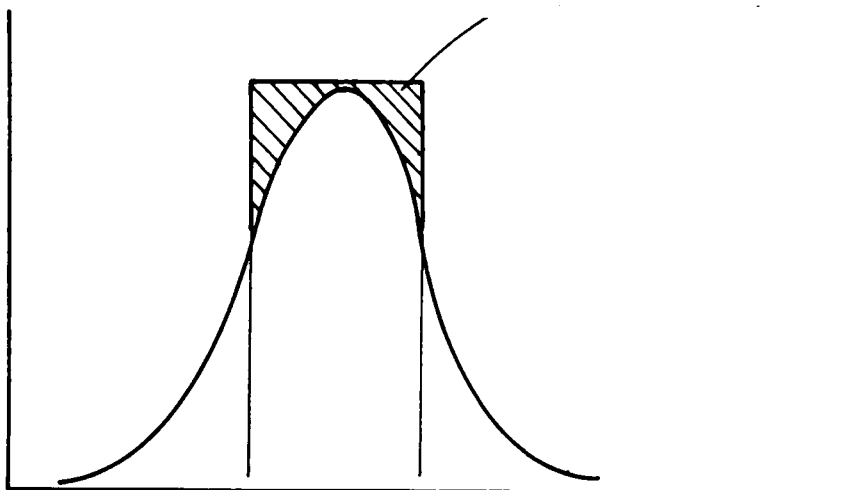


Figure (1-8). Effect of ballistic deficit on energy spectrum. The shape of the energy line is determined by the distribution of charge collection times.

Ballistic Deficit

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Ballistic Deficit

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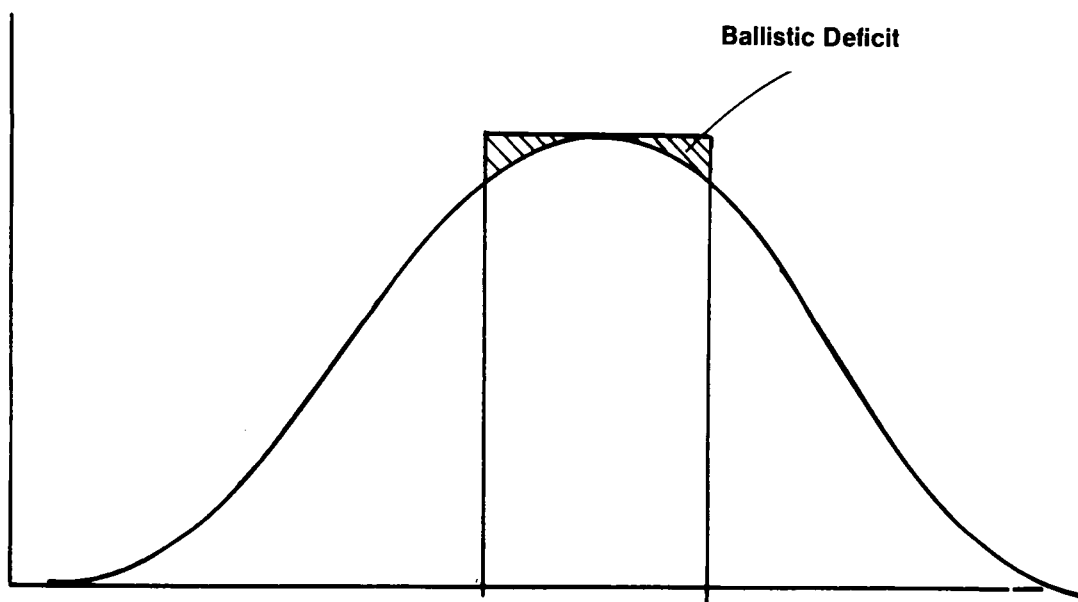


Figure (1-9). Effect of $t_{1/2}$ on ballistic deficit.

reducing the magnitude of the ballistic deficit. However, a longer $t_{1/2}$ also reduces the count rate capability of the spectroscopy amplifier since the width of each output pulse becomes larger.

Radeka proposed a time variant spectroscopy amplifier with a flat-topped impulse response, shown in figure (1-10), that would minimize ballistic deficit, but maintain high count rate performance [19]. The Radeka amplifier consisted of a time invariant semi-Gaussian prefilter followed by a gated integrator as illustrated in block diagram form in figure (1-11). The flat-top was achieved by having the integration time longer than the width of the prefilter pulse at the base [19]. Figure (1-12) depicts the convolution of the current waveform from figure (1-6) with the impulse response of figure (1-10). As long as the flat-top region of the impulse response of the Radeka amplifier is wider than T_{max} , no ballistic deficit will be present. The Radeka amplifier is now widely known as the gated integrator.

The gated integrator allows for good energy resolution at high count rates due to its relatively short overall width and its wide, flat top. However, the gated integrator is extremely sensitive to low-frequency noise (parallel and $1/f$ noise terms) and near dc baseline fluctuations [7]. Also, the noise performance of the semi-Gaussian amplifier is superior to the gated integrator at equivalent count-rate

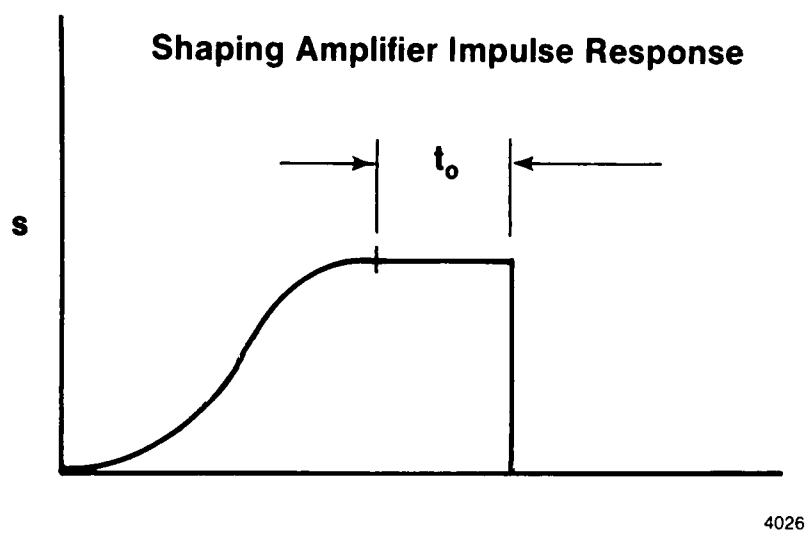
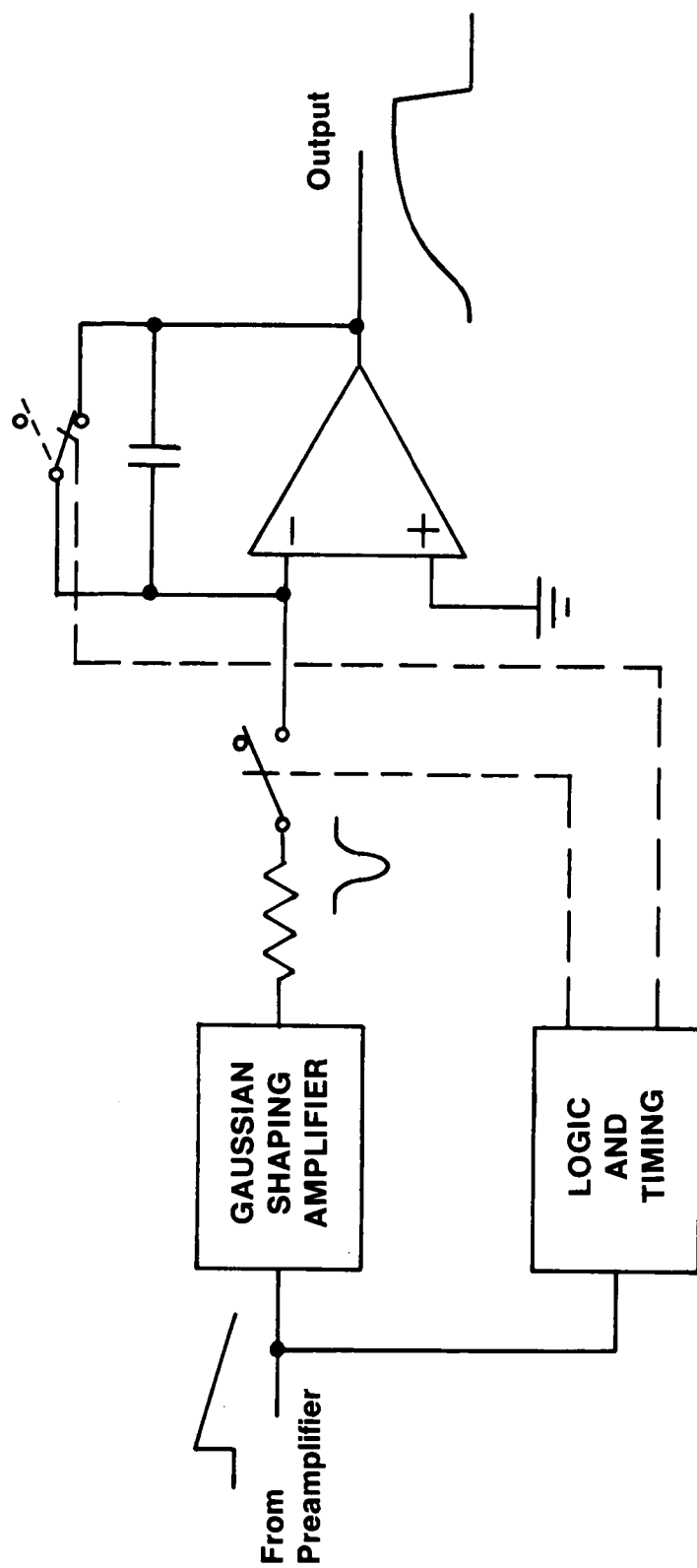


Figure (1-10). Impulse response of Radeka shaping amplifier.



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Figure (1-11). Block diagram of gated integrator.

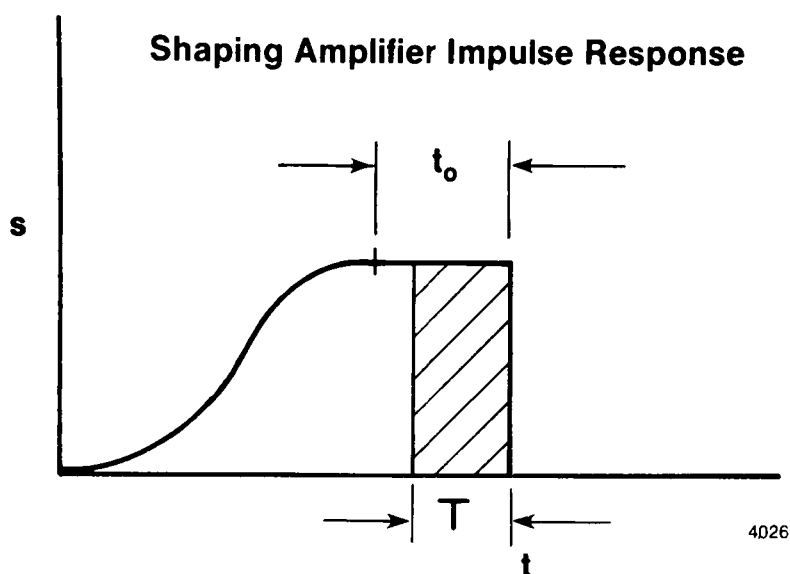


Figure (1-12). Convolution of impulse response of figure (1-10) with current signal of figure (1-6). Since the width of the flat top is greater than the width of the current signal, there will be no ballistic deficit.

capabilities. These factors led Goulding and Landis to develop a ballistic deficit correction circuit that retained the benefits of the semi-Gaussian amplifier [7].

Using the theoretical work of Loo, Goulding, and Gao [13] which showed that

$$G(t_d) \sim K(t_d)^2, \quad (1-14)$$

where K is a constant for a fixed shaping time, Goulding and Landis developed a circuit which calculated the ballistic deficit and added a correction to the spectroscopy amplifier output signal on a pulse-by-pulse basis. Figure (1-13) shows a block diagram of the Goulding and Landis circuit. The Goulding and Landis circuit, now known as the ballistic deficit correction (BDC) circuit, does not suffer from the disadvantages of the gated integrator, since the noise performance is that of the semi-Gaussian. However, Simpson et al. [24] have shown that the BDC circuit does add some noise to the semi-Gaussian, and does not make a complete correction for the ballistic deficit. Thus, neither the gated integrator nor the Goulding and Landis BDC circuit can be said to offer the best resolution under all conditions [24].

Hinshaw and Landis [10] also have reported a circuit which adds a correction to the spectroscopy amplifier output to compensate for ballistic deficit. The Hinshaw and Landis

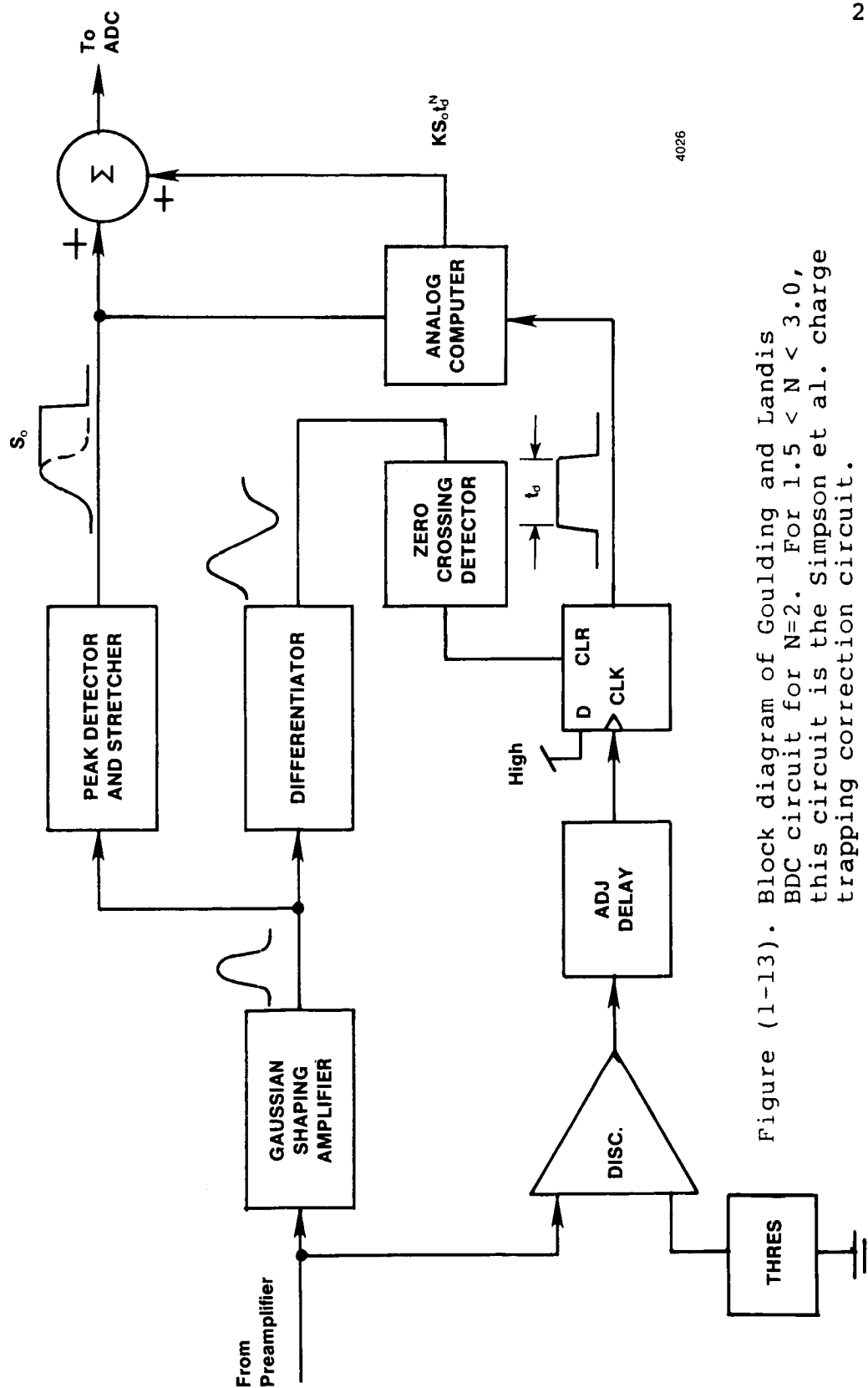


Figure (1-13). Block diagram of Goulding and Landis BDC circuit for $N=2$. For $1.5 < N < 3.0$, this circuit is the Simpson et al. charge trapping correction circuit.

circuit is based on the concept that the ballistic deficit is proportional to the difference between the peak amplitudes of two spectroscopy amplifier output pulses with different peaking times. This circuit is shown in block diagram form in figure (1-14).

The final energy line broadening mechanism to be considered in this work is deep-level charge trapping. Figure (1-15) depicts a cross sectional view of a coaxial detector with the electric field lines directed radially out from the center. Electrons generated in the detector will travel to the inner electrode, while the holes will travel to the outer electrode. This type of device, commonly referred to as a reverse electrode detector in the literature, is usually fabricated from N-type Ge, with electrons being the majority carrier.

Even the most pure Ge will contain deep-level trapping centers which can capture electrons from the conduction band and/or holes from the valence band. When a gamma-ray interacts in the detector, a non-equilibrium concentration of carriers is introduced into the Ge [2]. As the carriers travel across the detector, a certain fraction will be captured by the deep-level traps. If n_0 and p_0 are the number of electrons and holes generated by the gamma-ray, the number of carriers at time t (the interaction occurs at $t = 0$) is given by [2]

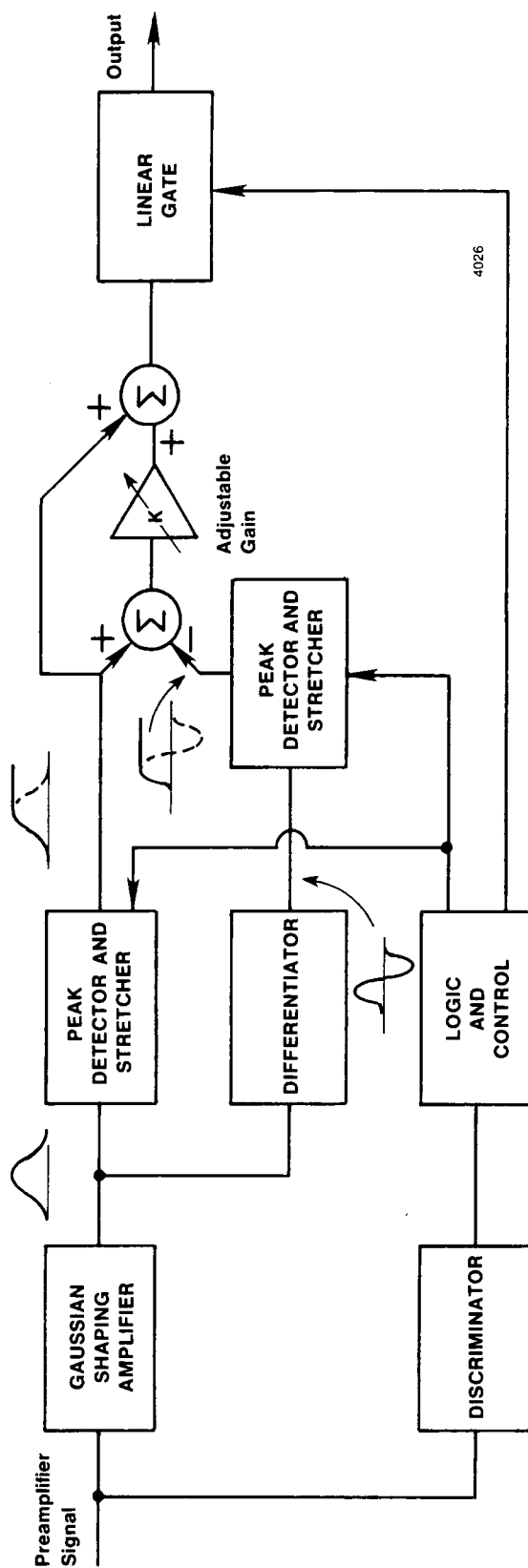


Figure (1-14). Block diagram of Hinshaw and Landis BDC circuit.

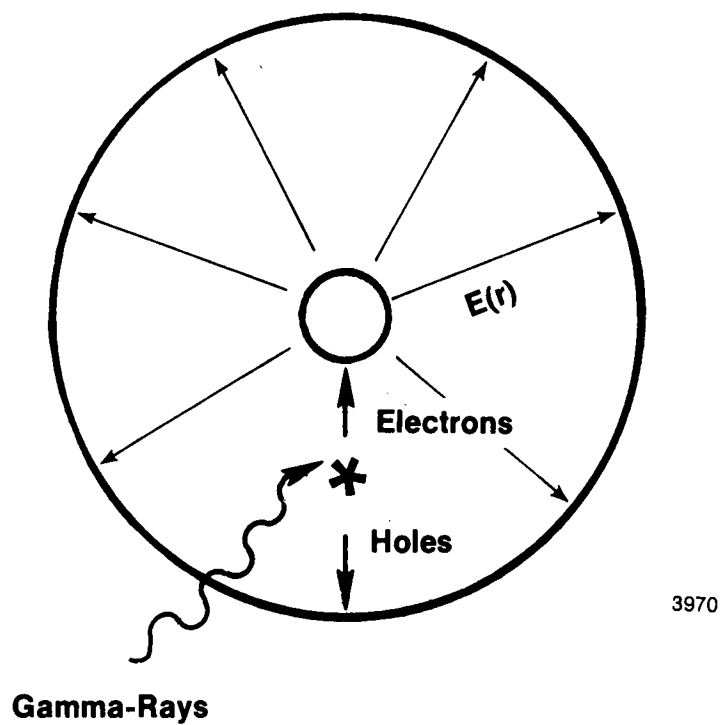


Figure (1-15). Cross section of coaxial, germanium, radiation detector.

$$n = n_0 \exp(-t/t_n) \quad (1-15a)$$

$$p = p_0 \exp(-t/t_p), \quad (1-15b)$$

where n is the number of electrons in the conduction band, p is the number of holes in the valence band, t_n is the majority carrier (electron) lifetime, and t_p is the minority carrier (hole) lifetime.

Consider the case where $t_p \gg t_n$. In this case, electron (majority carrier) trapping dominates over hole trapping and the charge carrier loss implied by equation (1-15b) can be neglected. As equation (1-15a) suggests, the deficit in the spectroscopy amplifier output due to majority carrier charge trapping is proportional to the number of electrons created by the gamma-ray interaction and to some function, U , of the electron charge collection time. Or

$$\begin{aligned} d &= S_0 U(t_{ce}) \\ &= S_0 U((r-r_1)/V_{sat}), \end{aligned} \quad (1-16)$$

where d is the deficit due to majority carrier charge trapping, t_{ce} is the electron charge collection time, r is the radius of interaction, r_1 is the inner radius of the detector, and V_{sat} is the saturation velocity of the electrons.

Energy line broadening is caused by majority carrier charge trapping in a manner analogous to the ballistic

deficit case. Interactions may occur at any r in the detector. Thus as equation (1-16b) shows, a range of deficits will result which will induce energy peak broadening. However, fluctuations in the number of carriers trapped for an interaction at a given radius also degrade the energy resolution [20]. This type of line broadening is much like that caused by charge generation fluctuations which were discussed earlier [20].

The case discussed above applies directly to the case of hole trapping in a conventional electrode detector if t_{∞} is replaced with t_{ch} , the hole charge collection time, and V_{sat} is the saturation velocity of holes in equation (1-16). In a conventional electrode detector, the field lines are directed radially in toward the center and holes travel to the inner electrode. The detector material in this case is usually P-type and holes are the majority carrier.

The argument above also applies to minority carrier trapping if equation (1-16) is rewritten as follows:

$$\begin{aligned} d &= S_0 W(t_{cm}) \\ &= S_0 W((r_2 - r)/V_{sat}), \end{aligned} \quad (1-17)$$

where W is a function that in general is not equal to U , r_2 is the outer radius of the detector, t_{cm} is the minority carrier (holes in reverse electrode and electrons in

conventional electrode detectors) collection time, and V_{sat} is the saturation velocity for the carrier being trapped.

Beginning with the theoretical work of Trammell [26], Simpson et al. [23] showed that

$$U(t_{cmj}) \sim A(t_{cmj})^M, \quad (1-18)$$

where A is a constant, t_{cmj} is the majority carrier collection time, and M is between 1.5 and 3.0 for most detectors. Simpson et al. also showed that

$$t_{cmj} \sim t_d, \quad (1-19)$$

for a large majority of gamma-ray interactions.

Substituting equations (1-18) and (1-19) into equation (1-16a) reveals

$$d \sim AS_0(t_d)^M, \quad (1-20)$$

for most gamma-ray interactions. Simpson et al. [23] reported a circuit, shown in figure (1-13), that corrected for majority carrier charge trapping deficit in a manner analogous to the BDC circuit of Goulding and Landis. In fact, for the case of $M=2$, the Simpson et al. circuit and Goulding and Landis circuit are identical and could be used to correct for ballistic deficit and majority carrier charge

trapping simultaneously. The BDC circuit and the charge trapping correction (CTC) circuit will be referred to collectively as resolution enhancement circuits in this work.

Like the BDC circuit, the CTC circuit also adds some noise to the system and does not perfectly correct for charge trapping deficit [24]. However, no other circuit has been shown to correct effectively for charge trapping deficit.

