SIGNAL PROCESSING FOR RADIATION DETECTORS BASED ON INTEGRATOR WITH CHARGE COMPENSATED RESET

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By

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ABSTRACT

SIGNAL PROCESSING FOR RADIATION DETECTORS BASED ON INTEGRATOR WITH CHARGE COMPENSATED RESET

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Master of Science in Electrical Engineering

A typical signal processing scheme for spectroscopy measurements with radiation detectors consists of an analog integrator (charge sensitive preamplifier) followed by a band pass filter and pulse height analysis of individual pulses. Whenever an analog integrator is used there has to be a mechanism for reset, when it approaches saturation. In conventional systems this is done through a continuous injection of current (most often by a resistor), or through a large current pulse applied for a short period of time. The first approach (among many other problems) introduces a slow exponential decay component that complicates signal processing and can still lead to saturation at high rate of radiation events. The second reset strategy is discontinuous and abrupt, leading to time variance, dead time due to “blinding” of subsequent electronics, and other problems. This work describes the development of a Multichannel Analyzer (MCA) system based on simple analog and digital architecture that takes advantage of a novel idea for resetting a charge integrating preamplifier. We have developed a so called Free Running Integrator, which is an integrator that allows resets to be transparent to signal processing, without loss of signal charge due to abrupt current pulse injection. Dead time and signal processing complications due to resets are eliminated. The response of the Free Running Integrator to a current impulse is a step function with no exponential decay, which makes possible the use of simplified signal processing. A prototype is described whose signal processing consists of an ADC working at a very low sampling rate followed by simple digital processing algorithm. This implementation lends itself to applications where low power consumption, low cost, and simplicity are needed.
Chapter 1

Purpose, Framework, and Introduction

1.1 Project Framework

This work presents an analysis, design, and application of a system commonly referred to as a Multichannel Analyzer (MCA) used for radiation spectroscopy measurements. It implements a novel idea for amplification of detector signal using an integrator with transparent reset, which solves many problems with similar systems. The design allows a very simple architecture that avoids complications in signal processing.

Chapter 1 provides introduction to radiation, radiation detectors, and signal formation. This introduction may be trivial for physicists and those who are familiar with the subject, but it may prove beneficial for technical readers who have not studied radiation detection and nuclear instrumentation. The information in this chapter is quite condensed and should serve as general overview.

Chapter 2 focuses on standard electronics for radiation measurement, and attempts to give somewhat substantial introduction to theory and practice of electronic design. The topics include signal processing, amplification, noise considerations, and others, keeping in mind the nature of a detector signal and its origins. This also is only an overview.

Chapter 3 is where some specific unsolved problems with conventional amplifiers and filters are identified. They are compared to a new idea for an electronic integrator that provides a transparent reset functionality. A particular application for the device is described, but the design should not be considered application specific. The nature of MCAs is quite universal, which is the case here as well. Potential sources of performance degradation are mentioned and analyzed.

Chapter 4 describes in detail the physical implementation of the design including parts' and parameter selection along with justification for these choices.
Chapter 5 presents results from examination of the physical design and compares them to the performance of conventional detector systems. Integrator reset performance, which is at the heart of the new idea is examined carefully.

Chapter 6 presents potential improvements. Technological improvements include integrated circuit design, which for now is out of reach. The rest of the improvements have been tried in other devices developed by our group. The future work described is based on tested concepts and good potential for implementation.

1.2 Introduction To Radiation Detection

In this work, radiation and radioactivity refers to the presence and emission of high energy photons (gamma rays and X rays) and particles such as neutrons. In this chapter the interaction of neutrons and gamma rays with matter will be summarized, and some theory necessary for the following chapters will be introduces.

1.2.1 Gamma rays, X-rays, and Their Interaction With Matter

Gamma rays and X-rays are electromagnetic radiation, basically of the same nature as viable light and radio waves, except their energy is several orders of magnitude greater. The distinction between the two types is based on their origin.

Gamma rays are generated inside the nucleus of an atom when there is rearrangement of energy levels following excitation such as spontaneous decay or neutron capture. We also call gamma-rays, photons released from matter and antimatter annihilation reactions like electron and positron annihilation.

X-rays are high energy photons generated outside of the nucleus. One common mechanisms for their production is electromagnetic energy emitted due to abrupt velocity change when an accelerated electron interacts with matter. This process is called Bremsstrahlung. X-ray photon emission also occurs when electrons in an atom's orbitals
that have been excited by an external event return to their ground state [1]. These are called characteristic X-rays because different elements have specific electron orbital energy levels, so the energy of a photon emitted from transition between two of these levels is characteristic for the particular element.

The distinction between gamma and X-rays is only technical. As far as detection is concerned, only the energies of the emitted photon is of importance. It's important to mention that commonly used unit for energy of radiation particles is an electron-Volt (eV), which is the energy it takes to move a charge equal to the charge of an electron \((1.6 \times 10^{-19} \text{ C})\) across one volt electric field potential. KeV and MeV refer to Kilo-electron-Volt and Mega-electron-Volt respectively.

In general, the X-rays spectrum starts at lower energy, overlaps somewhat with gamma rays while gamma-ray energy extends to higher levels. The range where X-ray energies begin and end is not well defined, because the definition of X and gamma rays refer to origin process yielding, not in wavelength ranges.

They typical gamma (and X-ray) energies for nuclear safeguard analysis ranges from 10-keV up to \(~5\)-MeV, to which correspond a wavelength \(\lambda=hc/E\gamma\) (h the Plank constant, c speed of light) in the range: \(10^{-8} \text{ cm} >\lambda>2 \times 10^{-11} \text{ cm}\).

1.2.2 Neutron Radiation and its Interaction With Matters

Neutrons are subatomic particles with no electric charge, and they are unaffected by the Coulomb barrier. Due to this feature neutrons can yield many typed of nuclear reactions. For example they can have elastic scattering with nuclei of the medium, as well as inelastic scattering where the nucleus is left in an excited state with the emission of gamma ray; neutrons can have capture reaction where the incident neutron enters the nucleus, with subsequent decay by gamma emission. Among many nuclear reactions, two are important for the next chapters [1]:

\[
^{3}\text{He} + n \rightarrow ^{3}\text{H} + p + 0.764 \text{ MeV}
\]
This is a nuclear reaction between $^{3}\text{He}$ and neutrons that produces $^{3}\text{H}$ and proton (p). This neutron reaction is very important because it is used for the neutron detection in the 3He proportional counters.

Another very important nuclear reaction is the nuclear fission induced by a neutron, as for example with 235U uranium:

$$n + ^{235}_{92}\text{U} \rightarrow F_1 + F_2 + \nu$$

Where $F_1$ and $F_2$ are the fragment fission nuclei produced and $\nu$ the average neutron production emitted (~2-3) with an energy of about few MeV.

The probability of nuclear reaction occurring is a function of the neutron energy. The following nomenclature is commonly used to classify this energy: thermal neutrons, being the neutrons having an average kinetic energy of the motion of the atoms of the materials (energy in the order of 25- meV at standard temperature), fast neutrons, refers to neutron with several MeV of kinetic energies as the ones produced from fission reactions, or the 14-MeV neutron generated in the fusion reaction of deuteron and tritium. Many nuclear reactions have higher probability to occur at low neutron energy (typically at thermal energies). For that reason an important process is the slowing down of neutrons (called moderation) by elastic collision in medium containing light nuclei, such as hydrogen. In a laboratory arrangement this is typically done using high density polyethylene (HDPE), which contains a lot of hydrogen and carbon.

### 1.2.3 Gamma-rays and their Interaction with Matter

Gamma ray photons can transfer part or all of their energy to electrons in the detector medium. In the process, the gammas either disappear, or are scattered. The electrons that are accelerated by the energy transfer, deposited their kinetic energy as they are stopped.

- Three types of interactions of gamma with matter are important: photoelectric effect, Compton scattering, and pair production. Photoelectric absorption is a process where the energy is completely absorbed by an electron within an atom. Such energy transfer can't be done to free electrons. After the interaction, a vacancy is left in the shell where the electron used to reside, which could be filled by another electron resulting in emission of characteristic X-rays. The X-ray
could be reabsorbed, or leave the detector volume.

- Compton scattering occurs when a gamma transfers an arbitrary portion of its energy to an electron and changes direction as a result of conservation of momentum. This electron is known as a recoil electron.

- When a gamma photon has energy larger than twice the rest-mass of an electron (1.022MeV), a pair of electron and positron can be created in a process called pair production. This happens when there is a heavy particle such as a nucleus to absorb the momentum of the photon. The difference between the 1.022MeV rest-mass energy, and the original gamma energy is transferred to the stopping atom. The created positron is annihilated soon (when it encounters an electron), and exactly 1.022MeV gamma is produced. This gamma may be reabsorbed or could escape the volume of the detector. Absorption through pair production becomes important only with energies greater than 5 MeV [2].

The probability of the three listed phenomena is function of the atomic number Z of the atoms involved and the photon energy.

1.3 Signal Formation in Detectors

The radiation interactions mentioned above yield energetic charged particles (electron or proton and tritium nucleus). Unlike neutrons and gammas, charged particles are quickly stopped by multiple collisions with surrounding atoms. In this process, energy is deposited in the detector volume in the form of ionized or excited atoms. The generated charge in the form of free electrons (or ones that have been excited to the conduction band in solids) can be measured directly by collection and electronic amplification. Also the initial charge can undergo multiplication in the detector as is the case in proportional counters.