1.3 Scope and Organization of Dissertation

This work has the following three major goals: to fully develop and unify the theories of charge trapping and ballistic deficit; to develop an equation from which energy line shapes can be calculated for arbitrary charge trap distributions and spectroscopy amplifier shaping times; and to analyze and develop design equations for resolution enhancement circuits, including a complete noise analysis and treatment of non-ideal qualities of the correction. Extensive experimental data will also be presented.

Chapter 2 will discuss coaxial, HPGe detectors. This chapter will lay the groundwork for much of the development in later chapters, including the formulation of the theories of ballistic and charge trapping deficit. Also, starting

with the works of Trammell [26] and Trammell and Walter [27], an equation will be developed that will allow the calculation of energy line shapes for arbitrary charge trap distributions and spectroscopy amplifier shaping times.

Chapter 3 will deal with the design and analysis of BDC circuits. This chapter will include a thorough analysis of the noise introduced by the correction, and will show most of the correction noise for the Goulding and Landis circuit can be attributed to the time jitter in the t_0 measurement [24]. Also, a comparison will be made between the gated integrator and the two BDC circuits. This discussion will show that, in general, BDC circuits provides better energy resolution at the larger values of $t_{1/2}$, while the gated integrator allows for better resolution at the smaller values of $t_{1/2}$.

Chapter 4 will contain a development of the Simpson et al. CTC circuit. The noise added as a result of this correction will also be shown to arise from jitter in the t_0 measurement. The subject of minority carrier charge trapping correction will be addressed in this section. Minority carrier charge trapping occurs in the important case of radiation damaged, reverse electrode detectors [20], and was treated theoretically by Raudorf and Pehl [20]. Also, Simpson et al. [23] proposed a circuit to correct for minority carrier charge trapping. This circuit will also be analyzed in chapter 4.

Unlike the BDC circuit, there is no competing instrument to compare with the CTC circuit. However, an instrument which uses a gated integrator to correct for ballistic deficit and a CTC circuit to correct for majority carrier charge trapping will be presented in chapter 4.

Chapter 5 will be devoted to experimental data collected using the various instruments discussed in the previous chapters. Spectra collected with conventional and reverse electrode detectors of several sizes will be presented along with spectra calculated using the methods of chapter 2. The energy range of the experimental spectra is from 100 keV to 2.6 MeV, while the calculations were performed over the 100 keV to 10 MeV range. The spectra were collected using semi-Gaussian spectroscopy amplifiers with peaking times from 1 μ S to 12 μ S and with gated integrators whose semi-Gaussian prefilters peaking times varied from .5 μ S to 12 μ S. Corrected and uncorrected spectra are presented for each detector studied. In this chapter, the BDC circuits are experimentally compared with the gated integrator. Also, spectra collected with a gated integrator which contains a CTC circuit are shown.

Finally, chapter 6 discusses the results and presents some conclusions.

CHAPTER 2

COAXIAL, HPGe, GAMMA-RAY DETECTORS

The quality of the data that is collected by any radiation spectrometer is ultimately determined by the properties of the detector. Coaxial, HPGe detectors have become the most frequently used devices for high-resolution, high-efficiency, gamma-ray spectroscopy [12]. This chapter will describe coaxial HPGe detectors, discuss the interaction of gamma-rays in these devices, and derive the output signals which occur due to these interactions. Also, the phenomena of ballistic deficit and charge trapping that were briefly treated in Chapter 1, will be presented in greater detail in this chapter. Finally, the Trammell-Walter relationship, which allows energy line shapes to be calculated, will be presented.

2.1 Coaxial HPGe Detector Fundamentals

Figure (2.1) shows cross sectional views of the two most commonly used coaxial, HPGe detectors. As this figure illustrates, these detectors are not truly coaxial, but have a closed-end configuration. The closed-ended geometry is preferred to avoid leakage currents at the front surface of the device [12]. However, true coaxial geometry is usually

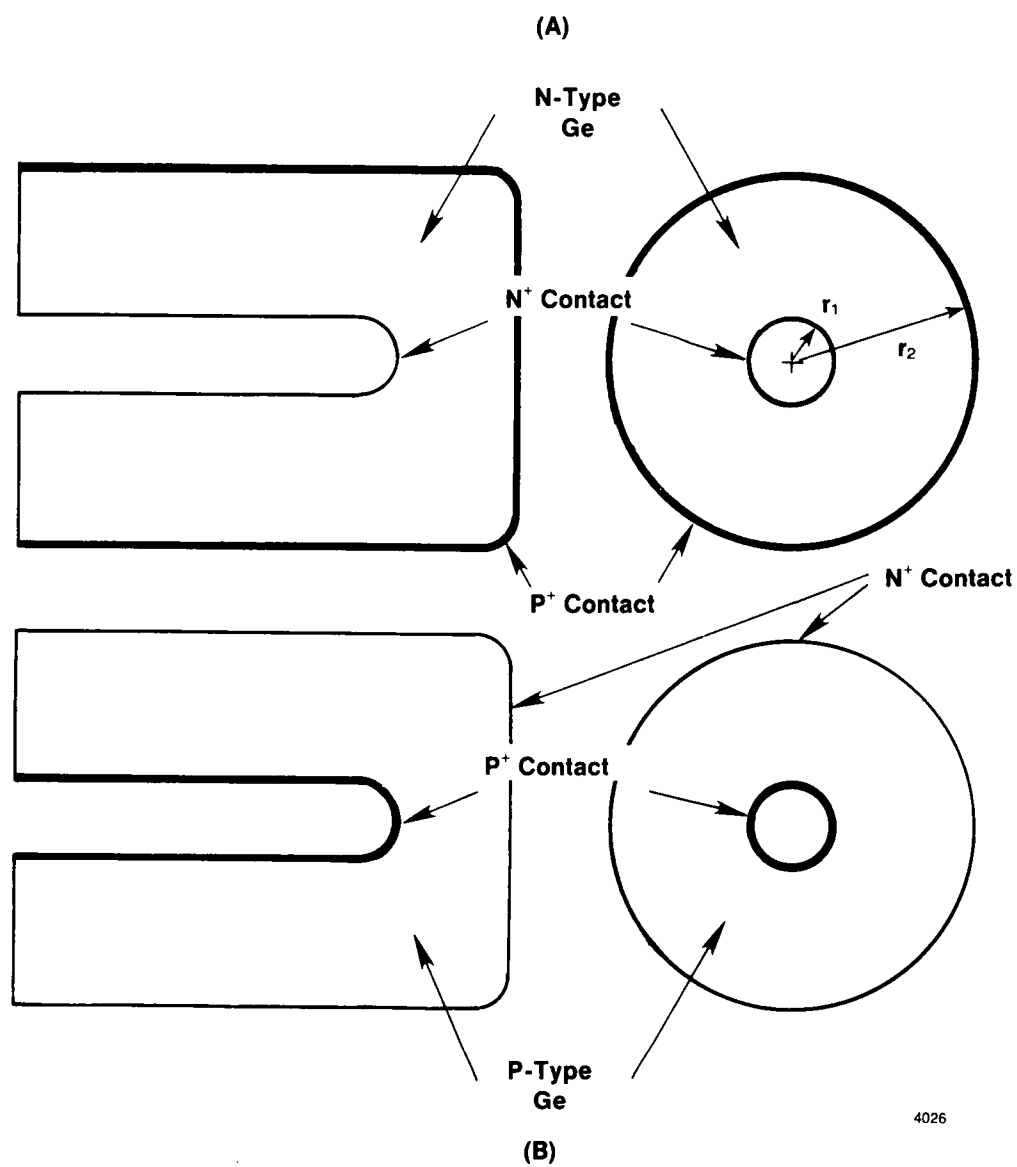


Figure (2-1). Cross-sectional views of (A) reverse electrode and (B) conventional electrode coaxial HPGe detectors.

assumed when performing calculations, to simplify the analysis. The detector may consist of either n-type or p-type material. In either case, the impurity concentration is very small (as the name would imply) when compared to doping levels used in the integrated circuit industry.

A p-n junction is formed in the p-type detector by evaporating and diffusing an n-type dopant, such as lithium, into the outer surface of the detector. This n-type diffusion is usually highly doped, and an ohmic contact can be made to the diffusion to form one electrode of the device. The other electrode is formed by diffusing a highly doped, p-type region into the inner surface of the detector where an ohmic contact is made. The resulting device meets the criteria of a one-sided junction diode [31]. This p-type configuration is often referred to in the literature as a conventional electrode detector.

In an analogous manner, a p-n junction is formed by implanting a p-type dopant, such as boron, into the outer surface of an n-type detector. This p-type region is highly doped, and an ohmic contact can be made to the diffusion to form one electrode of this detector. The other electrode is formed by diffusing a highly doped, n-type region into the inner surface of the detector where an ohmic contact is made. This one-sided junction diode is referred to as a reverse electrode detector.

For either case, when a reverse bias is applied to the p-n junction, a depletion region will form. However, nearly all of the depleted area will reside in the bulk of the detector due to its extremely light doping level compared to the highly doped outer electrode. If a large enough reverse bias, V_D , is applied to the detector, the depletion region will extend throughout the lightly doped material, and an immobile bulk charge will exist in the active region of the detector. For biases greater than or equal to V_D , the electric field, $E(r)$, in the detector is given by [12]

$$E(r) = (qN/(2K))r + [V + (qN/4K)(r_2^2 - r_1^2)]/(r \ln(r_2/r_1)), \quad (2-1)$$

where true coaxial geometry has been assumed, q is the charge of an electron, N is the impurity concentration in the lightly doped region of the detector, K is the dielectric constant of Ge, r_1 is the inner radius of the detector, and r_2 is the outer radius of the detector. The voltage needed to completely deplete the detector, V_D , is found by setting $E(r_1) = 0$ in equation (2-1), which yields

$$V_D = qN/(2K)[r_1^2 \ln(r_2/r_1) - (1/2)(r_2^2 - r_1^2)]. \quad (2-2)$$

For a detector with $r_1 = .5$ cm, $r_2 = 3$ cm, and a desired

depletion voltage of 2000 V, the impurity concentration from equation (2-2) is $9 \times 10^9 \text{ cm}^{-3}$.

The efficiency of HPGe detectors can be specified in several manners. For example, the efficiency of a detector can be defined as the number of counts which get placed in an energy peak by the spectrometer divided by the number of gamma-rays of that energy emitted by the radiation source. This efficiency, then, depends both upon the location and the energy of the radiation source. However, HPGe detectors are usually specified by the relative efficiency, which is defined as the number of counts placed in the 1.33MeV energy peak by the HPGe spectrometer divided by the number of events that a 3"X3"X3" NaI scintillation spectrometer would place in the same energy peak. The radiation source is placed 25cm away from the detectors in both measurements. It should be noted that relative efficiencies greater than 100% are possible.

2.2 Gamma-Ray Interactions in Ge

To detect and quantify a gamma-ray, an electrical signal must be produced when the gamma-ray interacts in the detector. In HPGe detectors, this signal consists of a number of electron-hole pairs that is linearly proportional to the energy of the incident radiation. However, this

charge can be generated by three methods. The dominant charge generation mechanism is a function of the energy of the interacting gamma-ray.

At low energies, the dominant charge production mechanism is photoelectric absorption. Photoelectric absorption occurs when the gamma-ray interacts with a Ge atom releasing an energetic electron from a tightly bound shell of the atom. In this process, the photon is completely absorbed, thus the energy of the photoelectron, E_{e^-} , is [12]

$$E_{e^-} = E_G - E_b, \quad (2-3)$$

where E_G is the energy of the incident gamma-ray and E_b is the binding energy of the electron in its original atomic shell. The interacting Ge atom may also emit an x-ray or an Auger electron as it fills the vacancy left by the photoelectron [12]. However, the x-ray is usually re-absorbed in a photoelectric process [12]. The electron-hole pairs are generated by collisions of the energetic photoelectron with Ge atoms. It is usually assumed that all the electron-hole pairs are produced in a region very near the original interaction [12] (i.e. all the charge is produced at one point in the detector).

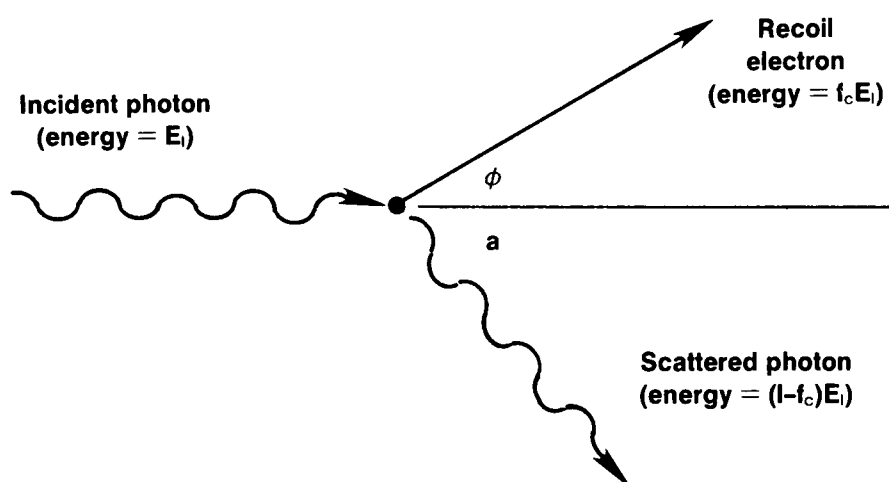
At higher energies, Compton scattering becomes the dominant mechanism for the transfer of energy from the

gamma-ray to the detector. As mentioned in Chapter 1, Compton scattering is the process whereby a portion of a gamma-ray's energy is transferred to an electron in the detector, and the gamma-ray is deflected through an angle a . This is shown graphically in figure (2-2). The fraction of energy transferred to the electron, f_c , is given by [12]

$$f_c = 1 - 1/(1 + E_G/.511(1 - \cos(a))), \quad (2-4)$$

where E_G is in MeV. As equation (2-4) shows, f_c is between 0 and a fraction less than one. Again, this energetic electron will produce electron-hole pairs by collisions with Ge atoms.

After this initial Compton scattering, several things may occur. (1) The scattered gamma-ray may exit the detector without depositing the remainder of its energy, thus contributing to the Compton scattered background; (2) the gamma-ray may be photo-absorbed, depositing the remainder of its energy and producing a photoelectron; or (3) the gamma-ray may experience another Compton interaction, depositing more of its energy in the detector. Process (3) will continue until the final scattered gamma-ray is photo-absorbed, or travels out of the detector. Any time a scattered gamma-ray escapes the detector, the event becomes part of the Compton scattered background and is of no experimental use. Therefore, those events will not be



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Figure (2-2). Graphical representation of Compton scattering.

considered further here. If process (2) occurs, the gamma-ray interaction takes place in two locations in the detector nearly simultaneously. If process (3) occurs, the gamma-ray interaction takes place in three or more locations at nearly the same time.

Figure (2-3) depicts the probability distribution function for the gamma-ray scattering angle, α . As this figure shows, at low energies, the scattering is almost isotropic. However, as the energy increases, small angle, forward scattering becomes the most likely occurrence. As figure (2-3) illustrates, closely spaced interaction radii for successive Compton scattered events is favored by this anisotropy if the entrance angle of the gamma-ray is normal to the front surface of the detector.

Finally, if the energy of the incident gamma-ray is greater than twice the rest mass of an electron (1.022 MeV), the production of an electron-positron pair by the annihilation of the gamma-ray is possible [12]. The energy of the gamma-ray in excess of the rest mass of the electron-positron pair is imparted to the particles as kinetic energy. An electrical signal is produced by the electron-positron pair in the following manner. Electron-hole pairs are produced by collisions of the electron and positron with Ge atoms in the detector. This will produce a signal that is proportional to the original gamma-ray energy minus twice the rest mass of the electron. However, after the positron

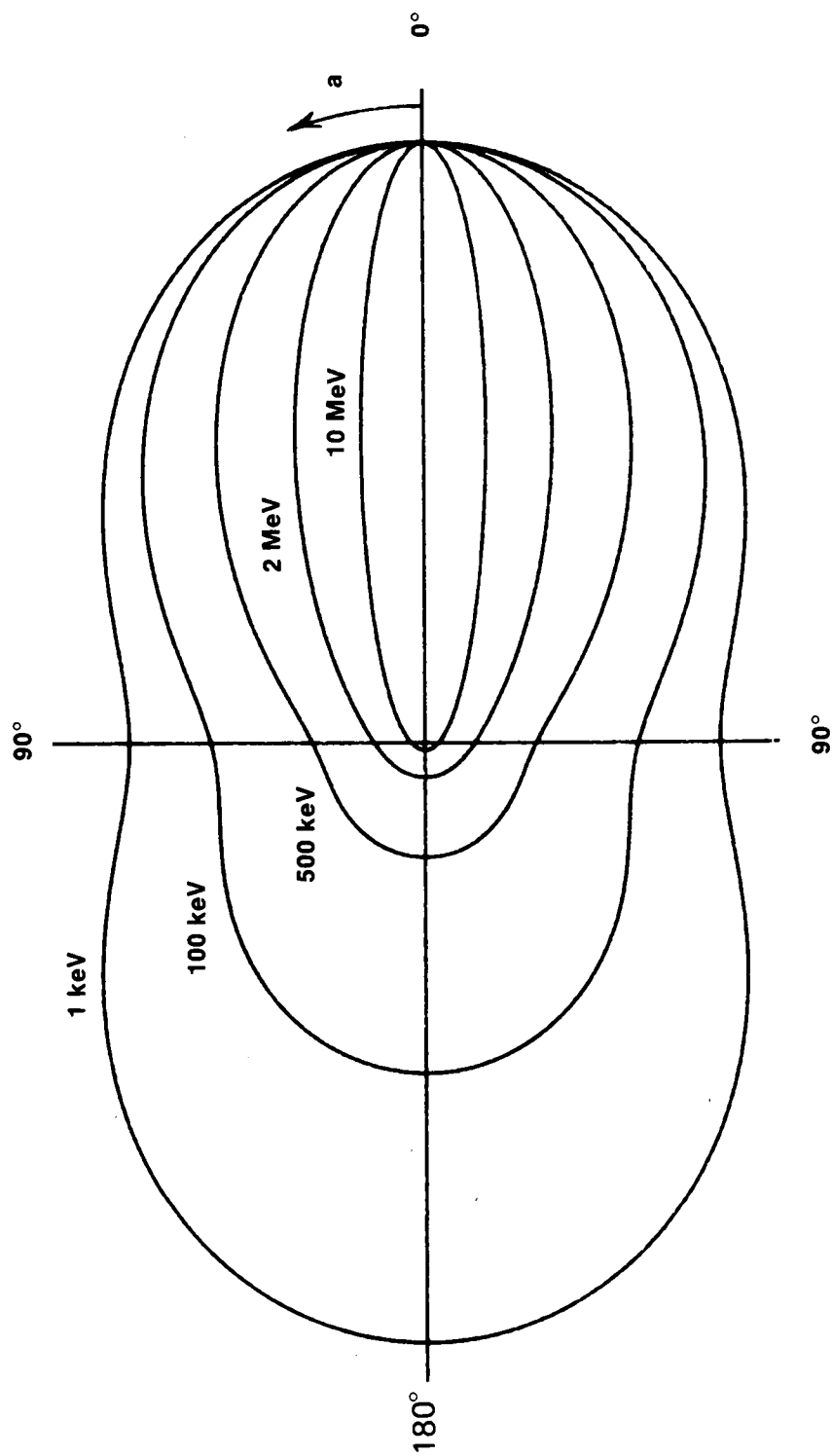


Figure (2-3). Probability distribution function for gamma-ray scattering angle in Compton scattering.

is brought to rest, an electron-positron annihilation will occur, producing two 511 keV gamma-rays whose directions of travel are usually 180° apart. These gamma-rays may interact with the detector in any of the methods discussed above (except the energy is too low for pair production), or either or both may escape from the detector without interacting. However, if both gamma-rays deposit all of their energy in the detector, the entire energy of the original gamma-ray would be deposited in the detector.

Each of the three major interactions mechanisms described above is dominant over some range of energies, as figure (2-4) shows. The single photoelectric interaction method is dominant for energies below ~ 140 keV [12], while pair production is dominant only at energies above 1.02 MeV. Therefore, for the majority of gamma-ray spectroscopy (100 keV to 10 MeV), the dominant interaction mechanism is multiple Compton scattering.

2.3 Electrical Signals Produced by Coaxial, HPGe Detectors

The charge produced by the interaction of a gamma-ray in a detector will result in a current signal at the electrodes of the detector. This signal will be processed by the preamplifier and the spectroscopy amplifier, and will eventually be counted in a channel (energy bin) of the MCA.

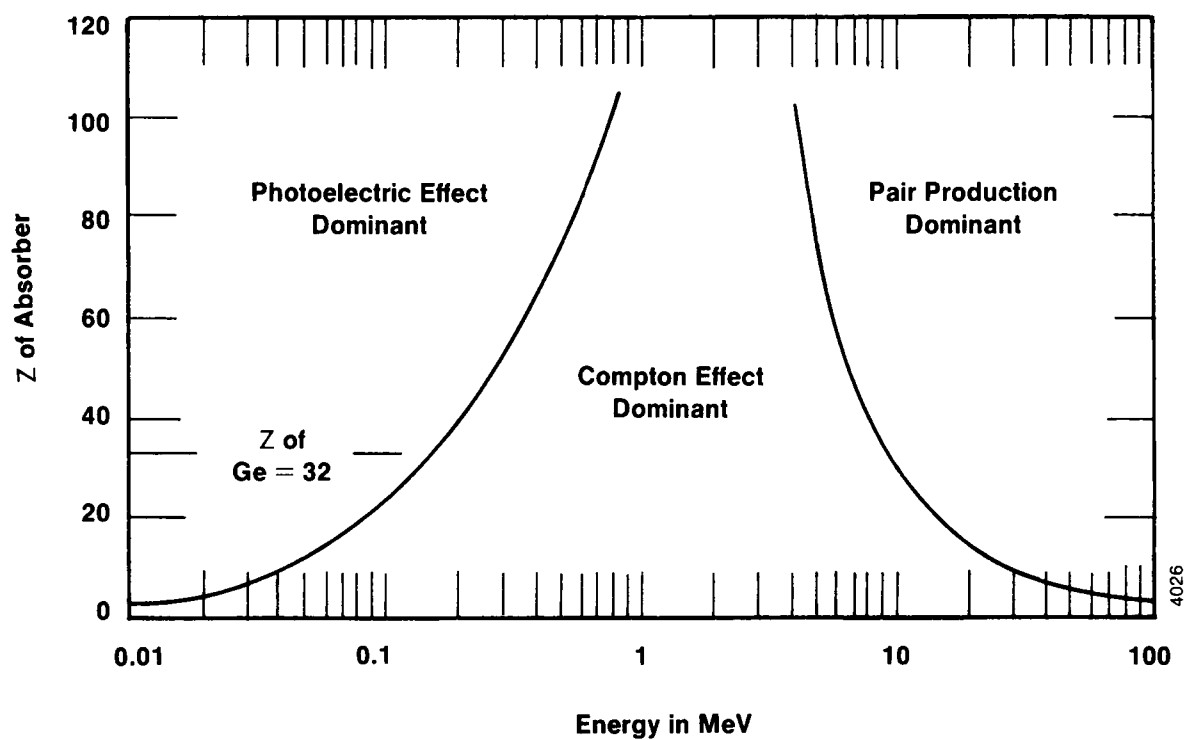


Figure (2-4). Dominant gamma-ray interaction mechanisms vs. energy.

To accurately predict the response of the spectroscopy system, however, the shape of the current waveform must be known.

To make the calculation of the current waveforms tractable, it is helpful to make the following assumptions [12]:

1. All the charge carriers are created at a fixed location in the detector. This assumption neglects the finite range of the secondary electrons and multiple interactions. However, since the detector is a linear device, superposition can be used to calculate the waveform resulting from a multiple interaction event.
2. Any charge carrier that is captured in a trap remains trapped for a time much greater than the charge collection time.
3. All charge carriers are created in the active volume of the detector where the electric field has its full expectation value.
4. The electric field in the detector is large enough to cause saturation of the drift velocity for

charge carriers of both polarities, and these carriers reach this velocity in a negligible amount of time.

5. True coaxial geometry is assumed.
6. The n-type detector and the conventions shown in figure (2-1) are used for this calculation. The same analysis can be used for p-type detectors if the proper polarities are used.

When a gamma-ray interacts in the detector, a quantity of charge is created at a fixed radius (see assumption 1 above). An immobile space charge exist throughout the detector due to the ionized impurities, and the electric field is given by equation (2-1). The charge carriers reach their saturation velocity very quickly (see assumption 4 above), and travel toward their respective electrodes. To determine the current, it is necessary to calculate the rate of change of the charge induced on the electrodes due to the electrons and the holes moving at the saturation velocity, v_{sat} . This problem was treated by Gunn [9], who found that

$$i_h(t) = q_h v_{sat} U(t_{ch} - t) dE/dV \quad (2-5a)$$

$$i_e(t) = -q_e v_{sat} U(t_{ce} - t) dE/dV, \quad (2-5b)$$

where $i_h(t)$ is the component of the current due to holes, $i_e(t)$ is the component of current due to electrons, q_h is the charge carried by holes, q_e is the charge carried by electrons, v_{sat} is assumed equal for electrons and holes, U is the unit step function, t_{ch} is the charge collection time for holes, t_{ce} is the charge collection time for electrons, and equation (2-5) is evaluated at $V = V_{op}$, the bias voltage of the detector. The total current is the sum of equations (2-5a) and (2-5b). Substituting equation (2-1) into equation (2-5) shows that

$$i_h(t) = q_h v_{sat} (1/(r \ln(r_2/r_1))) U(t_{ch}-t) \quad (2-7a)$$

$$i_e(t) = -q_e v_{sat} (1/(r \ln(r_2/r_1))) U(t_{ce}-t) \quad (2-7b).$$

If the time dependence of the radius in equation (2-7) is included explicitly, the result is

$$i_h(t) = q_h v_{sat} (1/((r_i + v_{sat} t) \ln(r_2/r_1))) U(t_{ch}-t) \quad (2-8a)$$

$$i_e(t) = -q_e v_{sat} (1/((r_i - v_{sat} t) \ln(r_2/r_1))) U(t_{ce}-t), \quad (2-8b)$$

where r_i is the interaction radius. Figure (2-5) shows example current waveforms calculated from equation (2-8).

The amount of charge collected is the integral of equation (2-8) with respect to time. Since a charge

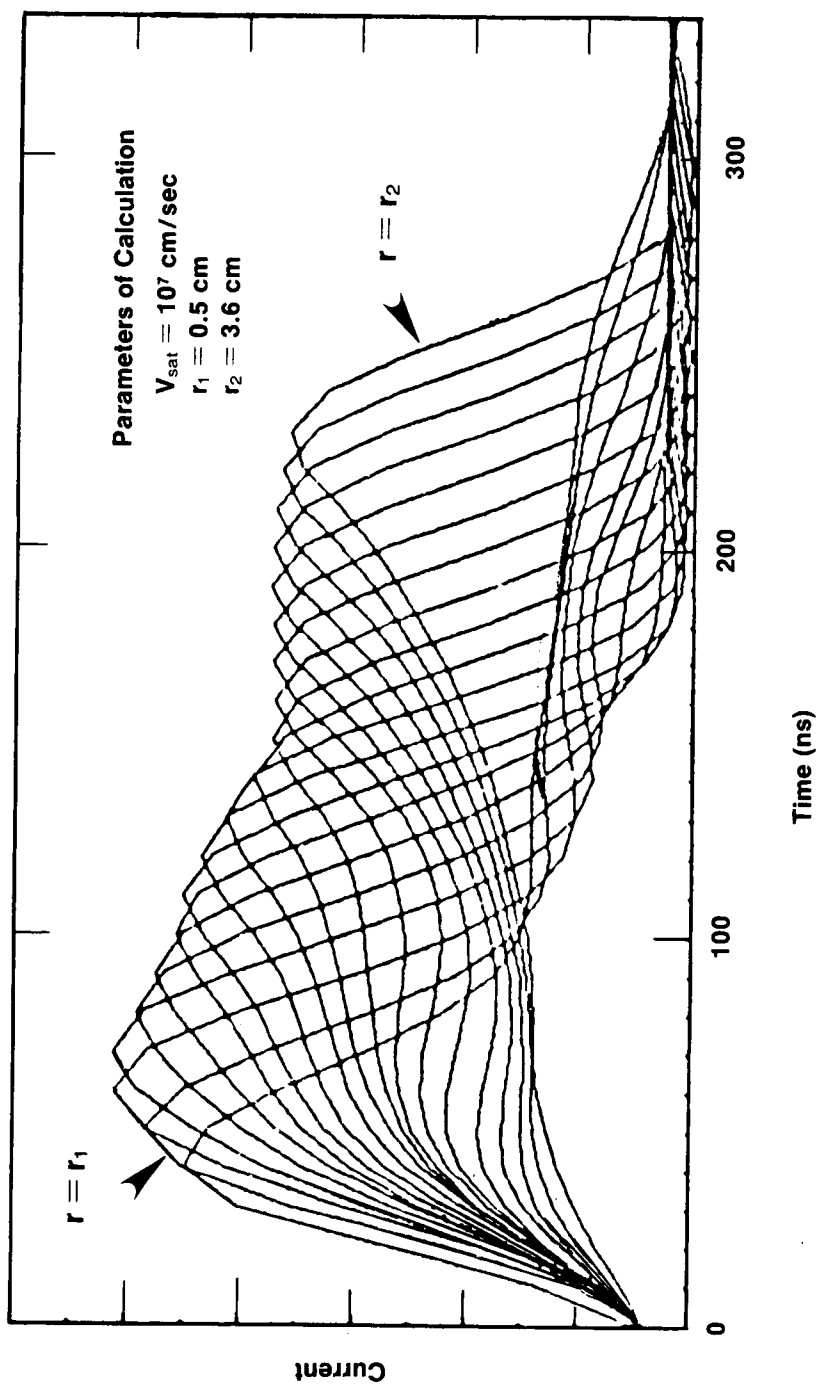


Figure (2-5). Current waveforms calculated with equation (2-8).

sensitive preamplifier is ideally an integrator, its output is proportional to this charge signal. These preamplifier signals have been studied both analytically and experimentally by several authors [21,22]. Figure (2-6) shows calculated preamplifier signals for the case of single and double interactions.

2.4 Ballistic Deficit

As described in Chapter 1 and shown in Section 2.3, the charge created by a gamma-ray interacting in the detector is not collected instantaneously. This leads to a ballistic deficit that was described qualitatively in Chapter 1. In this section, ballistic deficit will be treated in more detail. A simplified analytical analysis will be performed, along with more exact PSpice [28] simulations.

The current signals shown in figure (2-5) are inputs to a linear system composed of the preamplifier and the spectroscopy amplifier. The response of this system to these stimuli, then, is given by the convolution integral

$$V_o(t) = \int_0^{\infty} i(z)h(t-z)dz, \quad (2-9)$$

where $V_o(t)$ is the output of the spectroscopy amplifier, i is the input current waveform, and h is the impulse response

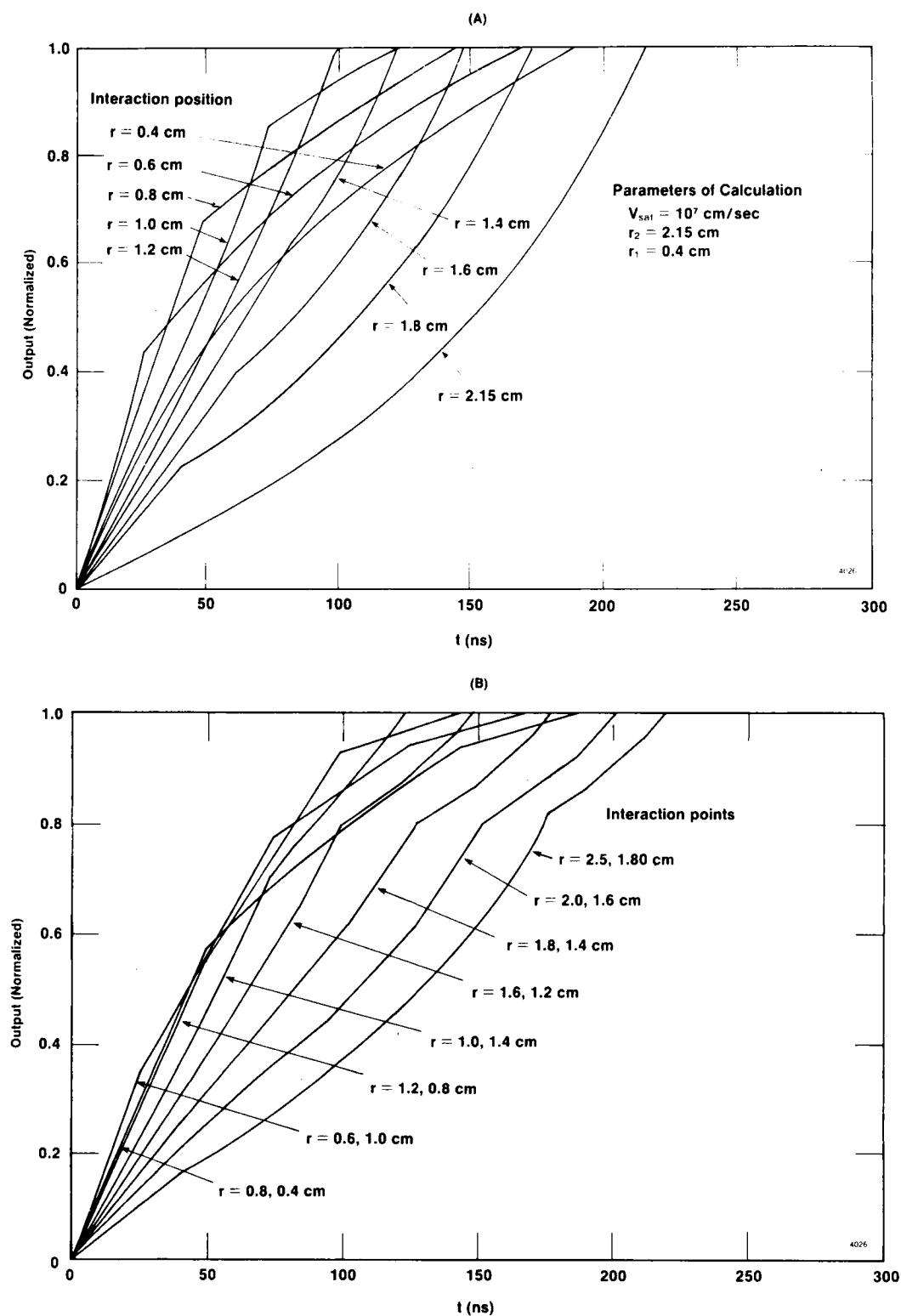


Figure (2-6). Calculated preamplifier signals for (A) single and (B) double interaction point cases.

of the combined preamplifier and spectroscopy amplifier. The calculation of equation (2-9) was shown graphically in Chapter 1.

To solve equation (2-9) analytically, it is helpful to model the current waveforms as a simple rectangular function as shown in figure (1-7) in chapter 1. This problem was addressed by Loo, Goulding, and Gao [13], who found that

$$V_o(t) = t_1/T \left\{ 1 - \exp(-t/t_1) \left[1 + \sum_{M=1}^N t^M / (M! t_1^M) \right] \right. \quad (2-10a) \\ \left. - [1 - \exp((T-t)/t_1) - \exp((T-t)/t_1) \right. \\ \left. \times \sum_{M=1}^N (t-T)^M / (M! t_1^M) \right] U(t-T) \right\}$$

$$t_p = T / (1 - \exp(-T/Nt_1)) \quad (2-10b)$$

where t_1 is the time constant of the integration stages, T is the duration of the detector current pulse, t_p is the peaking time of the semi-Gaussian spectroscopy amplifier and N is the number of integrators in the semi-Gaussian filter. While equation (2-10) is exact for the case of a rectangular current waveform, it is cumbersome to use. Thus, beginning with the work of Baldinger and Franzen, Loo, Goulding, and Gao also developed the following simplified equation for the relative ballistic deficit (RBD):

$$\text{RBD} = (N/24) (T/t_p)^2, \quad (2-11)$$

where t_p is the peaking time of the spectroscopy amplifier and semi-Gaussian shaping has been assumed. Figure (2-7) shows the relative ballistic deficit predicted by equation (2-11) as a function of the radius of interaction. In this figure, the current waveform was assumed to be a rectangular function whose width was taken to be the longer of the collection time for the holes or the collection time for the electrons.

Equation (2-11) suggests that a pulse-by-pulse correction could be made for ballistic deficit if the duration of each current waveform could be measured. Goulding and Landis developed a correction circuit based on equation (2-11). Ballistic deficit correction will be the topic of Chapter 3.

Equations (2-10) and (2-11) are extremely useful for calculating approximations to the ballistic deficit. However, for more exact calculations, actual current waveforms like those shown in figure (2-6) must be used in equation (2-9). It would be very difficult and time consuming to solve this equation analytically for arbitrary current waveforms. However, excellent computer simulations can be performed by approximating the current waveforms of figure (2-6) as piecewise linear functions. Figure (2-8) shows the ballistic deficit versus the interaction radius calculated for a seventh-order, semi-Gaussian filter using the circuit simulation program PSpice [28] with the

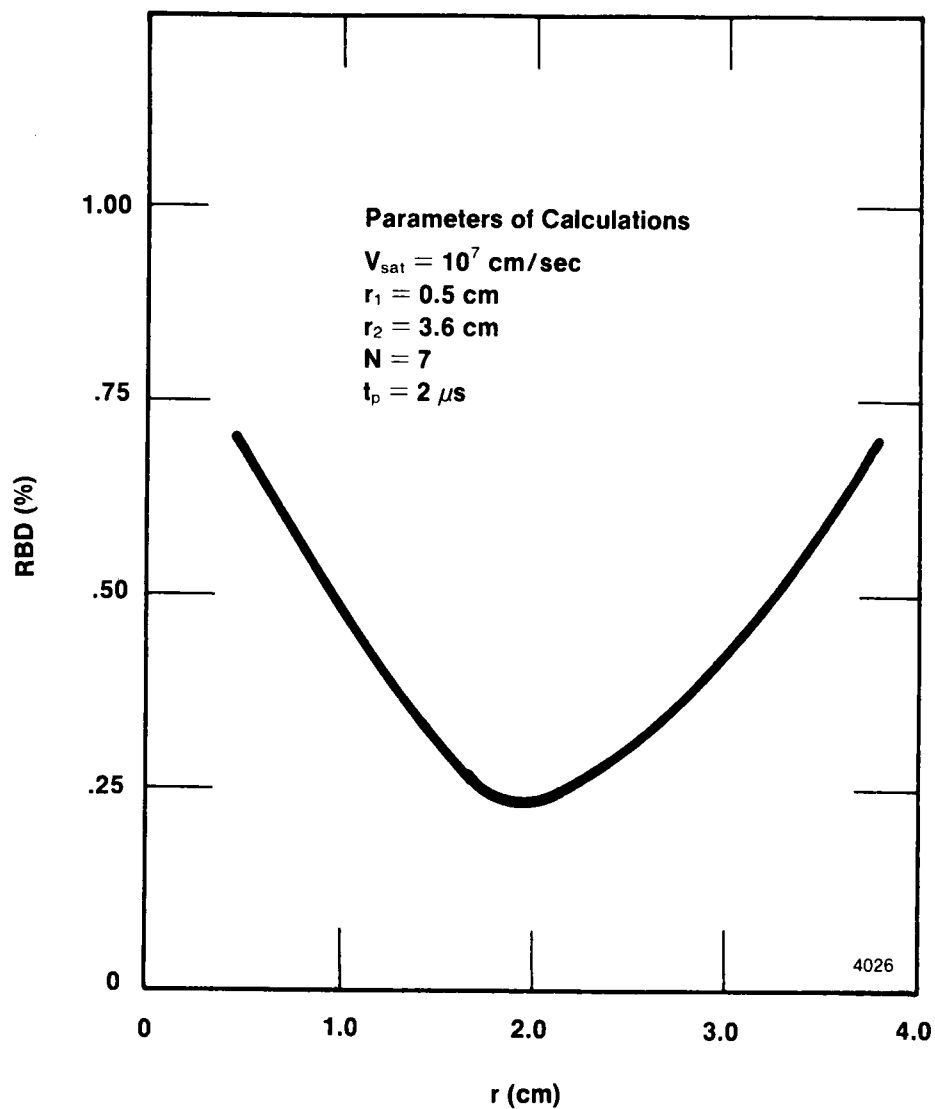


Figure (2-7). RBD vs. radius of interaction as predicted by equation (2-11).

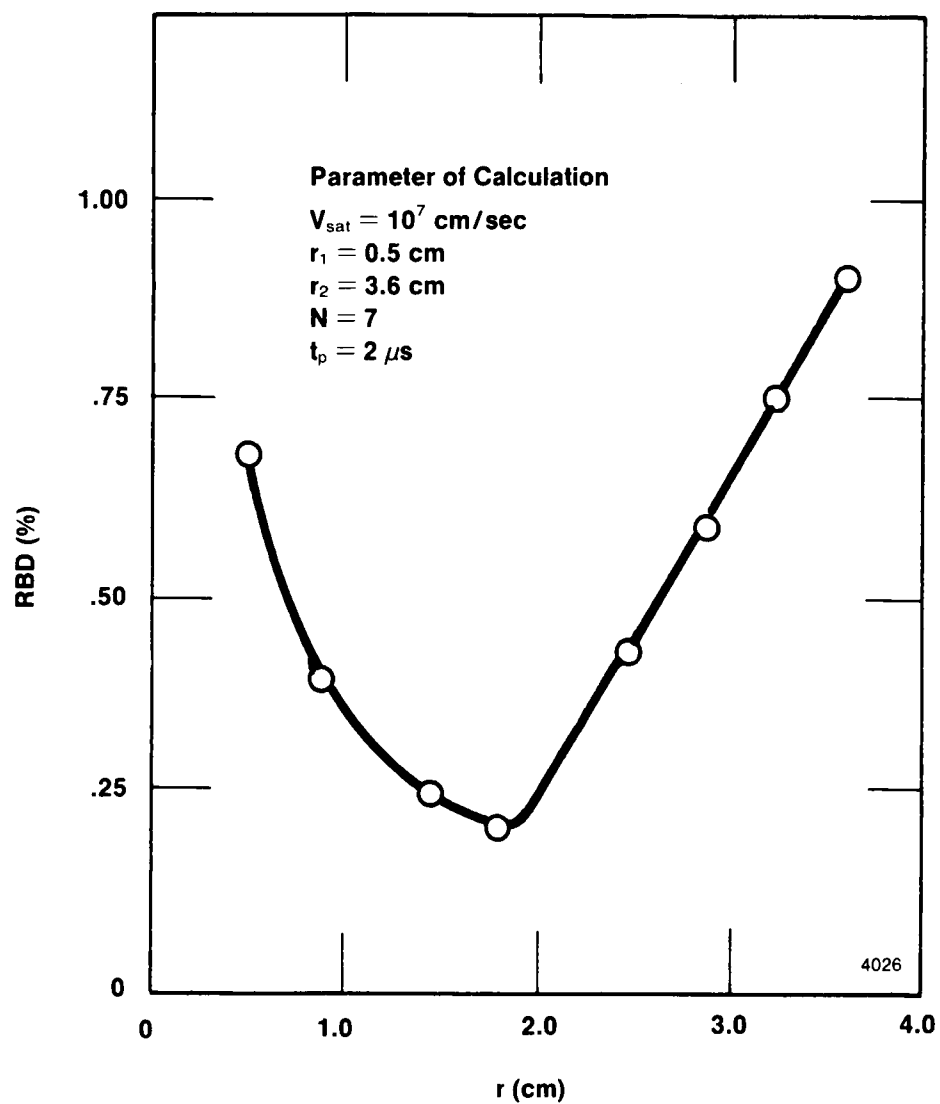


Figure (2-8). RBD vs. radius of interaction as calculated with PSpice.

waveforms of figure (2-6) as inputs. A comparison between figures (2-7) and (2-8) reveal that both curves are similar in shape. However, the magnitude of the ballistic deficit as predicted by the computer simulation is greater than that predicted by equation (2-11). Also, the PSpice calculations show that the ballistic deficit is worse at the outer radius than it is at the inner radius, while equation (2-11) predicts that the ballistic deficit is equal for these two cases. Since the computer simulations used actual preamplifier waveforms as inputs, figure (2-8) should be more accurate.

2.5 Charge Trapping

The equations developed in section 2.3 for the detector current assumed that none of the charge generated by a gamma-ray interaction was impeded from reaching the electrodes. However, as described in Chapter 1, even the purest Ge will contain charge traps that capture electrons and holes as they travel through the detector. This section will describe the three ways in which charge trapping can degrade the energy resolution of HPGe detectors. Methods of charge trapping correction will be treated in Chapter 4.

Trammell and Walter gave the first accurate theoretical description of the effects of charge trapping on energy

resolution in 1969 [27]. They considered the case of a planar, lithium-drifted, Ge detector, like the one shown in figure (2-9). An uniform distribution of traps with long detrapping times is assumed to exist throughout the active area of this detector. For an interaction that occurs at x_0 and produces N_0 electron-hole pairs, the mean number of actual charge carriers induced on the electrodes is reduced by trapping to [27]

$$N = n(x_0)N_0, \quad (2-12)$$

where $n(x_0)$ is defined as the charge collection efficiency. The probability that an electron will travel a distance x and be trapped in dx is [27]

$$P_{nT}(x) = (1/L_n) \text{EXP}(-x/L_n) dx, \quad (2-13)$$

where L_n is the mean free path before being trapped for an electron. The amount of induced charge lost due to the trapping of this electron is [27]

$$F(x, x_0) = q(x_0 - x)/d, \quad (2-14)$$

where d is the width of the detector as shown in figure (2-9). Thus, the mean fraction loss of induced charge due to electron trapping for an interaction occurring at x_0 is [27]

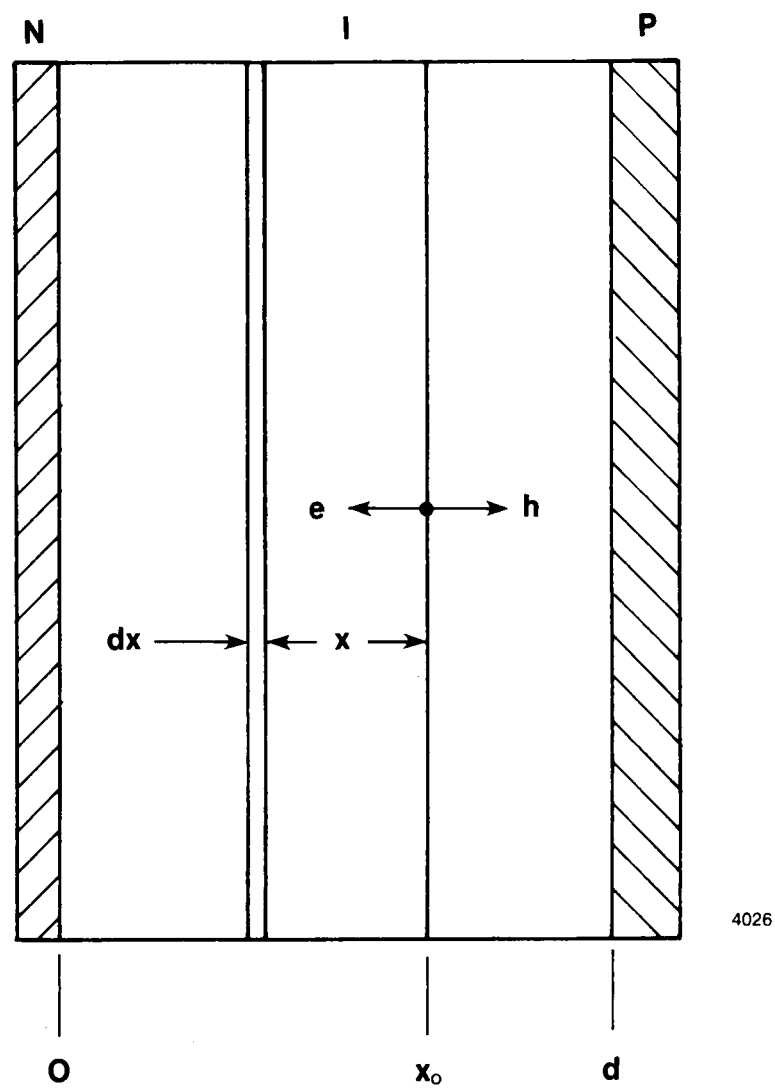


Figure (2-9). Planar, Li-drifted, Ge detector.

$$\begin{aligned}
 d_n(x_0) &= \int_0^{x_0} (1/L_n) \exp(-x/L_n) \{(x_0-x)/d\} dx \\
 &= \{(x_0/d) - (L_n/d)(1 - \exp(-x_0/L_n))\}.
 \end{aligned}
 \tag{2-15}$$

In a similar manner, the mean fractional loss due to hole trapping for an interaction occurring at x_0 is

$$d_p(x_0) = \{(d-x_0)/d - L_p/d(1 - \exp(-(x_0-d)/L_p))\}
 \tag{2-16}$$

where L_p is the mean free path before trapping for a hole. Thus, the charge collection efficiency is [27]

$$\begin{aligned}
 n(x_0) &= 1 - d_n(x_0) - d_p(x_0) \\
 &= (L_n/d)\{1 - \exp(-x_0/L_n)\} \\
 &\quad + (L_p/d)\{1 - \exp(-(d-x_0)/L_p)\}
 \end{aligned}
 \tag{2-17}$$

This result was derived by another method in reference [32].