Detectors using scintillating materials generate signal indirectly. The de-excitation causes light emission (often visible light) in the material. The amount of produced light can be measured with a photodetector such as a photo multiplier tube (PMT) described later.
1.4 Quality of Energy Determination

An accurate measurement of the energy of radiation relies on proportionality in all of the above mentioned processes. The energy deposited by a gamma photon in the detector medium, for example should be proportional to the original energy of the photon (this can be true only when the gamma is absorbed completely). The conversion from deposited energy to charge has to be proportional. If there are additional steps like conversion to light, or charge multiplication (explained next), that process should be proportional to the original charge generated. And in the end the electronics that performs the measurement should give a result proportional to the charge was collected. The proportionality requirement, of course is not completely fulfilled most of the time. For some cases these conditions are sufficiently satisfied in a particular energy range and for a given detector; sometimes the conversion happens according a known dependence that can be taken into account after the measurement.

Measurements of the same deposited energy will have a statistical distribution with a certain mean and standard deviation \( \sigma \). The ability to resolve two mono-energetic distributions depends on how wide they are (which is related to their standard deviations). The quantity most often used to express the width is the full width of half maximum (FWHM), which is defined as the width of the distribution at half of its peak value [1]. Very often, spectroscopy measurements produce Gaussian distributions. In that case FWHM\(=2.35\sigma \), where \( \sigma \) is the standard deviation.

Many processes can degrade the energy resolution (such as noise in the electronics, measurement accuracy drift, etc.). However, the fundamental lower limit of the standard deviation, comes from the fact the number of charges or photons generated for given energy of interaction is discrete. If if their generation occurs independently from one another, then the probability density for the number of carriers \( n \) generated is a Poisson distribution

\[
P(n) = \frac{(N)^n e^{-N}}{n!}
\]

with an average number of charges \( N \) produced for given energy, and a standard deviation
\[ \sigma = \sqrt{N}. \] If \( N \) is large than the Poisson distribution can be approximated by Gaussian. The FWHM of a Gaussian is \[ FWHM = 2.35 \cdot \sqrt{N}. \] Many detectors show better than expected energy resolution probably because each charge carrier is not generated independently [1].

The only way to reduce this statistical uncertainty is to generate more charge carriers per deposited energy. For this reason semiconductor detectors have high resolution due to their low energy of ionization (the amount of energy needed to put carriers in the condition band). In all detectors the highest energy resolution is fundamentally determined by the lowest number of discrete carriers in the detection chain. Sometimes, the lowest number is not the number of charges generated after the interaction. For example in scintillation detectors light has to be converted to electrons. Commonly used light sensor is the Photomultiplier Tube (PMT), which has usually about 30% conversion efficiency. This reduces the number of carriers (called photo-electrons) by more than 1/3. Then the number of photo electrons will be the fundamental resolution limiting factor.

Another, less straight forward source of error is the statistical nature of the time of arrival of events. If two events occur close to one another in time, they could be indistinguishable and the measured amplitude will include parts of both pulses. This is called pileup. The probability of it happening increases exponentially with the average frequency of detected events (the count-rate).

### 1.5 Specific Detectors of Interest

In this work, we use two types of detectors: The first type is called proportional counter, which collects the primary ionization directly but instead of immediately amplifying the resulting current with an electronic circuit, the electrons from the ionization are accelerated and produce an avalanche of secondary ionization, as they encounter other atoms. This multiplies the available signal charge and makes amplification easier. The second type is a scintillation detector. As explained earlier, it uses light sensitive device attached to a scintillator material. The most common light detector is a Photomultiplier Tube (PMT).
1.5.1 $^3$He Proportional Counter

A very common type of slow neutron detector is a tube filled with $^3$He gas (Fig. 1.1). This isotope of helium has a high cross-section (probability of interaction) with thermal neutrons. The neutron capture reaction that occurs was described in section 1.2.2. The products are a triton (a tritium $^3$H nucleus) and a proton with kinetic energy of 0.764 MeV [1]. Some events deposit all that energy in the detector fill gas, but many have one or both fragments reach the tube wall before they are fully stopped. This is why if a distribution of amplitudes of the electric pulses produced by the interactions is collected in a histogram (so called pulse height spectrum), it will be a continuum instead of a single peak that corresponds to the full energy of the capture reaction.