Equation (2-17) shows that for each possible interaction position, $0 \leq x \leq d$, a different n exist. This gives rise to a distribution of the measured energy of the interacting gamma-ray that is a function of x . For example, consider the important case where $L_p \gg L_n$. In this case,

$$n(x) \sim 1 - d_n(x).
 \tag{2-18}$$

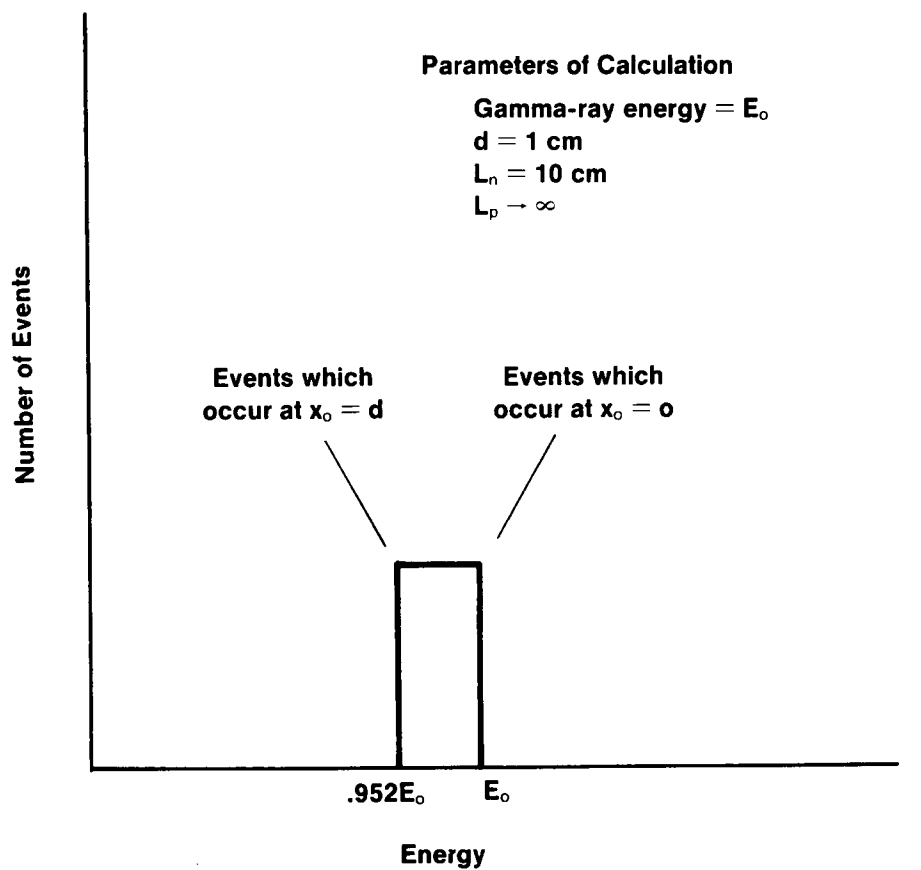
For this condition, if all non-ideal effects except charge trapping are neglected, the measured energy spectrum would appear as shown in figure (2-10). This figure illustrates one way charge trapping can degrade energy resolution.

As noted by Trammell and Walter [27] and discussed in detail by Raudorf and Pehl [20], fluctuations in the number of carriers trapped for events at a single interaction point will also degrade the energy resolution. Consider the case of a collimated radiation source where all the gamma-rays interact at position x_0 in the detector, producing N_0 electron-hole pairs. The mean number of carriers trapped per event is $d_n(x)N_0 + d_p(x)N_0$. However, due to the random nature of charge trapping, the number of carriers trapped will be statistically distributed.

Raudorf and Pehl, in a manner analogous to the derivation of charge production statistics, assumed

$$\sigma_T^2 = KN_T \quad (2-19)$$

where σ_T is the standard deviation of the number of trapped carriers, N_T is the average number of trapped carriers and K is a constant. From the experimental work of McMath and Martini [14], Raudorf and Pehl calculated a value of 340 for K . Thus, equation (2-19) can be rewritten as [20]



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Figure (2-10). Energy spectrum broadened only by charge trapping deficit.

$$\sigma_T(r) = \{340\epsilon E_0[1-n(x)]\}^{0.5} \quad (2-20)$$

where E_0 is the energy of the incident gamma-ray, ϵ is the average energy required to produce an electron-hole pair, and the units of equation (2-20) are eV.

The third way in which charge trapping may degrade resolution is by causing ballistic deficit. Up to this point, the equations derived for charge trapping assumed that captured carriers were not released during the pulse processing time. This type of trap is commonly referred to as a deep-level trap [24]. However, the detrapping times will be continuously distributed over a range from 0 to effectively infinity. Thus, some of the trapped charge carriers will be released quickly enough to contribute to the output current of the detector. In this case, the charge is delayed, not lost, and degrades the resolution by producing a ballistic deficit. This type of charge trapping is often referred to as shallow-level trapping [24].

2.6 Trammell-Walter Equation

Trammell and Walter also considered the problem of calculating the shape of the energy line obtained from a detector with a significant amount of charge trapping. They considered a slice of the detector in figure (2-9) between x and $x + dx$. Assuming no y dependence of the charge trap distribution, the energy spectrum obtained from a mono-energetic source interacting in this infinitesimal volume is a Gaussian centered around $n(x)E_0$. The shape of the energy line is found by summing the contributions of each of these volumes. Which, in the limit as dx approaches 0, becomes [27]

$$\frac{dN(E)}{dE} = \frac{1}{(2\pi)^{0.5}} \int_0^d \frac{1}{\sigma(x)} G(x) \times \exp\left\{-\frac{(E-n(x)E_0)^2}{2\sigma(x)^2}\right\} dx, \quad (2-21)$$

where E_0 is the energy of the incident gamma-ray, $\sigma(x)$ is the standard deviation of the normal distribution, $dN(E)$ is the fraction of events occurring in the energy range E to $E + dE$. $dN(E)/dE$ is experimentally proportional to the number of counts in a MCA channel [20]. $G(x)$ is a geometrical weighting function that gives the probability of the interaction occurring at x . For the planar geometry considered here, $G(x)$ is uniform over the entire detector

(i.e. under uniform irradiation the interaction is equally likely to occur at any x in the detector). For other geometries, this weighting function will not be uniform.

Although derived for the case of a planar detector, equation (2-21) is valid for other geometries if the proper functions for G and n are used. In particular, equation (2-21) rewritten for the case of coaxial geometry gives [20,26]

$$\begin{aligned} dN(E)/dE = & 1/(2\pi)^{0.5} \int_{r_1}^{r_2} G(r)/\sigma(r) \\ & \times \exp\{-(E-n(r)E_0)^2/(2\sigma(r)^2)\} dr \end{aligned} \quad (2-22)$$

where r is the radius of interaction, r_2 is the outer radius of the detector, and r_1 is the inner radius of the detector.

The standard deviation for the normal distribution in the integrand of equation (2-22), $\sigma(r)$, is given by

$$\sigma(r) = (\sigma_F^2 + \sigma_n^2 + \sigma_T^2)^{0.5}, \quad (2-23)$$

where σ_F is the broadening due to charge production statistics, σ_n is the contribution due to electronic noise, and σ_T , the broadening due to trapping fluctuations, is given in equation (2-20). Since this last term is dependent on r , $\sigma(r)$ must remain inside the integral in equation (2-22).

Equation (2-22) is written for the case of Φ and Z independence of the charge trap distribution. A Φ and Z dependence could be included by converting equation (2-22) into a volume integral. However, in practical calculations, a trap distribution that varies with only r is usually assumed. For example, a charge distribution which is larger on the outer periphery than on the inner radius of the detector can be handled via equation (2-22). However, a distribution which increases as the distance into the detector increases would require an integration with respect to Z also.

Equation (2-22) accounts for electronic noise, charge generation fluctuations, charge trapping fluctuations, and charge trapping losses. However, ballistic deficit effects are not included in this expression. Ballistic deficit effects can be included by introducing into equation (2-22) a term, g , to model the ballistic efficiency. The ballistic efficiency accounts for ballistic deficit by allowing E_0 to apparently depend upon radius, r , and the peaking time, t_p , of the spectroscopy amplifier. Thus, equation (2-22) becomes

$$\begin{aligned} dN(E)/dE = & 1/(2\pi)^{0.5} \int_{r_i}^{r_f} G(r)/\sigma(r) \\ & \times \exp\{-(E-g(r,t_p)n(r)E_0)^2/(2\sigma(r)^2)\} dr \end{aligned}$$

(2-24)

The following method can be used to determine g . As described in section (2-4), the ballistic deficit can be calculated using PSpice and piecewise linear approximations to the preamplifier output signals. Then,

$$g(r) = 1 - \text{RBD}, \quad (2-25)$$

where RBD is the relative ballistic deficit defined in equation (1-10). The t_p^2 dependence of the ballistic deficit predicted by equation (2-11) can be included by modifying equation (2-25) to become

$$g(r, t_p) = 1 - (K_g)(\text{RBD})/t_p^2, \quad (2-26)$$

where K_g is a normalizing constant which makes equation (2-25) and (2-26) equal for the t_p used in the calculation of (2-25). Comparisons between measured and calculated results will be shown in Chapter 5.

CHAPTER 3

BALLISTIC DEFICIT CORRECTION

In the previous two chapters, the cause and effect of ballistic deficit has been documented. Figure (1-7) illustrates how ballistic deficit comes about, while figure (1-8) shows the resulting effect on the energy resolution. In Chapter 2, a more rigorous mathematical development of the theory of ballistic deficit is given. The relationships derived by Loo, Goulding and Gao [13] are shown in equations (2-10) and (2-11). Also, the results from a more accurate computer analysis are shown in figure (2-8). Finally, equation (2-24) presents the generalized energy distribution function expression which includes the effects of ballistic deficit along with charge generation fluctuations, electronic noise, and charge trapping effects.

This chapter will deal with correcting for ballistic deficit. The gated integrator, which was first developed by Radeka [19], will be briefly described. Then, the ballistic deficit correction (BDC) circuit reported by Goulding and Landis [7] will be presented in depth. This treatment will include an extensive analysis of non-ideal effects of the BDC circuit. Also, the ballistic deficit correction circuit developed by Hinshaw and Landis [10] will be described.

The last section of this chapter will present a comparison of the three methods of ballistic deficit

correction treated here. Other methods to correct for ballistic deficit, such as the circuit designed by White [29] are not widely used in nuclear spectroscopy and will not be considered here.

3.1 Gated Integrator

Since 1970 when Radeka [19] first described the gated integrator, it has been the instrument of choice for high-rate, high-resolution nuclear spectroscopy. Figure (1-11) shows a block diagram of a gated integrator, and figure (1-10) shows the impulse response obtained from a gated integrator.

As described in Chapter 1, the gated integrator is insensitive to ballistic deficit due to the flat-top of the impulse response. Figure (1-12) depicts the convolution of a current pulse with the impulse response of the gated integrator, graphically showing the insensitivity to ballistic deficit.

Since the gated integrator automatically and completely corrects for ballistic deficit, the motivation to develop other instruments to perform this function is not immediately obvious. However, as noted by Goulding and Landis [7] and in Chapter 1, a gated integrator with a Gaussian or triangular prefilter has poor low-frequency

performance. This is illustrated in figure (3-1), where a triangular pulse with amplitude A and width T is the input to a gated integrator. This input signal is corrupted by a low-frequency noise component with RMS value of B . The input signal-to-noise ratio (SNR) is defined here as A/B .

The response of the gated integrator to the triangular input signal is simply the area of the triangle, or

$$V_0 = AT/2, \quad (3-1)$$

while the standard deviation of V_0 due to the input noise is given by [15]

$$\sigma_{V_0} = [2 \int_0^T h(v) \{ \int_0^v h(u) \Phi(u-v) du \} dv]^{0.5}, \quad (3-2)$$

where Φ is the auto-correlation function of the input noise, and h is the impulse response of the integrator, which is the unit step function. Φ can be approximated as a constant for the time duration considered in the integral of equation (3-2) since low-frequency noise was assumed. Thus, equation (3-2) reduces to

$$N_0 = BT. \quad (3-3)$$

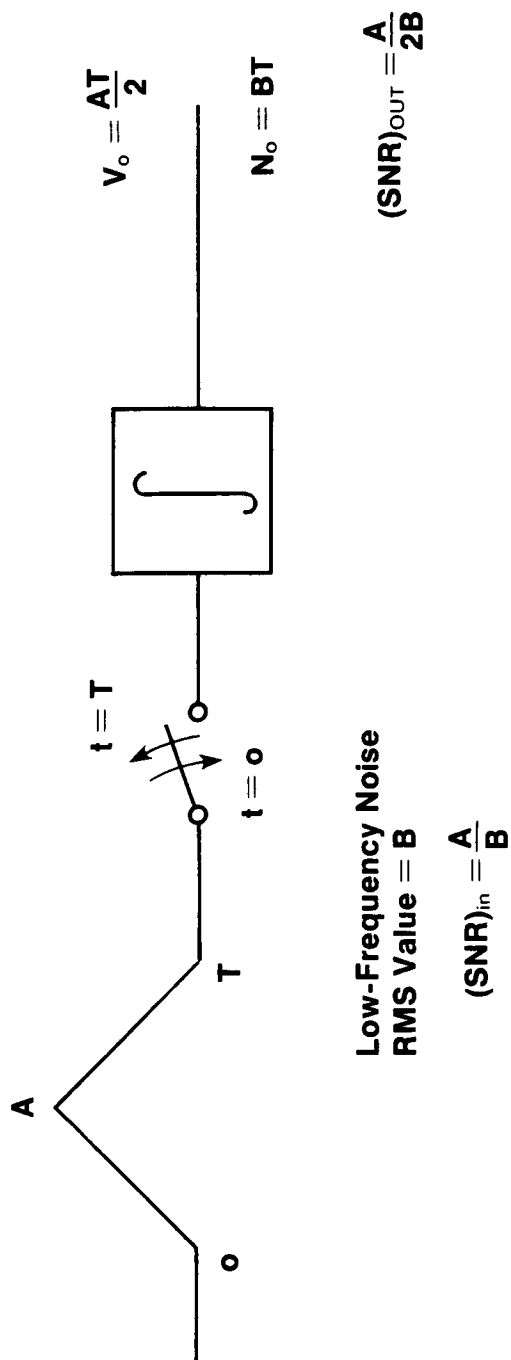


Figure (3-1). Low-frequency noise analysis of gated integrator.

The output SNR then is

$$(\text{SNR})_{\text{out}} = (\text{SNR})_{\text{in}} / (2). \quad (3-4)$$

As equation (3-4) shows, any low-frequency noise at the output of the prefilter will seriously degrade the SNR at the output of the gated integrator. This includes dc baseline fluctuations that may be caused by incomplete pole-zero cancellation or non-ideal baseline restoration of the prefilter signal at high input rates. Thus, an instrument that could correct for ballistic deficit while retaining the noise performance of the semi-Gaussian filter would be advantageous.

3.2 Goulding and Landis BDC Circuit

Citing the complexity and low-frequency noise performance of the gated integrator as motivation, Goulding and Landis [7] developed a ballistic deficit correction circuit which, to first order, retained the noise performance of the semi-Gaussian filter. This circuit was briefly discussed in Chapter 1, and is shown in block diagram form in figure (1-13).

This circuit is based on equation (2-11) which shows that the relative ballistic deficit is proportional to the

square of the charge collection time, T . Or, rewriting equation (2-11),

$$BD = GS_0(T/t_p)^2, \quad (3-5)$$

where BD is the ballistic deficit, G is a constant, t_p is the peaking time of the spectroscopy amplifier and S_0 is the amplitude of the pulse from the spectroscopy amplifier excluding the effect of ballistic deficit. In the Goulding and Landis scheme, the amplitude of the spectroscopy amplifier output and the charge collection time are measured and the ballistic deficit is calculated. A correction signal, given by equation (3-5) is then added to the output signal to compensate for the ballistic deficit.

In the Goulding and Landis circuit, the charge collection time is not measured directly. Rather, the unipolar peak delay time (UPDT) is found. However, as shown in equation (1-8), these two quantities are linearly related if a rectangular shape (see figure (1-6)) is assumed for the current pulse from the detector. Thus, a necessary assumption for the ideal operation of the Goulding and Landis BDC circuit is that current pulses from the detector have a shape like that shown in figure (1-6) and only vary in amplitude and duration [7]. The validity of this assumption will be considered in section (3.2.1). Using this assumption, Goulding and Landis [7] showed that

equation (3-5) could also be written as

$$BD = S_0(t_d/t_p)^2. \quad (3-6)$$

A second assumption is imposed on this circuit by equation (2-11). Semi-Gaussian shaping was assumed in deriving this equation, which would seem to impose this constraint on the Goulding and Landis BDC circuit. However, Loo, Goulding, and Gao [13] showed that equation (2-11) can be applied to any pulse shape if the region near the top of the pulse is approximately parabolic. Fortunately, nearly all pulse shapes used in nuclear spectroscopy fall into this category.

In the ideal case, then, the Goulding and Landis circuit operates in the following manner (refer to figure (1-13)):

1. The signal from the charge sensitive preamplifier is differentiated and then directed to the spectroscopy amplifier and a fast discriminator.
2. The fast discriminator generates a logic signal that indicates that a valid signal has been received by the spectroscopy amplifier. This signal initiates a pulse processing cycle.

3. The discriminator output is delayed by an amount of time equal to the peaking time of the impulse response of the spectroscopy amplifier. This delayed signal sets a flip-flop to the high state.
4. The output of the spectroscopy amplifier is differentiated, producing a bipolar signal whose zero crossing is indicative of the peaking time of the spectroscopy amplifier's output. A logic signal generated by this zero crossing resets the flip-flop in step 3 to the low state. The output of this flip-flop then, is a pulse whose duration is equal to the UPDT. This signal is one input to an analog computer.
5. The other input to the analog computer is the spectroscopy amplifier output. The output of the analog computer is a constant times the square of the UPDT times the amplitude of the spectroscopy amplifier output.
6. The properly scaled output of the analog computer is summed with the peak stretched spectroscopy amplifier output. After this summed signal has settled, it is gated out to the ADC.

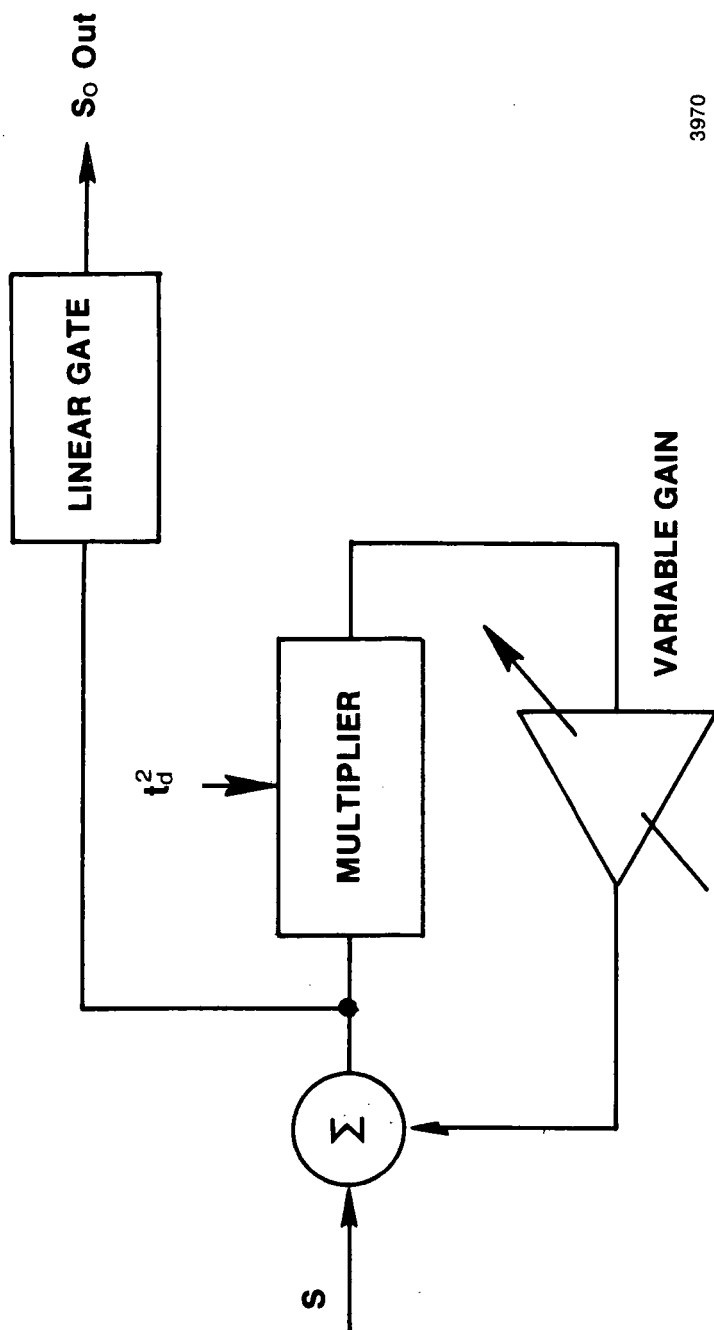
3.2.1 Non-Ideal Behavior of the Goulding and Landis BDC Circuit

Goulding and Landis did not consider in what ways their BDC circuit deviated from the ideal. In this section, three non-ideal properties of the Goulding and Landis BDC circuit are described.

One non-ideal property of the Goulding and Landis BDC circuit can be understood by considering equation (3-6). This equation shows the ballistic deficit to be proportional to the amplitude of the signal from the spectroscopy amplifier excluding the effect of ballistic deficit. However, the calculation of the correction is performed using the signal from the spectroscopy amplifier which includes ballistic deficit. Thus, the correction signal is given by

$$\begin{aligned} V_c &= (S_0 - S_0(t_d/t_p)^2)(t_d/t_p)^2 \\ &= S_0(t_d/t_p)^2 - {}^2S_0(t_d/t_p)^4. \end{aligned} \quad (3-7)$$

The first term in equation (3-7) is the desired correction, while the second term is a deviation from the ideal correction. This error is negligible when long peaking times are used where $t_p \gg t_d$. However, Simpson, et al. [24], showed that errors of 1.6 keV or more are possible with short peaking times. Figure (3-2) shows a circuit



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Figure (3-2). Circuit to compensate for incomplete ballistic deficit correction.

proposed by Simpson, et. al. [24], to compensate for the error term in equation (3-7). This circuit sums the correction with the peak stretched signal and then recalculates the correction continuously.

As mentioned earlier, equation (3-6) assumes that the current signals from the detector have a rectangular shape as shown in figure (1-6). This is equivalent to assuming that the signals from the preamplifier exhibit a linearly rising voltage pulse. However, as depicted in figure (2-6), actual preamplifier signals deviate greatly from this linear approximation. The effect of preamplifier signals which exhibit a non-linear rise was calculated in the following manner:

1. The preamplifier signals of figure (2-6) were represented as piecewise linear functions and used as inputs to a PSPICE model of a semi-Gaussian amplifier.
2. PSPICE was used to calculate the ballistic deficit and the UPDT.
3. Using the UPDT from step 2, the ballistic deficit was calculated with equation (3-6).

4. An error term was formed by subtracting the ballistic deficit calculated in step 3 from that calculated in step 2.

The steps above were performed for two different spectroscopy amplifier shaping times and the errors are plotted in figure (3-3) versus the normalized radius of interaction. As this figure shows, the error is much greater at the shorter peaking time.

The poor low-frequency noise performance exhibited by the gated integrator is not a problem with the Goulding and Landis BDC circuit. However, this circuit requires the accurate measurement of a pulse amplitude and a delay time to calculate the correction for ballistic deficit. Statistical variations in these two parameters will generate a statistical variation in the correction signal, which will broaden the energy peak. A derivation of this correction noise is given below.

From equation (3.6), the correction voltage is

$$V_c = (1/t_p)^2 S t_d^2. \quad (3-8)$$

The mean and mean squared values of V_c are given by

$$\overline{V_c} = (1/t_p)^2 \overline{S t_d^2} \quad (3-9a)$$

$$\overline{V_c^2} = (1/t_p)^4 \overline{S^2 t_d^4}, \quad (3-9b)$$

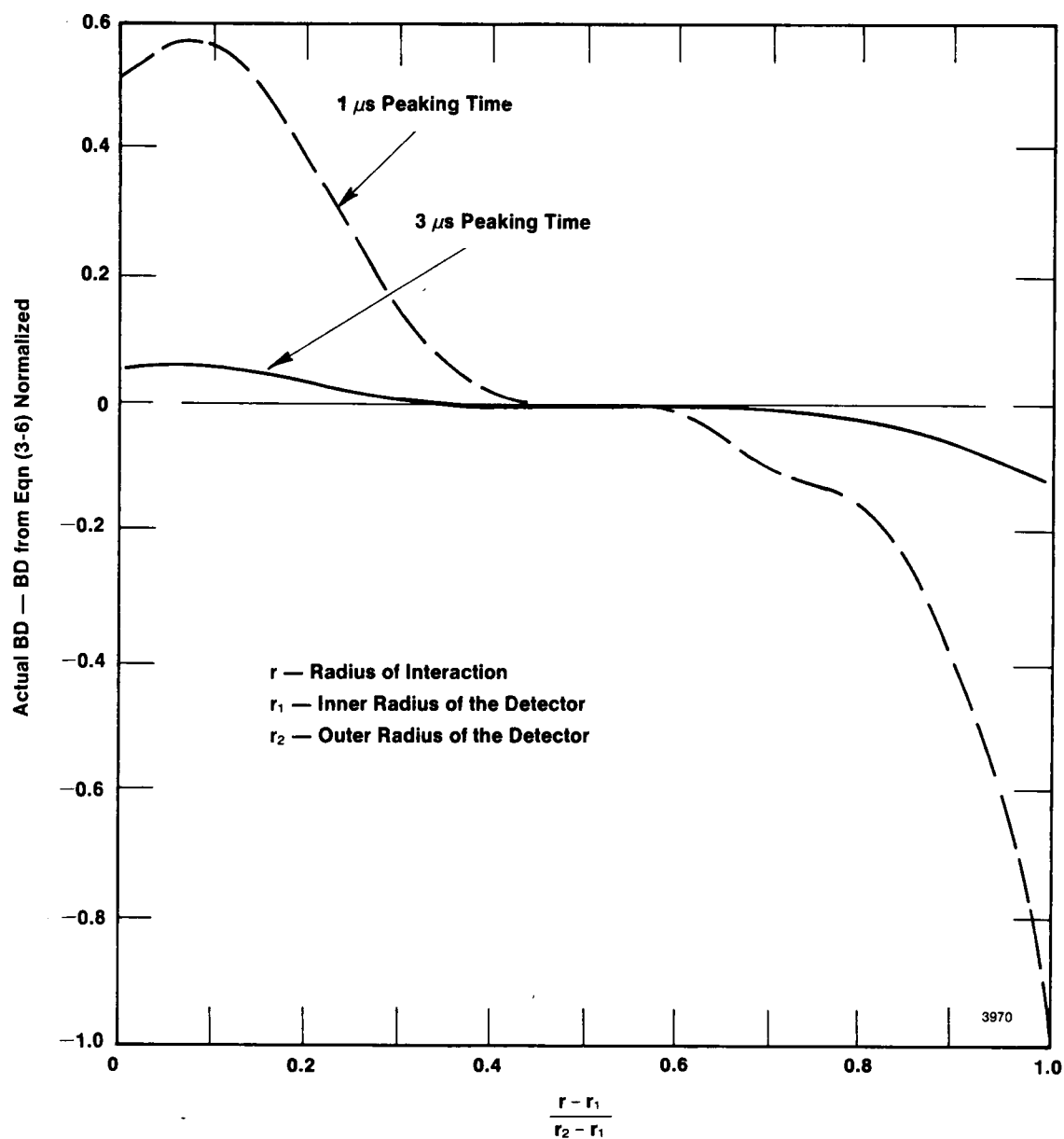


Figure (3-3). Error in ballistic deficit correction due to preamplifier pulse shapes.

where S and t_d are assumed to be statistically independent.

Since

$$\overline{S^2} = (\overline{S})^2 + \sigma_s^2, \quad (3-10)$$

where σ_s is the standard deviation of S , equation (3-9b) can be rewritten as

$$\overline{V_c^2} = (1/t_p)^4 [(\overline{S})^2 + (\sigma_s)^2] \overline{t_d^4}. \quad (3-11)$$

However, over the energy range where ballistic deficit is important, $S \gg \sigma_s$ in a high resolution system such as a Ge spectrometer. Thus, σ_s^2 can be dropped from equation (3-10). Then σ_{vc} , the standard deviation of the correction signal, is given by

$$\begin{aligned} \sigma_{vc} &= (\overline{V_c^2} - (\overline{V_c})^2)^{1/2} \\ &= (1/t_p)^2 \overline{S} [\overline{t_d^4} - (\overline{t_d^2})^2]^{1/2} \\ &= (1/t_p)^2 \overline{S} [\sigma(t_d^2)], \end{aligned} \quad (3-12)$$

where $\sigma(t_d^2)$ is the standard deviation of t_d^2 . Appendix A shows that

$$\sigma(t_d^2)^2 \sim 2\overline{t_d} \sigma(t_d), \quad (3-13)$$

where $\sigma(t_d)$ is the standard deviation of t_d . Thus

$$\sigma_{vc} \sim 2(1/t_p)^2 \overline{S} t_d \sigma(t_d). \quad (3-14)$$

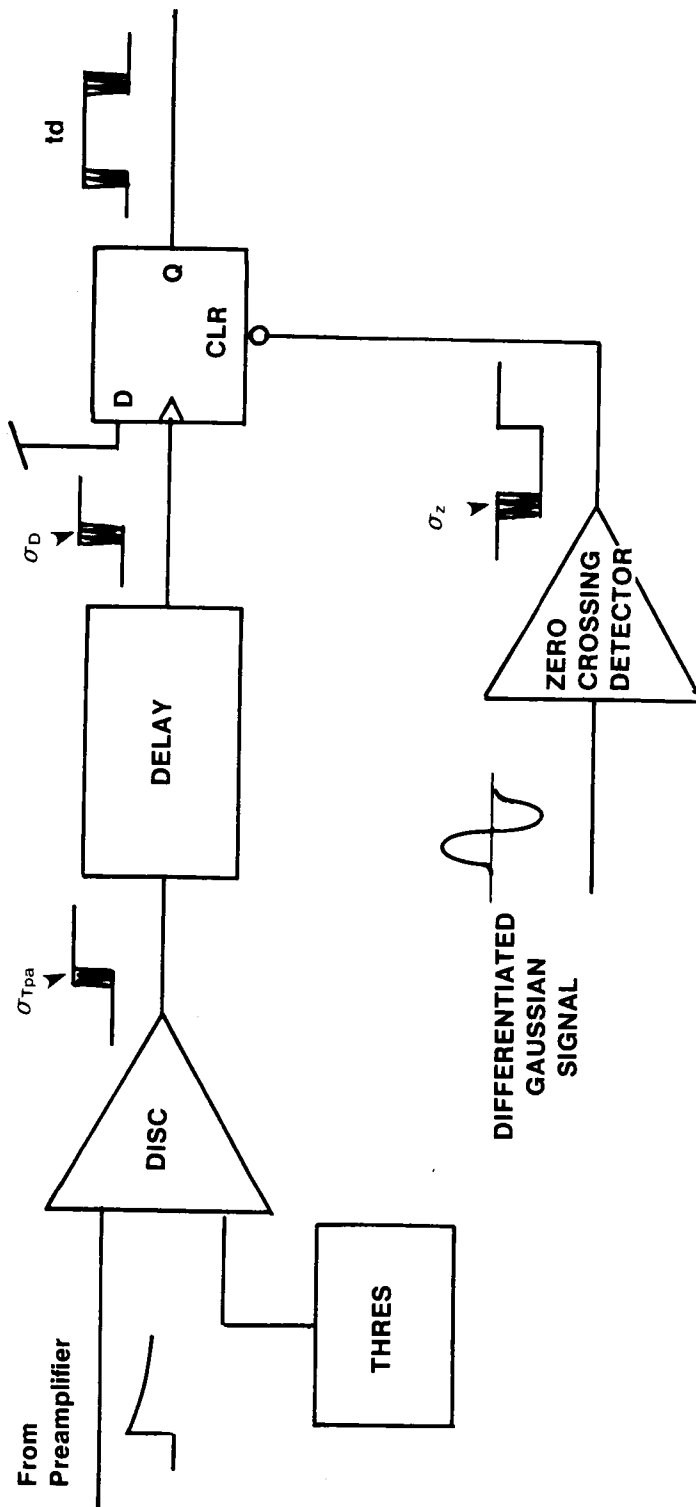
Equation (3-14) shows that the noise created by the Goulding and Landis BDC circuit is mainly due to the uncertainty or jitter in the UPDT measurement. Figure (3-4) shows a block diagram of the UPDT measurement circuit and identifies the three components of the measurement jitter. The first of these components is the jitter in the discriminator output which initiates the pulse processing cycle. Using the well known triangle rule [16],

$$\sigma_{Tpa} = \sigma_{vpa} / M_{pa}, \quad (3-15)$$

where σ_{Tpa} is the RMS timing jitter in the discriminator output, σ_{vpa} is the RMS voltage noise at the input to the discriminator, and M_{pa} is the slope of the signal at the input to the discriminator. Once again, it will be convenient to assume that the preamplifier signals exhibit a linearly rising voltage pulse. In this case,

$$M_{pa} = K_p S_o / T, \quad (3-16)$$

where K_p is the conversion gain of the charge sensitive preamplifier, and S_o and T are as defined earlier.



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Figure (3-4). Block diagram of UPDT measurement circuit.

Substituting equation (3-16) into (3-15),

$$\sigma_{Tpa} = (\sigma_{vpa})(T)/(K_p S_o). \quad (3-17)$$

The second component of timing uncertainty in the UPDT measurement is generated by the fixed delay between the output of the discriminator and the setting of the UPDT flip-flop. This delay is nominally equal to the peaking time of the spectroscopy amplifier, and is typically greater than 1 μ S. Thus, this delay is usually produced by an active device rather than a passive delay line. Delays are normally produced in active devices by comparing a ramp signal to a fixed threshold. The length of the delay is controlled by the slope of the voltage ramp. Again, applying the triangle rule,

$$\sigma_D = \sigma_v t_p / V_r, \quad (3-18)$$

where σ_D is the RMS timing jitter in the delay, V_r is the amplitude of the voltage ramp, σ_v is the rms input noise of the comparator circuit in the active delay and t_p is the peaking time of the spectroscopy amplifier. The rms noise added due to this timing jitter is found by substituting equation (3-18) into equation (3-14) which gives

$$\sigma_{vc}(\text{delay}) = 2\sigma_v S t_d / (t_p V_r). \quad (3-19)$$

As a typical example, the following calculation is presented. The highest energy of interest in a gamma-ray spectrometer would generally be $\leq 20\text{MeV}$, and a large detector could have a $t_d = 500\text{nS}$. If $V_r = 2\text{V}$ and $\sigma_v = 100\mu\text{V}$ ($10\text{nV}/(\text{Hz})^{1/2}$ in a 100MHz bandwidth), then

$$\sigma_{vc}(\text{delay}) = .001/t_p \text{ (eV)}. \quad (3-20)$$

The peaking time of commercially available quasi-Gaussian shaping amplifiers is $\geq .5\mu\text{S}$. Thus, the largest $\sigma_{vc}(\text{delay})$ would be 2KeV at the 20MeV energy line, which is negligible compared to charge generation fluctuations. Thus, this component of timing jitter will be neglected in the remainder of this development.

The final component of UPDT measurement uncertainty is generated by the jitter in the zero-crossing-detector which determines the actual peaking time of the spectroscopy amplifier output by detecting the zero-crossing of the differentiated spectroscopy amplifier signal (bipolar signal). Once again the triangle rule can be applied to give

$$\sigma_z = \sigma_{bp}/M_{bp}, \quad (3-21)$$

where σ_z is the standard deviation in the time of zero-crossing detection, σ_{bp} is the rms voltage noise of the

bipolar signal, and M_{bp} is the slope of the bipolar signal at the zero-crossing.

It is reasonable to assume that the slope of the bipolar signal at the zero-crossing is linearly related to the amplitude of the bipolar signal. It can also be shown analytically, by measurement, or by computer simulation that

$$M_{bp} \sim K_{bp} S / t_p, \quad (3-22)$$

where K_{bp} is a constant determined by the type of shaping amplifier used and the quality of the differentiation circuit employed to form the bipolar signal. Substituting equation (3-22) into (3-21) gives

$$\sigma_z = \sigma_{bp} t_p / (S K_{bp}). \quad (3-23)$$

The total rms timing jitter is given by the quadrature sum of equations (3-23) and (3-17) which yields

$$\sigma_{Td} = 1/S_0 (\sigma_{bp}^2 t_p^2 / K_{bp}^2 + \sigma_{vpa}^2 T^2 / K_p^2)^{1/2}, \quad (3-24)$$

where it was assumed that $S \sim S_0$. Substituting equation (3-23) into equation (3-14) shows

$$\sigma_{vc} = 2(1/t_p)^2 \overline{t_d} (\sigma_{bp}^2 t_p^2 / K_{bp}^2 + \sigma_{vpa}^2 T^2 / K_p^2)^{1/2}. \quad (3-25)$$

Equation (3-25) shows the correction noise to be worse at shorter peaking times, which would be expected since the magnitude of the correction would be largest at the smallest peaking times. Also, the correction noise is independent of energy, and thus is more significant at the lower energies where charge generation fluctuations are less important. This noise source is proportional to t_d and therefore T , making the magnitude of the noise term greater for larger detectors. Equation (3-25) gives the correction noise for a particular t_d and T . Therefore, to make noise calculations it is necessary to perform a weighted average of σ_{vc} over the entire range of t_d and T . However, t_d and T are distributed over a range of only ~2:1 [21], and an average value for these parameters can be used in equation (3-25) without large errors. Figure (3-5) shows calculated and measured σ_{vc} versus t_p for a typical detector.

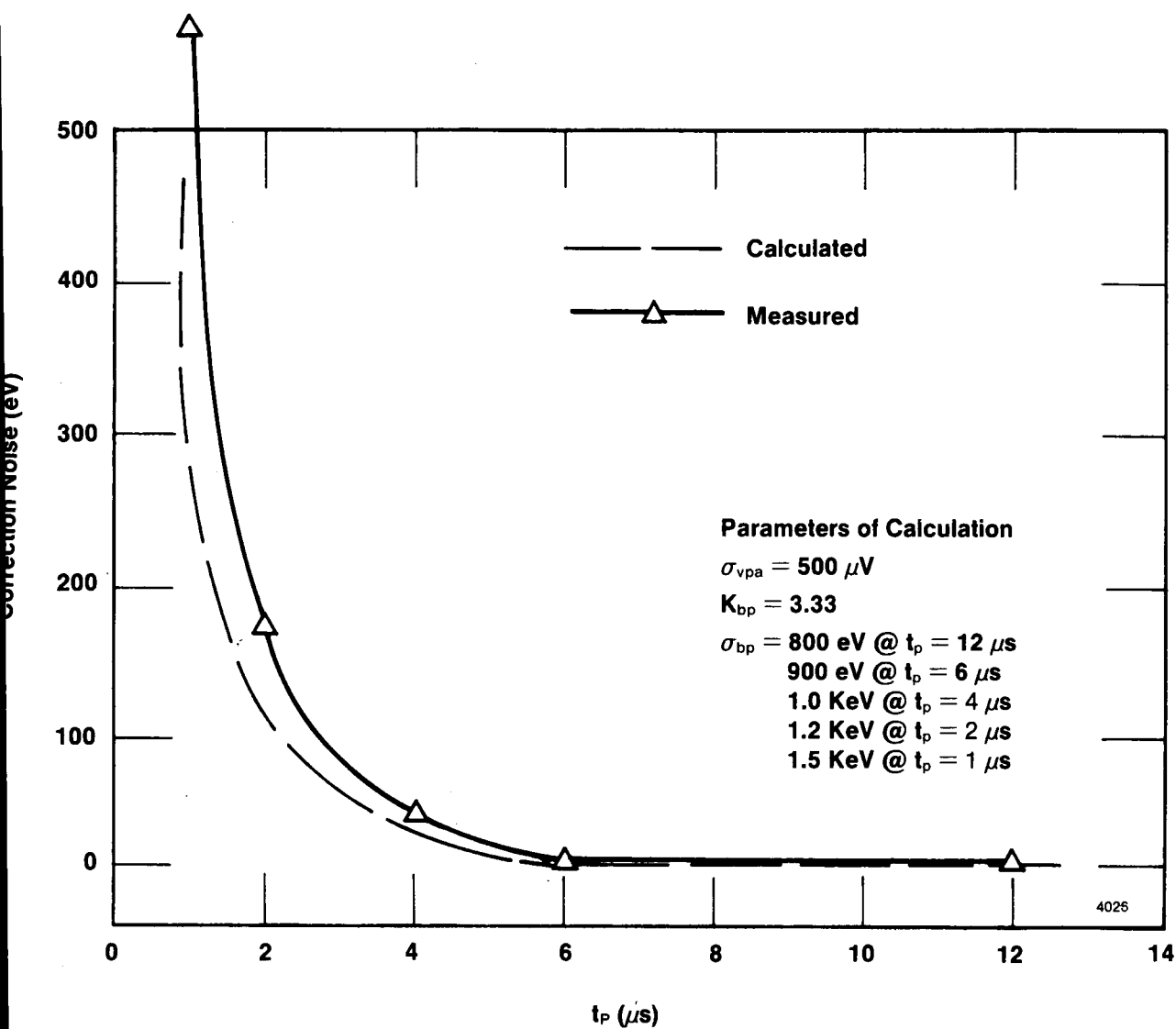


Figure (3-5). Measured and calculated correction noise versus t_p .

3.3 Hinshaw and Landis BDC Circuit

As stated earlier, equation (3-6) was derived assuming linearly rising voltage pulses from the preamplifier. However, as shown in section 3.2.1, this assumption leads to some error in the Goulding and Landis BDC circuit. If the preamplifier pulse shape restriction is removed, equation (3-6) becomes

$$BD = S_0 f(t_d) / t_p^2, \quad (3-26)$$

where f is some arbitrary, single-valued function of t_d only.

Consider the case of two shaping amplifier pulses formed from the same preamplifier signal with peaking times t_{p1} and t_{p2} where $t_{p2} > t_{p1}$. The difference in the peak values of the two spectroscopy amplifier signals would be

$$\begin{aligned} S_2 - S_1 &= BD_1 - BD_2 \\ &= S_0 f(t_d) [(t_{p2}^2 - t_{p1}^2) / (t_{p1} t_{p2})^2]. \end{aligned} \quad (3-27)$$

Equation (3-27) can be scaled by a constant to yield

$$C(S_2 - S_1) = S_0 f(t_d) / t_{p2}^2, \quad (3-28)$$

where

$$C = t_{p2}^2 / (t_{p2}^2 - t_{p1}^2). \quad (3-29)$$

Equation (3-28) gives the necessary ballistic deficit correction signal for the shaping amplifier with the longer peaking time and is the basis of the Hinshaw and Landis BDC circuit [10]. This circuit was briefly discussed in Chapter 1 and is shown in block diagram form in figure (1-14). In the Hinshaw and Landis implementation, the signal with the shorter peaking time is produced by differentiating the signal with the longer peaking time.

The Hinshaw and Landis BDC circuit has two major advantages over the Goulding and Landis BDC circuit. First, since no time measurement is required, there is no correction noise due to timing jitter. Second, the Hinshaw method places less stringent restrictions on the shape of the preamplifier pulse.

Hinshaw and Landis [10] did report, however, that their BDC circuit began to lose accuracy when $T/t_p > 0.2$. This inaccuracy is due to the assumption in equation (3-26) that f is a function of t_d only. When T (and therefore t_d)

becomes a significant fraction of t_p , this assumption is no longer valid. Thus, like the Goulding and Landis circuit, this instrument begins to lose accuracy as the peaking time of the spectroscopy amplifier becomes smaller. Goulding, Landis, and Hinshaw [8] report, however, that the Hinshaw and Landis circuit more completely corrects for ballistic deficit than the Goulding and Landis circuit when short shaping times are employed.

3.4 Comparison of Ballistic Deficit Correction Methods

The three methods of ballistic deficit correction described in this chapter were designed to allow for good energy resolution at short shaping times. However, these methods differ greatly in their implementation.

The gated integrator eliminates the ballistic deficit by virtue of its flat-topped impulse response. The correction is performed automatically (no adjustments are needed), and the correction is complete as long as the flat-top portion of the impulse response has a greater duration than the widest current pulse from the detector. The gated integrator does not have the noise performance of a Gaussian or triangular shaping amplifier, and is particularly sensitive to low-frequency noise components in the prefilter.

The Goulding and Landis BDC circuit was designed to avoid the disadvantages of the gated integrator. This circuit retains much of the noise performance of the Gaussian or triangular main shaping amplifier, but reduces the effects of ballistic deficit by adding a correction signal to each pulse. Over a certain range of shaping times, the Goulding and Landis BDC circuit will provide better resolution than the gated integrator due to its superior noise performance. However, at shorter shaping times the performance of the BDC circuit will degrade due to timing jitter noise and non-ideal qualities of the correction computation as detailed in section 3.2.1.

The Hinshaw and Landis BDC circuit retains the noise performance of the Gaussian or triangular main shaping amplifier yet does not add timing jitter noise to the spectrometer. However, at shaping times when T/t_p exceeds 0.2, the Hinshaw and Landis BDC method begins to deviate significantly from the ideal. Thus, at extremely high input counting rates when the width of the pulse from the shaping amplifier becomes very short, the gated integrator probably will provide for the best resolution.

CHAPTER 4

CHARGE TRAPPING CORRECTION CIRCUITS

The theory of charge trapping was considered in Chapters 1 and 2. In Chapter 1, the effect of charge trapping on the energy resolution of an HPGe detector was described in an intuitive manner as being similar to the effect of ballistic deficit. Equations (1-16) through (1-20) described the charge trapping deficit in a simplified form. In Chapter 2, however, a more rigorous approach to charge trapping was taken in Section 2.6, where the Trammell-Walter equation was presented in equation (2-22) and in modified form in equation (2-24).

This chapter will deal with circuits which compensate for charge trapping energy deficits. In particular, the class of circuits considered by Simpson, et. al. [23,25], will be described and analyzed in section 4.2 after the theory of charge trapping has been further described in section 4.1. Section 4.3 will consider the Hinshaw/Landis circuit's ability to correct for charge trapping, while section 4.4 will describe a circuit which has a gated integrator to correct for ballistic deficit, but also uses a charge trapping correction circuit.

4.1 Charge Trapping in Coaxial HPGe Detectors

In general, both majority and minority carriers are trapped in HPGe detectors. However, as shown by Raudorf and Pehl [20], with coaxial detectors a much greater concentration of minority charge traps is required to have the same effect on the energy resolution as majority carrier traps. The concentration of minority carrier traps required to significantly degrade the performance of a coaxial HPGe detector is not generally seen in modern commercial devices. However, hole traps can be induced in Ge detectors by radiation damage [20], leading to significant amounts of minority carrier trapping in reverse electrode detectors. Thus, the two cases of majority carrier dominated and minority carrier dominated charge trapping will be considered.

4.1.1 Majority Carrier Trapping

The reverse electrode detector shown in figure (4-1) is used in this discussion. However, the same arguments are applicable to the conventional electrode case if the proper voltage and charge polarities are used.

As in Chapter 2, the electric field, E , in figure (4-1) is assumed large enough that free carriers produced in the

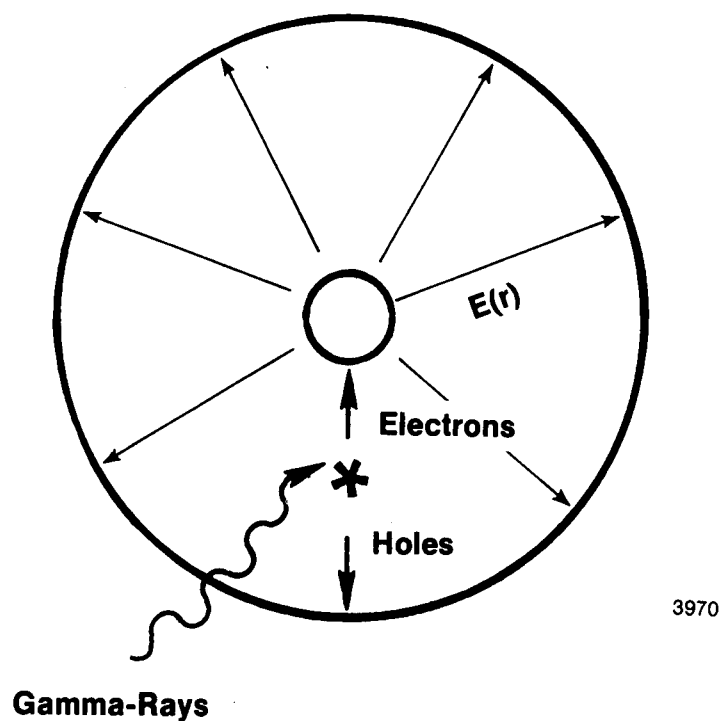


Figure (4-1). Cross section of coaxial, germanium, radiation detector.

detector reach saturation velocity, V_{sat} , very quickly. Furthermore, the saturation velocity for electrons and holes are assumed equal. Then,

$$t_{cmj} = (r - r_1)/V_{sat}, \quad (4-1)$$

where t_{cmj} is the majority carrier charge collection time, r is the radius of the gamma-ray interaction, and r_1 is the inner radius of the detector. Rearranging equation (4-1) gives

$$r = r_1 + V_{sat}t_{cmj}. \quad (4-2)$$

Assuming that charge trapping is dominated by majority carrier traps, the fractional loss of pulse height due to trapping, d_n , is given by equation (2-15) for planar detectors. Trammell also derived d_n for the coaxial detector case and found [26]

$$d_n = (1/(L_N \ln(r_2/r_1))) [r \ln(r/r_1) - (r - r_1)], \quad (4-3)$$

where the parameters are as described in section 2.5 and L_n assumed to be constant for all values of r . Equation (4-2) can be substituted into equation (4-3), making d_n a function

of t_{cmj} only. Then, to an excellent approximation [23], d_n can be expressed as a simple power law of the form

$$d_n \sim A t_{cmj}^N, \quad (4-4)$$

where A and N are constants determined by r_1 , r_2 , L_n and V_{sat} . For $V_{sat} = 10^7$ cm/sec and $r_1 = 0.5$ cm, $N = 1.61$.

However, in a coaxial detector, L_n is not constant for all values of r , but is inversely proportional to the product of the cross section and the density of the charge traps [23]. The cross section varies inversely with E [23], and equation (2-1) shows that E varies as $1/r$. Thus, L_n is directly proportional to $1/r$.

Since equation (4-3) was derived assuming constant L_n , it is not strictly correct to substitute an L_n with an r dependence into equation (4-3). However, Raudorf and Pehl [20] showed that this could be done with less than a 10% error. For this case, using the parameters given above, $N = 2.38$ if a uniform distribution of traps is assumed throughout the detector [23]. Often, however, the distribution of traps is greater on the periphery of the detector than toward the center [23]. For this case, N may be larger than 2.38. Simpson, et. al. [23], assumed an N between 1.5 and 3.0 would cover most practical cases.

4.1.2 Minority Carrier Trapping

This discussion considers hole trapping in reverse electrode detectors. Again, the same arguments apply to electron trapping in a conventional electrode detector if the proper polarities are used.

In a manner analogous to the majority carrier case,

$$t_{cm} = (r - r_2)/V_{sat}, \quad (4-5)$$

where t_{cm} is the minority carrier charge collection time. Assuming only holes are trapped, the fractional loss of charge due to trapping is given by [26]

$$d_p = 1/(L_p \ln(r_2/r_1))[(r_2 - r) - r \ln(r_2/r)], \quad (4-6)$$

where the parameters are defined in Section 2.5. Again, L_p is assumed directly dependent on $1/r$. Substituting equation (4-5) into equation (4-6) gives a d_p that depends on t_{cm} only, and an equation of the form of equation (4-4) can be assumed for d_p . For this case, $N = 1.48$ if the traps are uniformly distributed.

However, the power fit for the case of minority carrier trapping is not as good as that for majority carrier trapping [23], especially at the largest values of t_{cm} where

d_p is greatest. Thus, any correction scheme based on a power law relationship will not work as well for minority carrier trapping as for majority carrier trapping.

4.2 Charge Trapping Correction

The focus of this section is on the practically important case of majority carrier charge trapping correction, with a brief presentation of minority carrier charge trapping correction. Problems which limit the wide spread use and study of this minority CTC circuit are also discussed.

4.2.1 Majority Carrier Charge Trapping Correction

The majority carrier charge trapping correction circuit is based on equation (4-4). A comparison of this equation with equation (3-6) for ballistic deficit shows the two are similar for the case of $N = 2$ and $t_{cmj} = t_d$. A value of 2 for N is consistent with the derivation given in the preceding section. However, the second assumption involving t_d and t_{cmj} requires further analysis.

The relationship between the majority carrier charge collection time and the UPDT was determined as follows. The

preamplifier output signals were represented as piecewise linear functions that were used as excitation sources in PSpice simulations. These simulations were used to determine t_d values for interactions at various locations in the detector assuming seventh order Gaussian shaping. The t_{cmj} for each of these interaction points was determined by the application of equation (4-1), and t_d was graphed versus t_{cmj} as shown in figure (4-2). As this figure illustrates, a nearly linear relationship exists between t_d and t_{cmj} for $r > 0.4(r_1 + r_2)$. Since the volume of coaxial detectors goes as r^2 , more than 80% of the interactions can be expected to produce a linear relationship between t_d and t_{cmj} . Thus, t_d can be used in place of t_{cmj} in equation (4-4) to implement a majority carrier CTC circuit. Also, with the proper adjustment of correction gain and the selection of $N=2$, this circuit simultaneously corrects for ballistic deficit and deep level, majority carrier, charge trapping.

Figure (1-13) shows a block diagram of a majority carrier charge trapping correction circuit [23]. This circuit allows N to be continuously variable between 0 and 4.

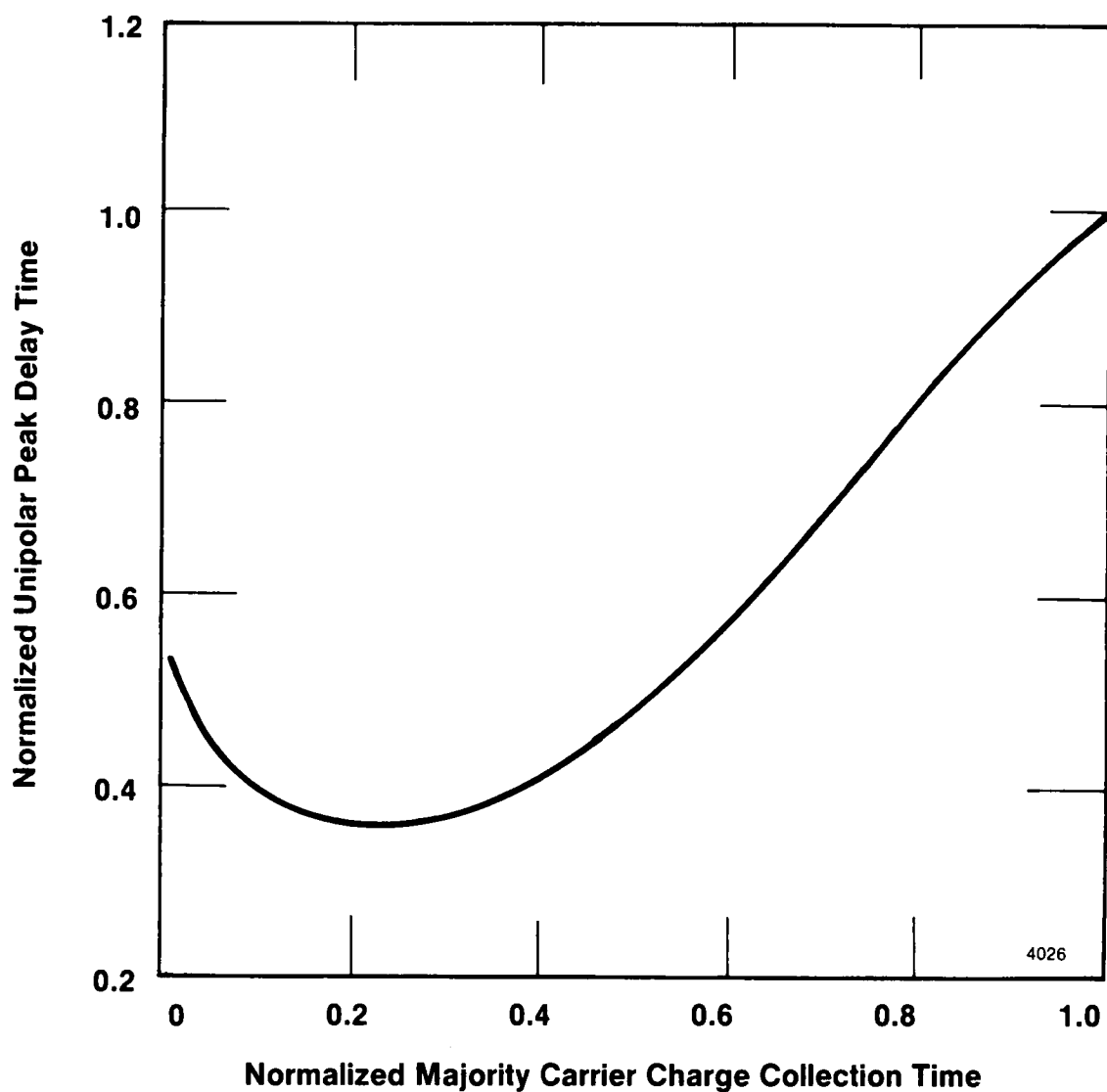


Figure (4-2). Normalized unipolar peak delay time vs. normalized majority carrier charge collection time for single gamma-ray interaction point preamplifier signals.

4.2.1-1 Non-Ideal Properties

Like the ballistic deficit correction circuit discussed in Chapter 3, the majority carrier charge trapping correction (CTC) circuit exhibits some non-ideal properties. In this section, four such properties are described.

First, the use of t_d rather than t_{cmj} results in over-estimation of the correction for 20% of the interactions. In practice, the correction gain can be adjusted to balance the correction and over correction to allow for the best resolution and peak symmetry.

The majority carrier CTC circuit also has the time jitter noise discussed for the case of ballistic deficit correction in Chapter 3. Equation (3-25) written for the case of charge trapping correction ($N=2$) becomes

$$\sigma_{vc} = (A) \overline{t_d} (\sigma_{bp}^2 t_p^2 / K_{bp}^2 + \sigma_{vpa}^2 T^2 / K_p^2)^{0.5}, \quad (4-7)$$

where A is the constant discussed in connection with equation (4-4), and the other parameters are as given in Chapter 3. Except for conventional electrode detectors with severe radiation damage, the magnitude of the charge trapping correction signal is much smaller than the ballistic deficit correction signal when short shaping times

are employed [24]. Thus, the noise in the correction signal is usually negligible for charge trapping correction.

Again, like the Goulding and Landis BDC circuit, the CTC calculation is dependent upon the full expected amplitude of the spectroscopy amplifier signal unperturbed by charge trapping. Including the effects of charge trapping, the correction signal is given by

$$V_c = A(S_0 - AS_0 t_{cmj}^N) t_{cmj}^N, \quad (4-8)$$

which leads to an error in the correction given by

$$V_{err} = A^2 S_0 t_{cmj}^{2N}. \quad (4-9)$$

A circuit like that shown in figure (3-2) could be used to compensate for this error. However, except for severely radiation damaged conventional electrode detectors, this error is quite small and can usually be ignored [24].

The derivations presented in this chapter assume single point interactions rather than the more typical multiple Compton events discussed in Chapter 2. To investigate this effect, consider the case of a gamma-ray which Compton scatters m times before being photo-absorbed in a detector ($m + 1$ total interactions). The charge trapping deficit for this case is

$$CTD = AS_o(J_1 t_{d1}^N + J_2 t_{d2}^N + \dots + J_{m+1} t_{dm+1}^N), \quad (4-10)$$

where J_m is the fraction of the gamma-ray energy lost in the m th interaction and t_{dm} is the UPDT associated with an event occurring at the point of the m th interaction. However, the CTC circuit calculates a correction of only

$$V_c = AS_o t_{deff}^N, \quad (4-11)$$

where t_{deff} is the actual UPDT measured by the CTC circuit. The relationship between t_{deff} and $t_{d1}, t_{d2}, \dots, t_{dm+1}$ is complex and depends upon the spacing between interaction radii and the fraction of energy lost at each interaction point. However, if the fraction of energy deposited at each interaction point is nearly equal, then t_{deff} is the weighted average of $t_{d1}, t_{d2}, \dots, t_{dm+1}$ [24], and equation (4-11) is a good approximation of equation (4-10).

The conclusion to be drawn from the discussion in this section is that the CTC circuit calculates an approximate correction for the effects of majority carrier charge trapping, yet complete correction for this phenomenon is not possible with this circuit. The experimental results presented in Chapter 5, though, will show that dramatic improvements in resolution can be made with this instrument.

4.2.2 Minority Charge Trapping Correction

As shown in Section 4.1.2, a minority carrier charge trapping correction circuit similar to the majority carrier circuit should be possible. However, for this circuit to operate the minority charge collection time must be measured. This measurement could be made in the following manner [23].

Define a parameter, t_x , given by

$$t_x = (r_2 - r_1)/V_{sat}. \quad (4-12)$$

t_x represents the largest charge collection time seen by the detector, since it represents the time required to travel from one electrode across the entire radius of the detector to the other electrode. For a given detector, t_x is a constant since it has no r dependence. Then, the minority charge collection time is given by

$$t_{cm} = t_x - t_{cmj}. \quad (4-13)$$

As shown in Chapter 3, t_{cmj} can be found by measuring the UPDT. Once t_{cm} is found, the correction for minority carrier charge trapping can be performed by the circuit shown in figure (1-13).

Up to this point, very little experimental work has been done with minority carrier charge trapping correction circuits. A small amount of data using a single detector has been reported [23]. This dearth of experimental work can be explained by a lack of need and means. As discussed in Section 4.1, the density of minority carrier traps needed to significantly degrade resolution is much higher than that required with majority carrier trapping. Thus, minority carrier charge trapping is only important for the case of severely radiation damaged, reverse electrode detectors. These detectors can be mostly repaired via a simple annealing process, making a minority carrier CTC circuit of limited practical use.

A radiation damaged, reverse electrode detector is needed to perform experimental work with minority carrier CTC circuits and detector manufacturers are understandably reluctant to damage detectors they hope to sell. Consequently, little experimental work will be done in this area until the need for minority carrier CTC circuits increases greatly.

4.3 Charge Trapping Correction and The Hinshaw/Landis BDC Circuit

Goulding and Landis serendipitously discovered that their BDC circuit could also be used to correct for deep-level, majority carrier charge trapping. This discovery suggests that the charge trapping correction performance of other instruments designed for ballistic deficit correction should also be investigated.

First, the gated integrator only operates on charge which reaches the electrodes and therefore will have no effect on deep-level charge trapping. However, as described in Chapter 2, charge captured by shallow-level traps does reach the electrodes, and the ballistic deficit-like effect caused by this increased charge collection time can be corrected by the gated integrator. For this correction to be complete, though, the total charge collection time must be less than the flat-topped region of the impulse response of the gated integrator (see figure (1-12)).

The Hinshaw and Landis circuit operates more like the Goulding and Landis BDC circuit since it relies upon a pulse-by-pulse correction. Goulding, Landis, and Hinshaw [8] report, however, that the Hinshaw and Landis circuit is ineffective in correcting for deep-level charge trapping. They explain this result in the following manner. Interactions which occur near the outer periphery of the

detector incur the greatest majority carrier charge trapping deficits since the majority charge generated in this region travels the greatest distance before collection. The UPDTs generated by these events are correspondingly large, and the Goulding and Landis circuit will calculate its largest correction for these interactions. Figure (3-4) shows that the ballistic deficit correction calculated by the Goulding and Landis circuit for these outer periphery events is too large. Thus, a circuit which corrects well for charge trapping, over corrects for the ballistic deficit of outer periphery interactions. Conversely, circuits which correct well for ballistic deficit under correct for the charge trapping deficit of outer periphery events. The Hinshaw and Landis circuit calculates a proper ballistic deficit correction, and thus under corrects charge trapping. In other words, the ability of the Goulding and Landis circuit to correct for charge trapping is the result of the over correction this circuit makes for the ballistic deficit of outer periphery interactions, while methods such as the Hinshaw and Landis circuit which accurately correct for the ballistic deficit of outer periphery events are ineffective for deep-level, charge trapping correction.

Like the gated integrator, the Hinshaw and Landis circuit can correct for shallow-level charge trapping. However, as noted in Chapter 3, the Hinshaw and Landis circuit loses accuracy when the total charge collection time

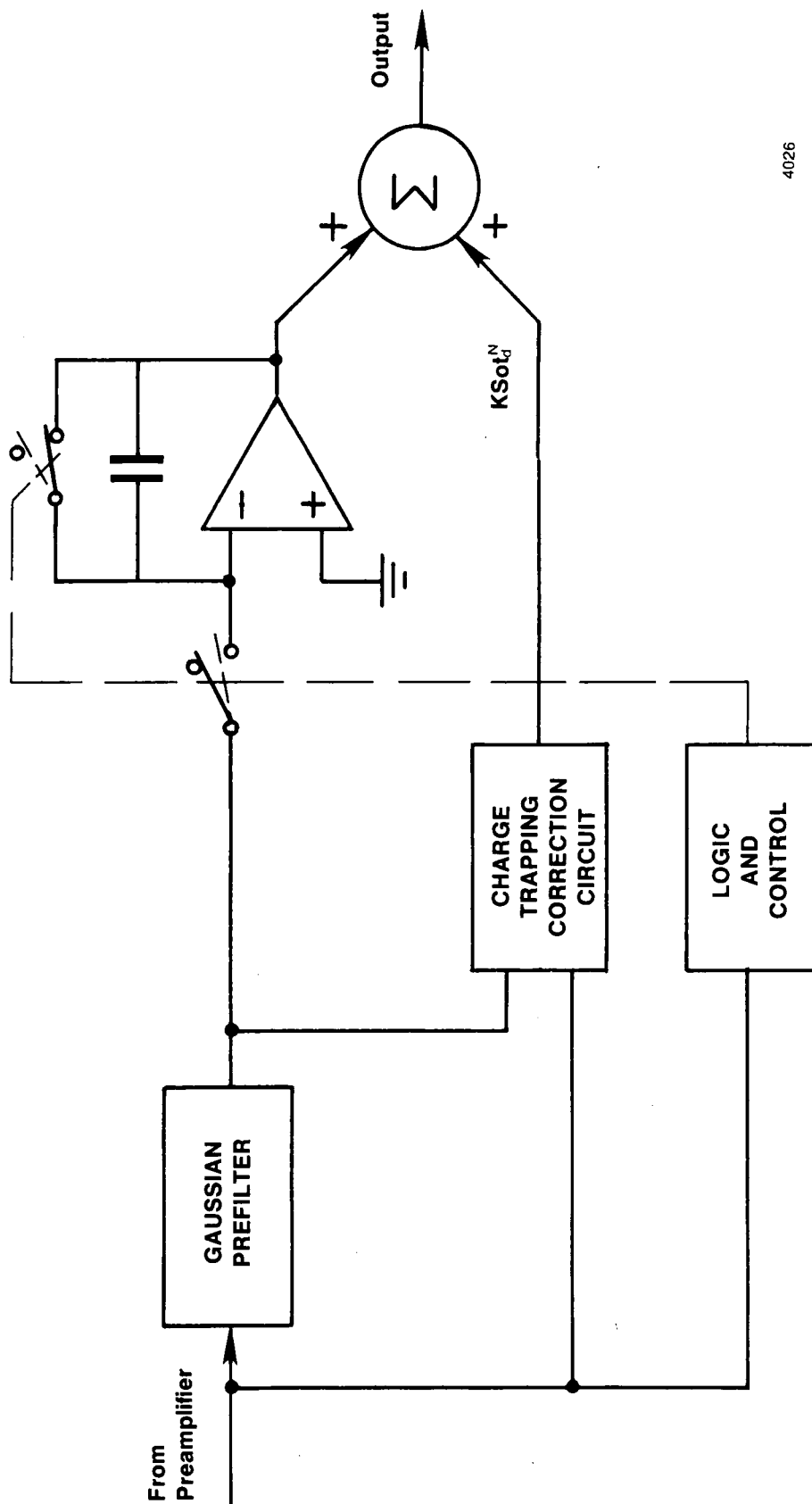
approaches 20% of the peaking time of the spectroscopy amplifier .

4.4 Gated Integrator with Charge Trapping Correction

The information presented in this chapter and in Chapter 3 indicate that the gated integrator and the Hinshaw and Landis BDC circuit do a very good job of correcting for ballistic deficit, but are poor deep-level charge trapping correctors. Conversely, resolution enhancement circuits such as the Goulding and Landis BDC circuit correct for deep-level charge trapping well, but lose accuracy in ballistic deficit correction when short shaping times are employed. A combination of two of these instruments is needed, then, to allow for the best performance in charge trapping and ballistic deficit correction.

Simpson et al. [24], reported an instrument which combines a gated integrator with a resolution enhancement circuit to correct for deep-level charge trapping. This circuit is shown in block diagram form in figure (4-3). The results obtained using this instrument are reported in Chapter 5.

Goulding, Landis and Hinshaw [8] have also proposed an instrument which combines a charge trapping correction



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Figure (4-3). Gated integrator with CTC block diagram.

circuit with the Hinshaw and Landis BDC circuit. Results obtained with this instrument are yet to be reported.

CHAPTER 5

EXPERIMENTAL RESULTS

The previous four chapters have detailed the theories behind charge trapping and ballistic deficit, and described instruments which compensate for these phenomena. In this chapter, the experimental data taken with some of these instruments over wide energy and shaping time ranges are presented and compared with the modified Trammell-Walter theoretical model given in Chapter 2.

5.1 Ballistic Deficit Correction

While it is possible to completely remove ballistic deficit by choosing a suitably long shaping time, the effects of charge trapping are independent of the spectroscopy amplifier. Thus, it is impossible to experimentally study "ballistic deficit only" correction. However, with very pure (trap free) Ge detectors, ballistic deficit effects will dominate over charge trapping effects when the charge collection times are greater than 20% of the spectroscopy amplifier peaking time.

Figure (5-1) illustrates the dramatic improvement in resolution possible with the Goulding and Landis BDC circuit. In this figure, a reverse electrode detector with

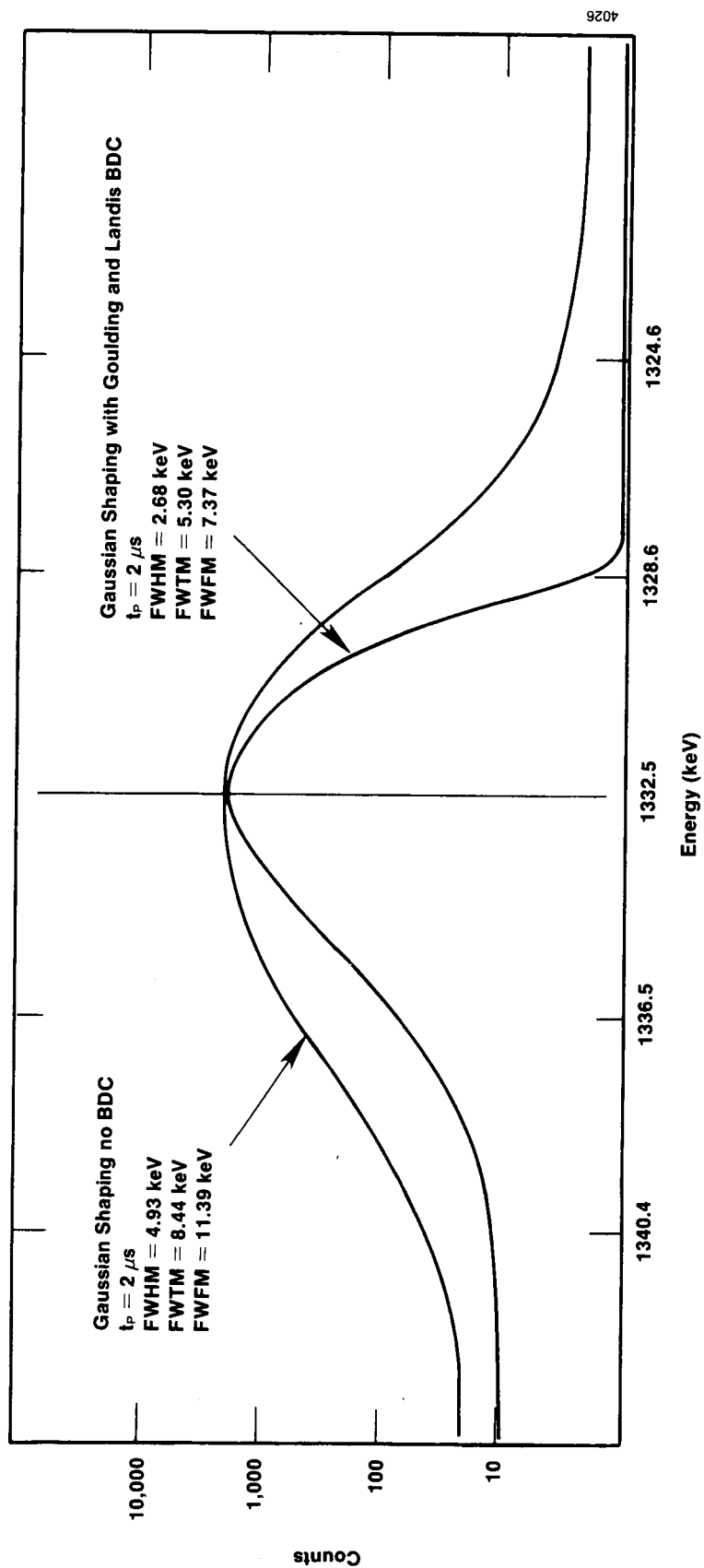


Figure (5-1). Corrected and uncorrected spectra with 33% reverse electrode detector.

a relative efficiency of 33% produced a 1.33 MeV energy peak with a FWHM resolution of 4.93keV. The peaking time of the spectroscopy amplifier was 2 us. However, when the Goulding and Landis circuit was used with the same detector and shaping time, the resolution improved to 2.68keV. Also, as figure (5-1) shows, large improvements were made in the full-width-at-one-tenth-maximum (FWTM) and full-width-at-one-fiftieth-maximum (FWFM) resolutions.

Figures (5-2), (5-3), and (5-4) show the FWHM resolution as a function of deadtime-per-pulse for detectors with relative efficiencies of 76%, 56%, and 29% respectively. The deadtime-per-pulse is the amount of deadtime for one nonpiled-up event as calculated using the Gedcke/Hale livetime correction method [11]. The use of this abscissa allows the direct comparison of resolutions obtained from spectroscopy amplifiers with vastly different shapes (such as a gated integrator and a Gaussian shaping amplifier). Also shown in figures (5-2) - (5-4) are the FWHM resolution predicted by equation (2-24) for the cases of 0% and 100% ballistic deficit and charge trapping correction.

A comparison of the results of figures (5-2) -(5-4) obtained with a Gaussian shaping amplifier alone with that obtained with the Goulding and Landis BDC circuit shows that the BDC circuit significantly improves the resolution at the shorter shaping times. Also, modest resolution improvements

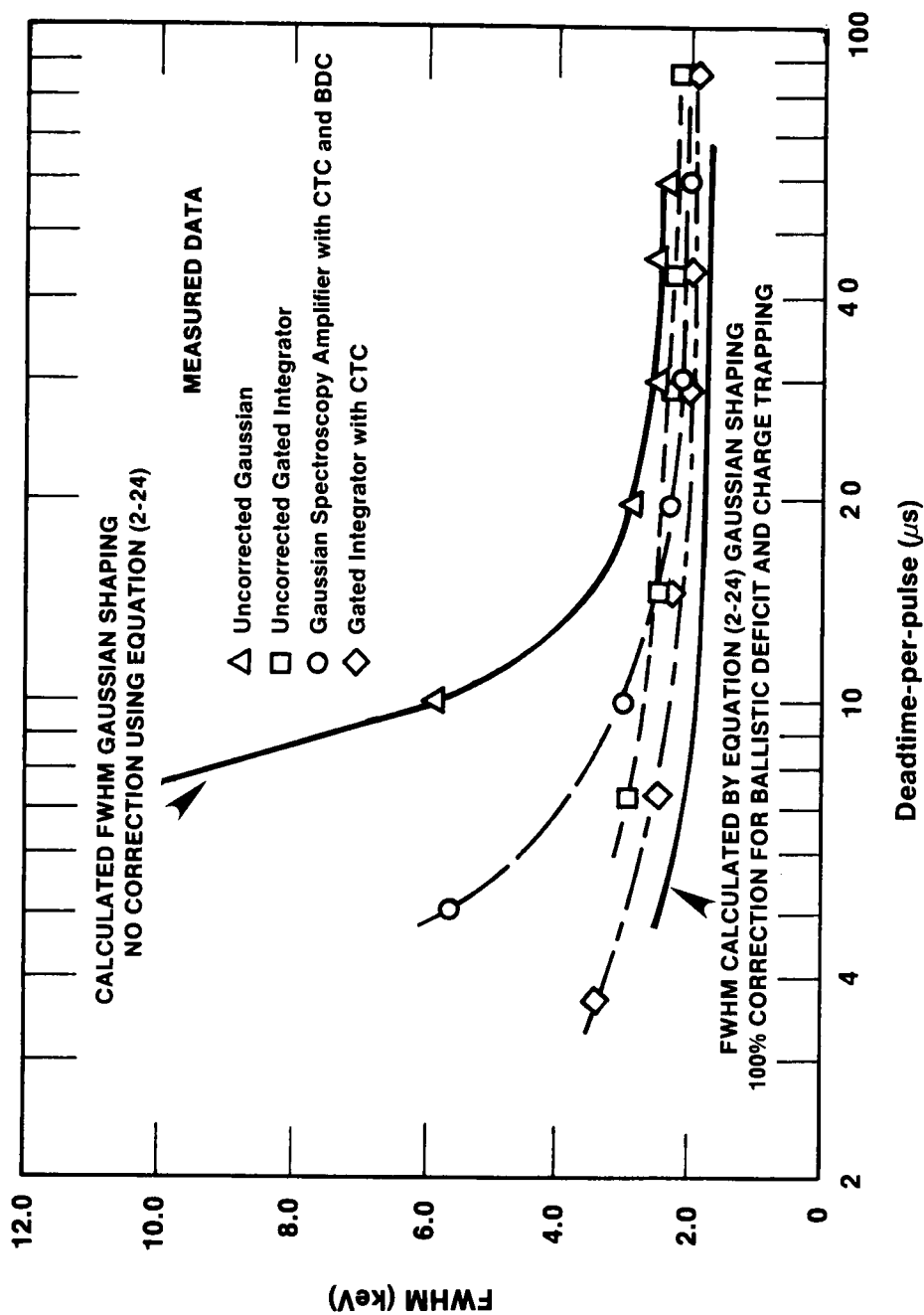


Figure (5-2). Measured and calculated FWHM vs. deadtime-per-pulse for a reverse electrode detector with 76% relative efficiency.

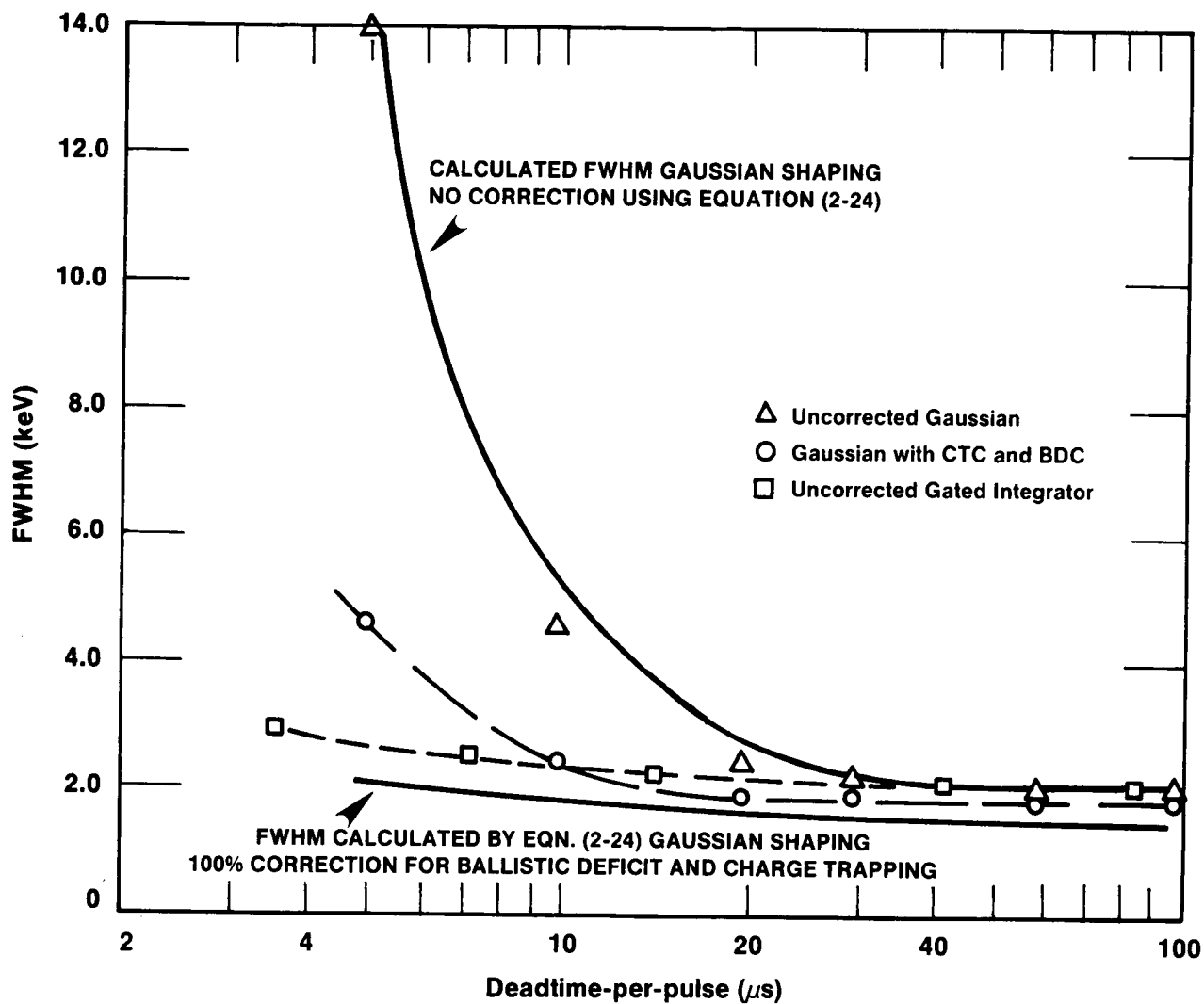


Figure (5-3). Measured and calculated FWHM vs. deadtime-per-pulse for a reverse electrode detector with a relative efficiency of 56%.

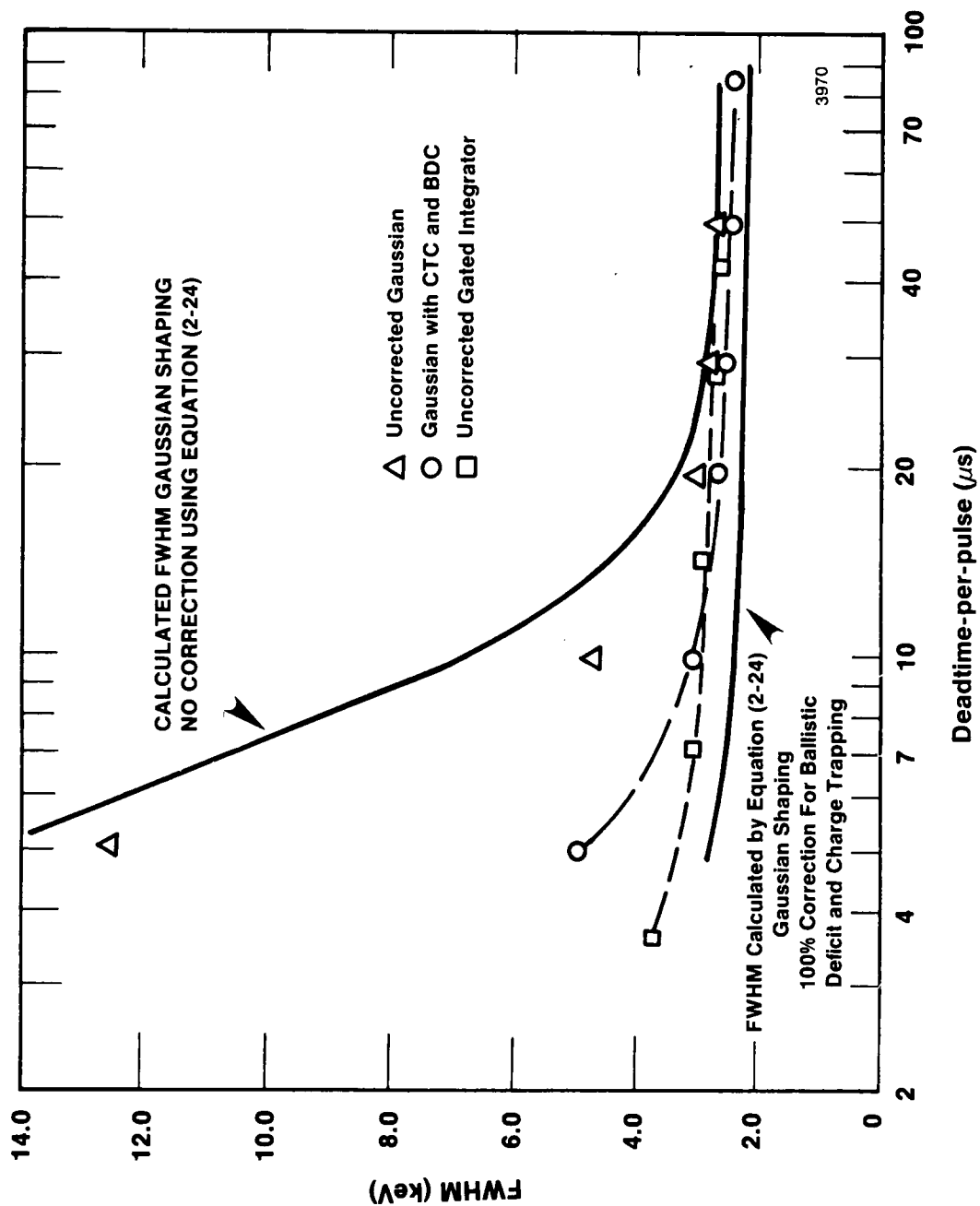


Figure (5-4). Calculated and measured FWHM vs. deadtime-per-pulse for a reverse electrode detector with a relative efficiency of 29%.

are seen at the longer shaping times due to charge trapping correction. The quality of the correction, as compared to the calculated 100% correction case, is seen to degrade as the shaping time is shortened. This is consistent with the discussion in Chapter 3.

At long shaping times, the gated integrator provided better resolution than the Gaussian shaper, but worse than the BDC circuit, at the 1.33 MeV energy line for the detector of figure (5-2). This result is due to the correction that the gated integrator makes for shallow level trapping, while making no correction for deep level trapping. It should be noted that at energies below about 800keV, where the electronic noise is a more important component of the resolution, the Gaussian shaping amplifier would provide better resolution than the gated integrator.

As the shaping time is reduced, the quality of the ballistic deficit correction made by the gated integrator does not decrease. Thus, at some deadtime-per-pulse the gated integrator provides better resolution than the BDC circuit. As shown in figures (5-2) - (5-4), this crossover point occurs at about 10 μ s of deadtime-per-pulse, which corresponds to Gaussian shaped pulses with peaking times of about 2 μ s. This crossover point is, in general, a function of the density of deep level traps in the detector. For example, the BDC will provide better resolution than the

gated integrator at a shorter shaping time if more deep level trapping is present.

5.2 Charge Trapping Correction

Figure (5-5) illustrates the improvement in resolution possible when charge trapping correction ($N=2.3$) is used. A radiation damaged, conventional electrode detector with a relative efficiency of 14% was used to generate these spectra. Without a CTC circuit, this detector produced a FWHM resolution at the 1.33 MeV energy line of 4.30 keV. Since the t_p of the Gaussian spectroscopy amplifier was >12 μ s, the effects of ballistic deficit were negligible, and this relatively poor resolution was due to the majority carrier trapping caused by the radiation damage. With the $N=2.3$ CTC circuit, the FWHM resolution improved to 2.00 keV. Also, the FW1/10M improved from 8.45 keV to 3.86 keV and the FW1/50M went from 11.0 keV to 5.55 keV.

Since radiation damage tends to provide a uniform distribution of majority carrier traps in conventional electrode detectors, this detector was used to determine the optimum value of N . Figure (5-6) depicts the FWHM resolution versus N . The resolution is a strong function of N for $N < 1.8$. However, for larger values of N , the resolution has only a weak N dependence. The best FWHM

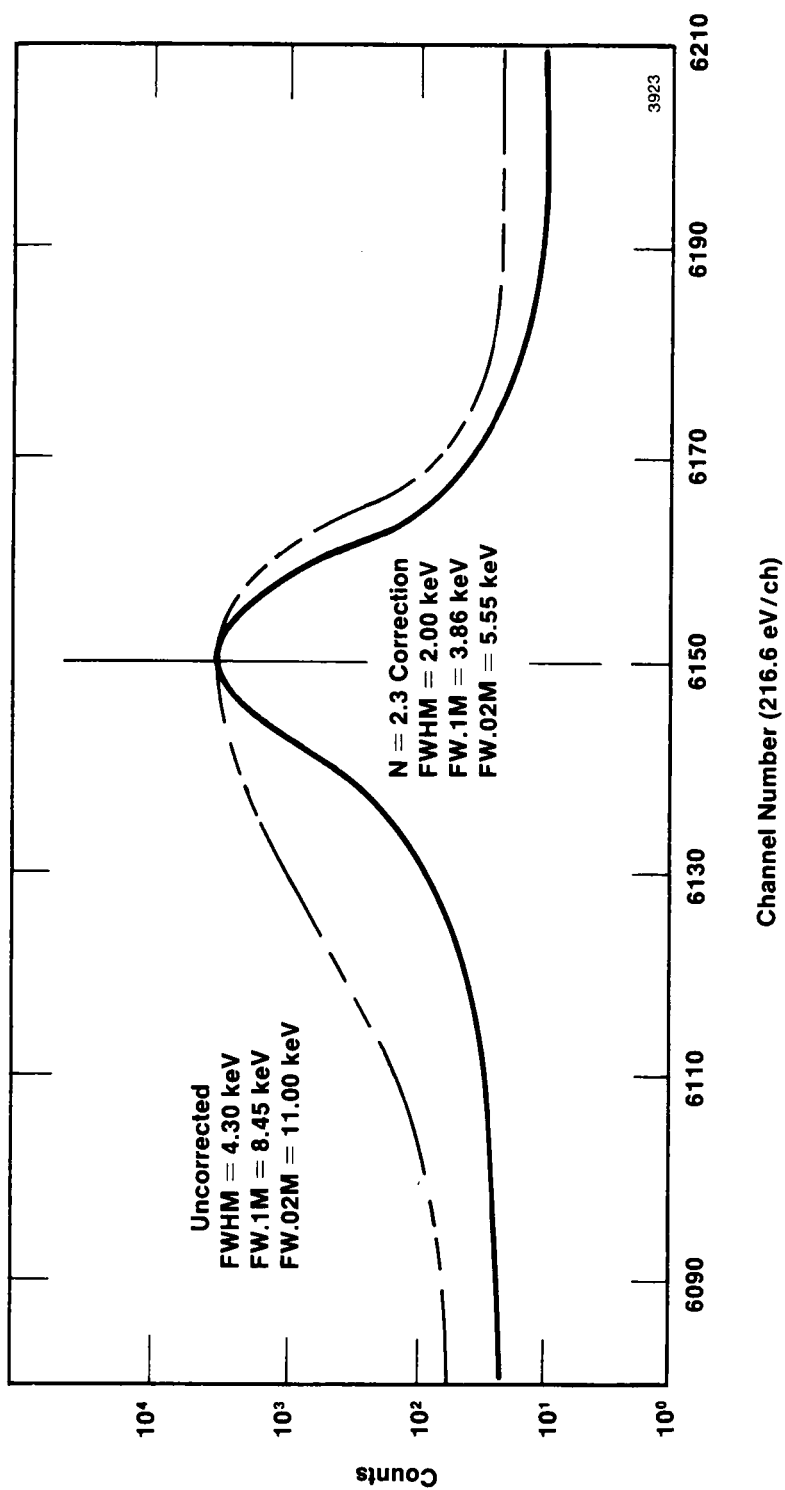


Figure (5-5). $N = 2.3$ corrected and uncorrected spectra for a radiation damaged, 14% efficient, conventional electrode detector.

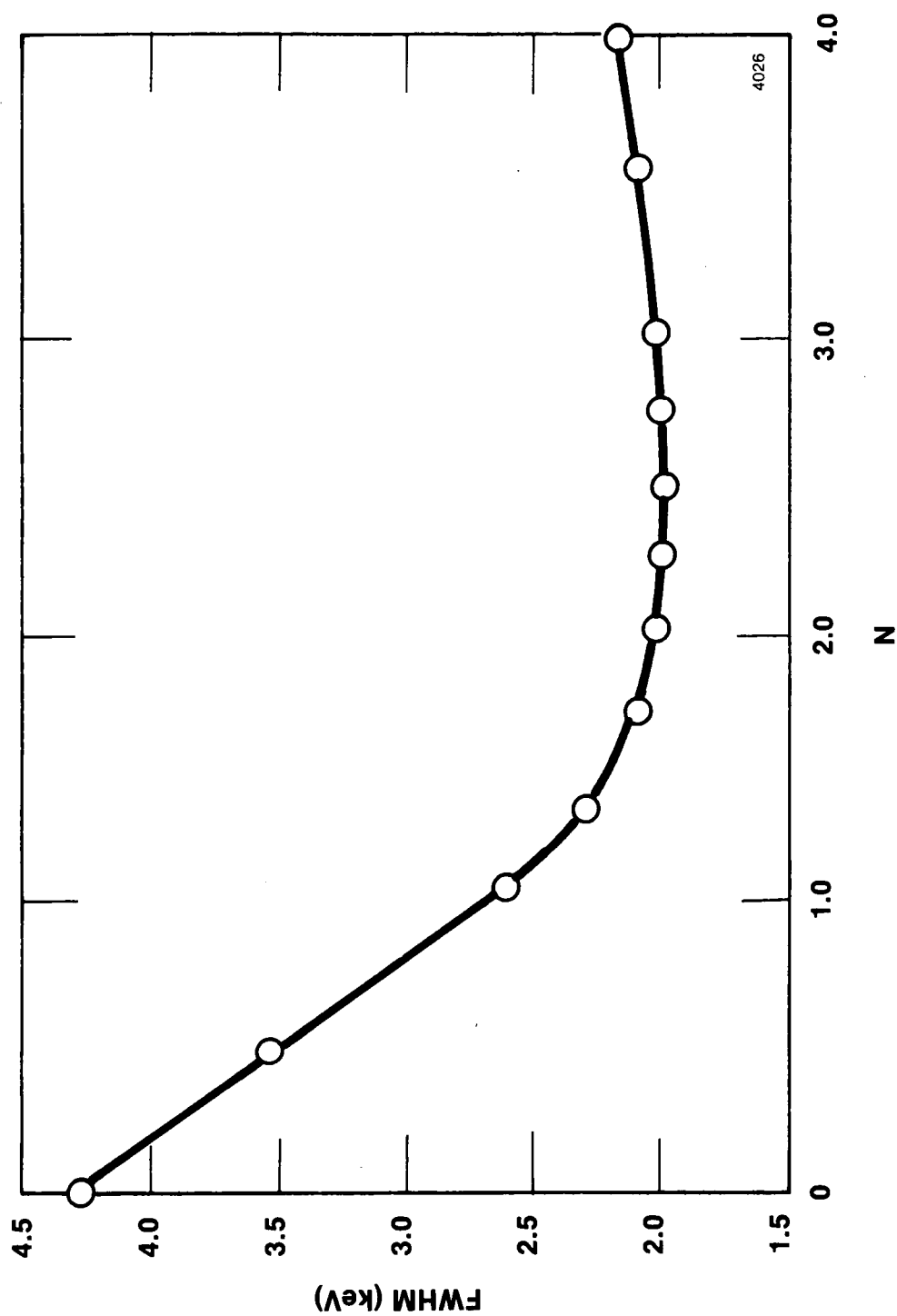


Figure (5-6). FWHM of the radiation damaged, conventional electrode detector vs. N.

resolution was obtained when $2.18 < N < 2.5$, which compares well with the value of 2.38 predicted in Section 4.1.1.

Chapter 4 predicts that the charge trapping deficit is strongly dependent upon the electric field in the detector. This dependence is illustrated in figure (5-7) where the corrected and uncorrected FWHM resolutions for a reverse electrode detector with a relative efficiency of 41% are plotted versus detector bias. The FWHM resolution obtained with a gated integrator is also plotted in this figure. As predicted, the resolution degradation increase as the bias decreases. This figure also shows that an $N=3$ correction produces better resolution than an $N=2$ correction at high bias levels. This was predicted in Chapter 4 for the case of a non-uniform trap distribution where the largest density of traps was near the outer periphery of the detector. A high bias level is required to obtain the $1/r$ electric field dependence needed for the $N=3$ prediction to be valid. As figure (5-7) shows, the gated integrator provides better resolution than the Gaussian amplifier at low bias levels. This suggest that more shallow level trapping is present at lower bias settings.

Another prediction of Chapters 2 and 4 was the energy dependence of the charge trapping deficit. This dependence is illustrated in figures (5-8) - (5-10) where the corrected and uncorrected FWHM resolution of the detectors from figures (5-2) - (5-4) are plotted versus energy. Again, the

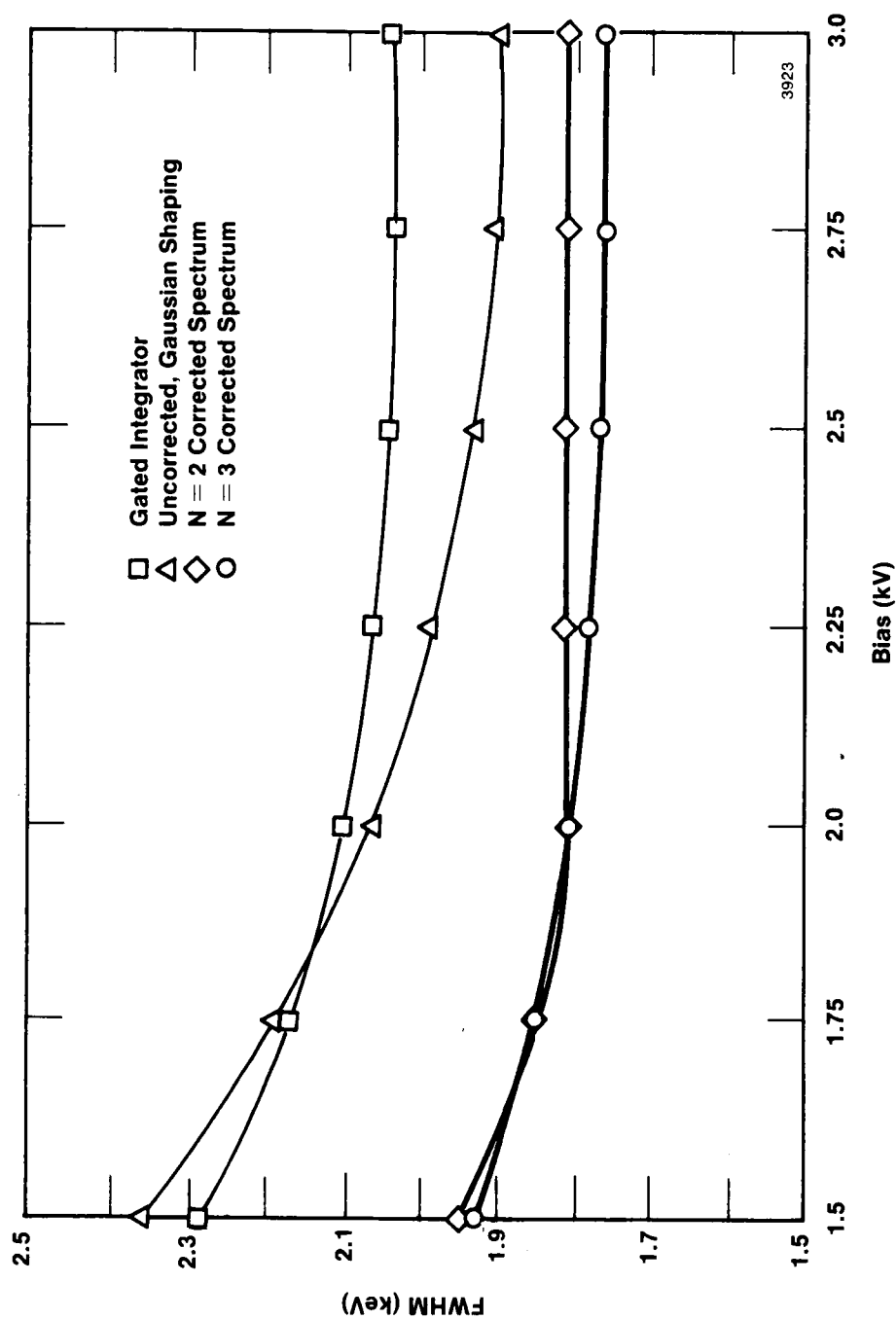


Figure (5-7). FWHM vs. detector bias for uncorrected Gaussian shaping, gated integrator, N=2 corrected and N=3 corrected spectra for the ^{60}Co 1.33 MeV line, 12 μs peaking time. The detector depletion voltage was 700 V, and the detector efficiency was 41%.

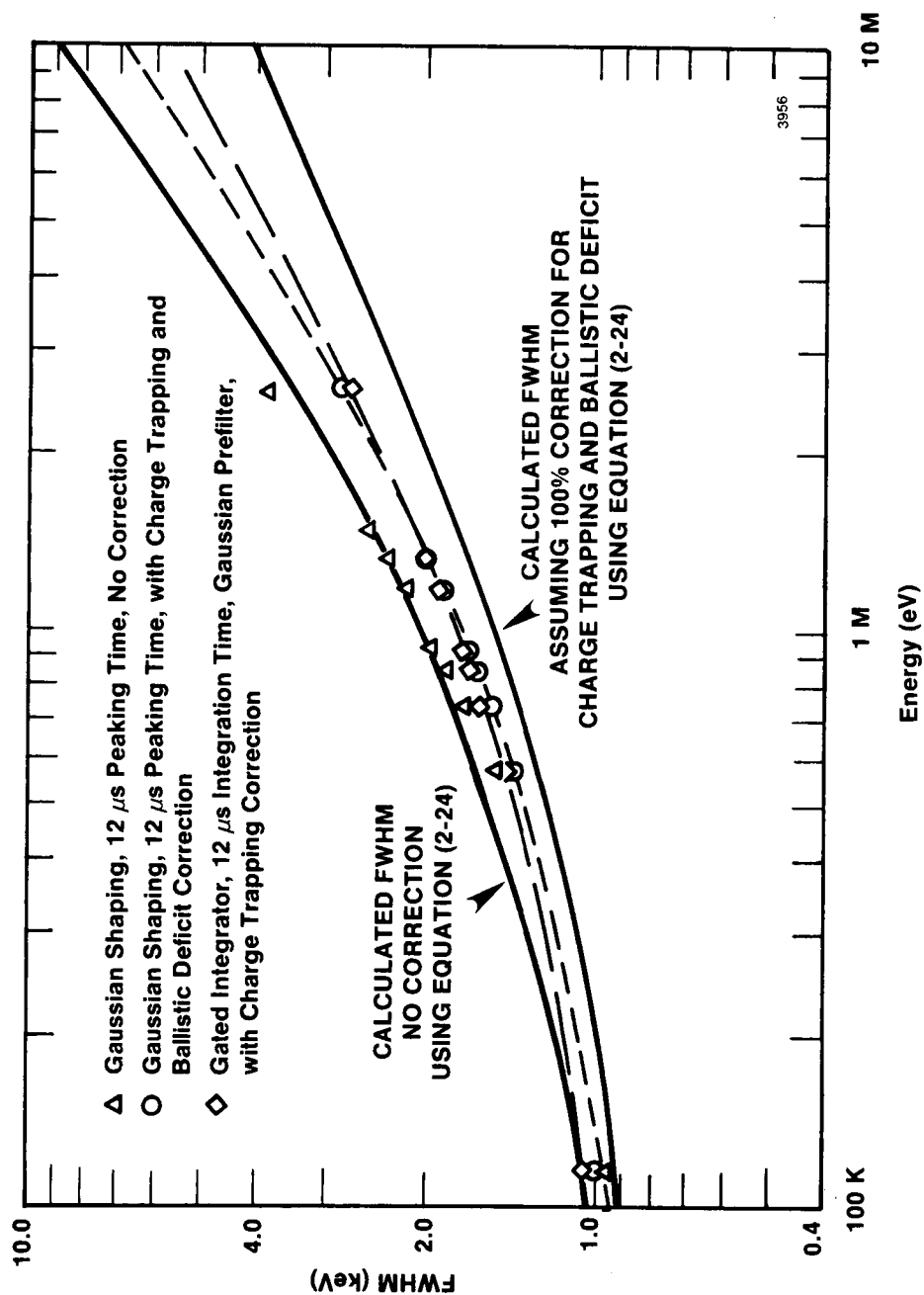


Figure (5-8). Calculated and measured FWHM vs. energy for a reverse electrode detector with 76% efficiency.

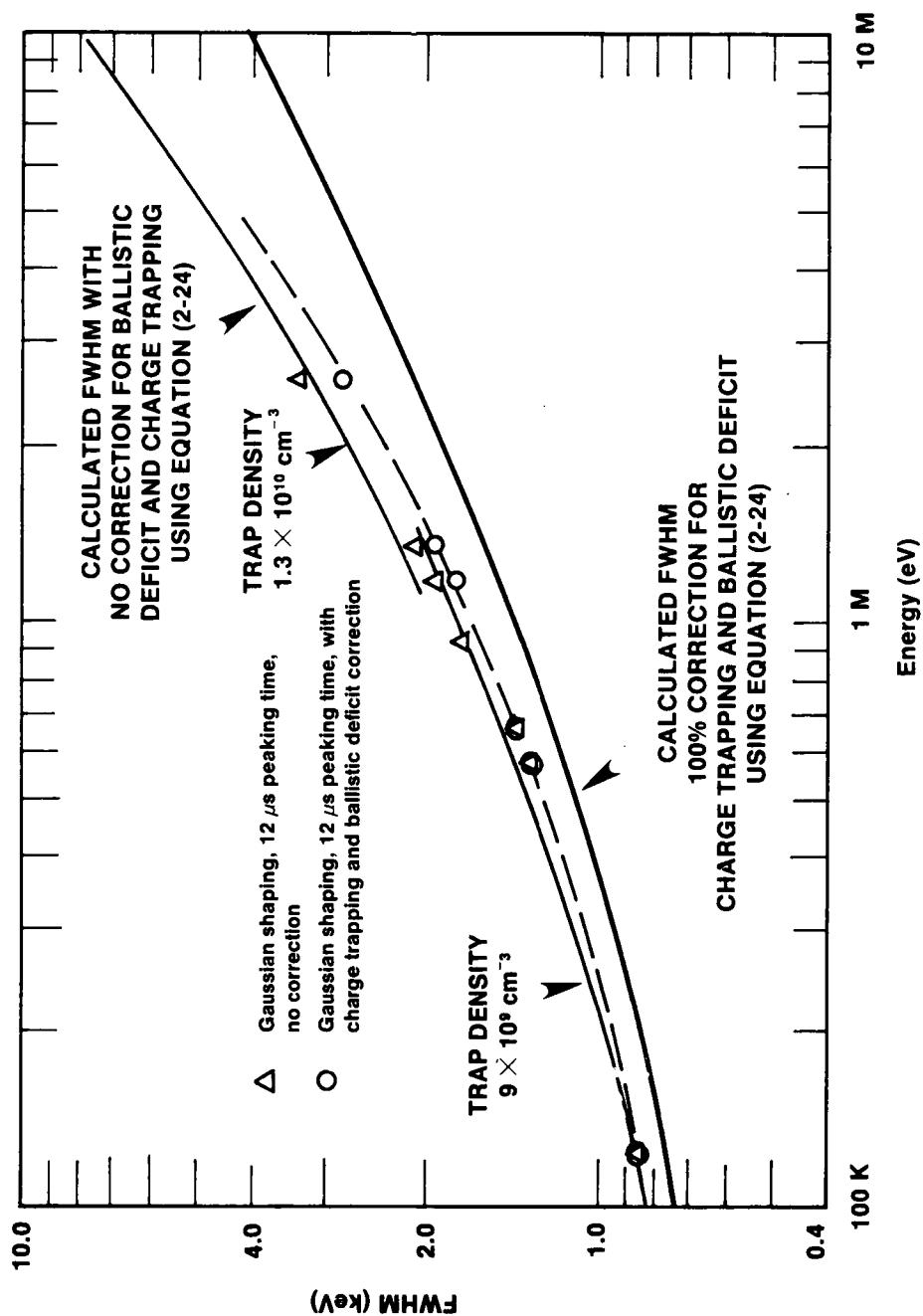


Figure (5-9). Calculated and measured FWHM vs. energy for reverse electrode detector with a relative efficiency of 56%.

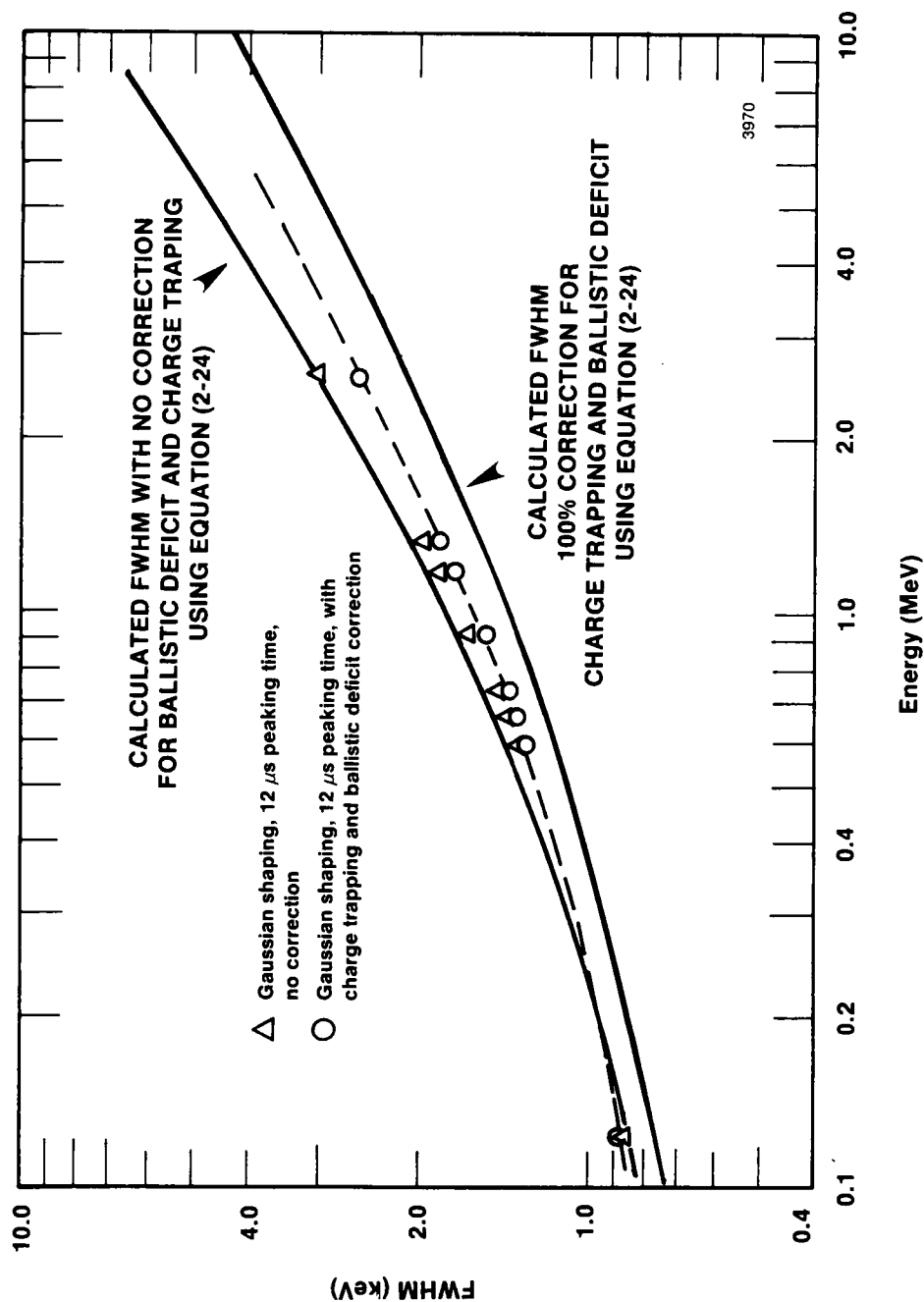


Figure (5-10). Calculated and measured FWHM vs. energy for reverse electrode detector with a relative efficiency of 29%.

shaping time of the amplifier was chosen large enough so that the ballistic deficit was small when compared to the charge trapping deficit. Also shown in these figures are the FWHM resolutions calculated with equation (2-24) for the cases of 0% and 100% correction for charge trapping deficits. For good agreement between the experimental and calculated results in figure (5-9), the calculation was split into a high and a low energy section. A higher charge trap density was required in the high energy region, which could be accounted for by a Z dependence of the trap distribution [24]. Higher energy gamma-rays are more likely to interact at locations deep in the detector. Thus, if the trap density is higher deep in the detector than nearer the surface, the trap density appears to be energy dependent.

Figures (5-8) - (5-10) show that resolution improvements made by charge trapping correction are more significant at the higher energies. Table (5-1) shows the corrected and uncorrected resolution for each detector at 1.33 MeV (measured) and at 10 MeV (calculated or extrapolated from the figures).

Detector from Figure:	Gaussian No Correction	Shaping Correction	Corrected Gated Int.	E
(5-2)	9.3keV	7keV	5.6keV	10MeV
(5-3)	8keV	6.7keV	_____	10MeV
(5-4)	7.3keV	5.6keV	_____	10MeV
(5-2)	2.2keV	1.9keV	1.9keV	1.33MeV
(5-3)	2.0keV	1.85keV	_____	1.33MeV
(5-4)	1.95keV	1.85keV	_____	1.33MeV

Table (5-1). Corrected and uncorrected FWHM resolution of the detectors in figures (5-2) - (5-4) at 1.33MeV and 10MeV.

5.3 Gated Integrator with Charge Trapping Correction

As described in Chapter 4, for the best resolution over wide shaping time and charge trap density ranges, an instrument which combines an efficient ballistic deficit correction with a charge trapping deficit correction is needed. Section 4.4 describes an instrument which combines a gated integrator with a charge trapping correction circuit. This section presents experimental results obtained with this instrument.

Figure (5-2), in addition to the results described earlier, shows the FWHM resolution obtained with the combination gated integrator/CTC circuit as a function of deadtime-per-pulse. This figure shows that this instrument offers the best resolution of all the instruments tested with this detector at the 1.33 MeV energy line. Figure (5-11) is identical to figure (5-2) except the gamma-ray energy is 2.6 MeV. This figure illustrates that charge trapping correction in combination with ballistic deficit correction becomes more important at higher energies, which is expected since both deficits are directly proportional to energy.

Figure (5-8) also shows the FWHM resolution obtained with this combination instrument as a function of energy. At the lower energies, the Gaussian shaping amplifier with or without charge trapping correction provide the best resolution, while at the higher energies the gated

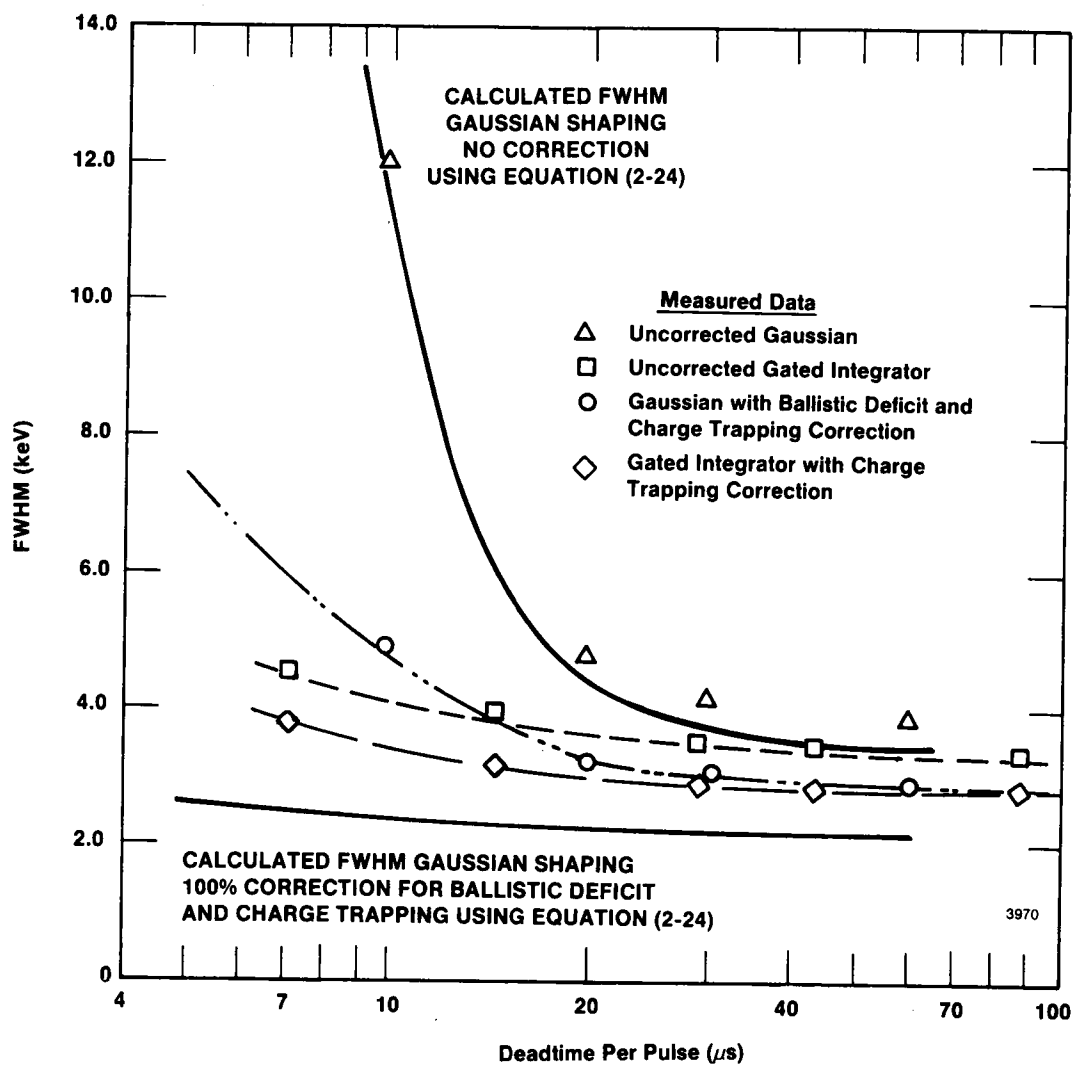


Figure (5-11). FWHM vs. deadtime-per-pulse at 2.61 MeV for a reverse electrode detector with 76% relative efficiency.

integrator with a CTC circuit allows for the best resolution. This can be explained as follows. At the lower energies, the major component of the resolution is electronic noise. Thus, the superior noise performance of the Gaussian amplifier provides the better resolution in this regime. However, as the energy is increased, charge trapping becomes a more important component of the resolution. Therefore, in the intermediate energy range the Gaussian shaping amplifier with a CTC circuit generates the best resolution. Finally, even at long shaping times, some ballistic deficit is present at high energies. Thus, an instrument which corrects well for both ballistic deficit and charge trapping is required for the best resolution at the higher energies.

CHAPTER 6

SUMMARY AND CONCLUSIONS

As stated in Chapter 1, the goals of this work are to fully develop and unify the theories of ballistic deficit and charge trapping; to develop an equation from which energy line shapes can be calculated for arbitrary charge trap distributions and spectroscopy amplifier shaping times; and to analyze and develop design equations for resolution enhancement circuits, including a complete noise analysis and treatment of non-ideal qualities of the correction.

Ballistic deficit is defined in equation (1-9), while its effect is shown in figure (1-8). In Section 2.4, approximate analytical expressions for the ballistic deficit are given in equations (2-10) and (2-11). Also, the results from a more accurate computer analysis using actual preamplifier signals as excitation sources are shown in figure (2-8).

In Chapter 3, the basis of the Goulding and Landis BDC circuit, equation (3-6), is derived, and this BDC circuit is thoroughly analyzed, including the description of non-ideal properties of the correction in Section 3.2.1. Short descriptions of the gated integrator and the Hinshaw and Landis BDC circuit are also presented in Chapter 3, which ends with a comparison of these three methods of ballistic deficit correction.

The subject of charge trapping is introduced in Section 1.2, and an intuitive treatment of the theory is presented in equations (1-15) - (1-20). In Section (2.5), a more detailed analysis of charge trapping is given, and figure (2-6) shows an energy line generated by a planar detector broadened only by charge trapping.

In Section 4.1.1, the basis of the majority carrier charge trapping correction circuit, equation (4-4) is derived. A similar equation is developed in Section 4.1.2 for the case of minority carrier trapping. Chapter 4 presents a majority carrier charge trapping correction circuit, including a complete analysis of non-ideal qualities in Section 4.2.1-1. The charge trapping correction capabilities of the gated integrator and the Hinshaw and Landis BDC circuit are explored in Section 4.3. Chapter 4 concludes by introducing a circuit which combines a gated integrator with a charge trapping correction circuit.

In Section 4.2.2 a minority charge trapping correction method is presented. Minority carrier charge trapping is most frequently encountered in the important case of radiation damaged, reverse electrode detectors. However, as stated in Chapter 4, these detectors can be repaired through a fairly simple annealing process. Thus, there has not been a great demand from users of HPGe detectors for a minority charge trapping correction circuit.

Equation (2-22) gives the Trammell-Walter equation for a coaxial detector. This equation allows the calculation of energy line shapes for charge trap distributions which can vary arbitrarily with r . Distributions which have Z or Φ dependence can be handled by expanding equation (2-22) to a surface or volume integral. Equation (2-24) gives a modified Trammell-Walter equation which includes the effects of ballistic deficit. This equation, then, allows energy line shapes to be calculated for arbitrary charge trap distributions (r dependence only) and spectroscopy amplifier shaping times. Again, Z or Φ dependence of the charge trap distribution can be handled by expanding equation (2-24) to a surface or volume integral. Figures (5-2) - (5-4) illustrate the accuracy of the modified Trammell-Walter equation for arbitrary spectroscopy amplifier shaping time, while figures (5-8) - (5-10) show the accuracy of this equation over a wide energy range with various levels of deep level charge trapping.

The results in Chapter 5 show the dramatic resolution improvements possible with the Goulding and Landis BDC circuit, the charge trapping correction circuit, and the gated integrator with a charge trapping correction circuit. As predicted in Chapter 3, figures (5-2) - (5-4) show that the Goulding and Landis BDC circuit does not fully correct for ballistic deficit at short shaping times. Hinshaw and Landis also report that their BDC circuit loses accuracy

when the preamplifier signal risetime exceeds 20% of the peaking time of the spectroscopy amplifier. Thus, when ballistic deficit is the dominant resolution limiting factor, the gated integrator provides better resolution than the other BDC circuits considered in this work.

Neither the gated integrator nor the Hinshaw and Landis BDC circuit effectively correct for deep level charge trapping. Therefore, when deep level charge trapping is the dominant resolution limiting factor, the charge trapping correction circuit provides for the best resolution. The gated integrator with charge trapping correction, however, may be the best instrument over all, since it allows for good resolution when ballistic deficit or charge trapping limited. The only disadvantage of the gated integrator with charge trapping correction is that the electronic noise is greater than that of the Gaussian spectroscopy amplifier. This is only important at low energies where the electronic noise is a significant component of the resolution.

Two rather surprising conclusions can be drawn from the results shown in figures (5-8) - (5-10). First, the resolution of detectors can be improved at high energies (>5 MeV) even if the detectors exhibits negligible effects from deep level charge trapping at the ^{60}Co 1.33 MeV energy line. Thus, essentially all reverse electrode detectors can benefit from charge trapping correction at high energies. Second, ballistic deficit can be significant at energies

above 2 MeV, even at spectroscopy amplifier peaking times as large as 12 μ s. Therefore, conventional, Gaussian amplifiers are not well suited for high energy gamma-ray spectroscopy. Rather, a Gaussian amplifier with BDC and CTC circuits, or preferably, a gated integrator with charge trapping correction should be used for spectroscopy at energies greater than 2 MeV.

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APPENDIX

Standard Deviation of t_d^2

The n th moment of t_d is given by [17]

$$\overline{t_d^n} = (-j)^n \{d^n C(w)/dw^n\}_{w=0}, \quad (A-1)$$

where j is the imaginary operator and $C(w)$ is the characteristic function of the random variable t_d [17]. The standard deviation of t_d^2 , then, is

$$\sigma(t_d^2) = \{[d^4 C(w)/dw^4 - (d^2 C(w)/dw^2)^2]^{0.5}\}_{w=0}. \quad (A-2)$$

Assuming t_d is a Gaussian distributed random variable [17],

$$C(w) = \exp[-w^2 \sigma(t_d^2)/2] \exp[jw \overline{t_d}]. \quad (A-3)$$

Substituting equation (A-3) into equation (A-2) yields

$$\sigma(t_d^2) = [2(\sigma_{td})^4 + 4(\sigma_{td})^2(\overline{t_d})^2]^{0.5}. \quad (A-4)$$

Since $\overline{t_d} \gg \sigma_{td}$, $2(\sigma_{td})^4$ can be neglected in equation (A-4).

Then,

$$\sigma(t_d^2) = 2\overline{t_d}\sigma_{td}. \quad (A-5)$$

VITA

Michael Leon Simpson was born in Knoxville, Tennessee on March 5, 1961. He attended elementary and secondary schools in that city, and was graduated from West High School in June 1979. The following September he entered The University of Tennessee, Knoxville, and was awarded a Bachelor of Science in Electrical Engineering in August 1983. In the fall of 1983, he accepted a graduate research assistantship at the Fusion Energy Division of The Oak Ridge National Laboratory and attended graduate school at The University of Tennessee, Knoxville. In March 1987, he was awarded a Master of Science in Electrical Engineering.

He immediately re-entered The Graduate School of The University of Tennessee, Knoxville, and began working toward a Doctor of Philosophy with a major in Electrical and Computer Engineering, while working full-time at Oak Ridge Associated Universities. In August of 1990, he completed all the course work for this degree.

The author is a member of Eta Kappa Nu, Mensa, and the IEEE, and is a Reserve Fire Fighter for the Rural/Metro Corporation in west Knox County, Tennessee. He has been employed at EG&G ORTEC as a Research and Design Engineer since February 1988. Mr. Simpson is married to the former Alice Faye McCorkle of Memphis, Tennessee.