![Diagram of $^3$He cylindrical tube shown on its side and in cross-section. There are indicated two neutron capture reactions that have tracks either parallel or perpendicular to the anode wire.](image)

Gamma-rays also deposit energy in the detector medium, but the charge produced is less than the neutron reaction. The gamma detector pulses are easily distinguished from from neutrons by their amplitude. This makes $^3$He excellent for separating gammas from neutrons. A typical pulse height spectrum taken with this type of detector is shown on Fig 1.2, where the neutron-gamma separation is visible.
The $^3$He detector is a gas proportional counter. Its energy deposition medium is a gas that fills the detector volume. Inside its enclosure there is a charge collection electrode (called anode), which is biased at high positive electric potential relative to the body. Free electrons created by ionization due to radiation interaction are attracted to the anode wire. If it has small enough diameter, the field strength will be large and it will accelerate the electrons to a point where a collision with gas atoms will cause further ionization freeing more electrons, which in turn are accelerated and cause ionization. The avalanche of electrons that is produced reaches the anode, which gives rise to a current signal at the detector output. There are heavy positive ions left behind that drift slowly towards the negative potential of the detector walls. They create a slow component that can lengthen signal the pulses if not filtered adequately.

Depending on the trajectory of the initial reaction products there can be different distributions of charge through the volume. Fig 1.1 shows two extremes. If the initial ionization track is parallel to the anode wire, all generated charge is the same distance from the anode, and it arrives at the same time creating a fast rising pulse. If the track is perpendicular to the anode, the time of arrival has a longer spread, and the pulse has a longer rise time. This gives rise a non-uniformity in detector pulse shapes which is illustrated by Fig. 1.3.
Fig 1.3 Non-uniformity in actual integrated $^3$He pulses showing both the time spread of the electron drift, and the slow ionic component. The corresponding pulses filtered with a fast shaper are shown on the left (signal processing will be discussed shortly). It should be noted that longer filtered pulses last beyond 2µs which is a drawback in this type of detector because it causes significant dead time.

The duration of a slow pulse can be couple of micro seconds. Incomplete charge collection by subsequent filtering stages results in variation in pulse height shown on the right plot on Fig 1.3. Fast filtering is done in many $^3$He systems in order to reduce dead time at the expense of pulse height uniformity. Details will be given in the following chapters.

1.5.2 NaI(Tl) Scintillation Detector

![NaI(Tl) Scintillator Diagram](image)

Fig 1.4 NaI(Tl) scintillator coupled to a PMT. The photo-cathode and dynode structure of the PMT are illustrated. This is a public domain image [3].

Scintillating materials like a NaI(Tl) crystal, have the property of luminescence, which is to produce a pulse of light (with wavelength of 415nm for this material [1]) when a radiation interaction causes excitation in the material. The following de-excitation
process results in visible light emission.

The time profile of the light emission can be approximated by exponential decay with one or more time constants. NaI(Tl) scintillators have two dominating decay components (a fast and slow) with time constants of 230ns, and 2.2μs. It has been also shown that the distribution of the signal between these components changes with temperature. The slow component becomes more significant at lower temperatures (contributing up to 40% of the light at -20°C), while near room temperature, it's negligible and the fast component dominates [4,5,6].

Despite its shortcomings (not so great resolution, non-negligible slow component in the time response, temperature dependence, and non-linearity), NaI(Tl) is by far the most popular scintillator material for spectroscopy. This is due to mature and inexpensive technology for crystal growth in many sizes and shapes, satisfactory resolution for many applications, and a wealth of experience and data from its use.

**PMT**

The Photomultiplier Tube is very commonly used with all kinds of scintillators. It offers very large gain and dynamic range, good linearity, low noise, and fast time response. Its principle of operation is as follows: Light photons are converted to electrons via photoelectric effect emission occurring on one of its electrodes called a photo-cathode. The efficiency of this conversion, regrettably is low (only 30% in normal PMT's and rarely more than 40% on devices with special photo-cathode materials). The emitted electrons are accelerated with an electric field towards another electrode called a dynode. In the collision a larger number of electrons are released via secondary electron emission. The acceleration and collision process can be repeated several times, multiplying the number of charges by a certain factor every step of the way. The last electrode, called anode, collects all the charges and outputs them as signal current.

An illustration of a PMT is shown on Fig 1.4 The dynode structure needs a supply of
gradually increasing high voltage potentials. The cumulative supply voltage for all of them could be from several hundred to thousands of volts depending on the desired multiplication factor and speed of the device.

It's important to gain appreciation for the role of the PMT. It is a device that both converts a very weak pulse of photons into a proportional number of electrons (with low efficiency unfortunately, only about 30%), and then amplifies the tiny amount of charge often to the order of $10^6$ or more. This amplification is performed with very little added electronic noise, which is mainly due to thermal electron emission in the photocathode (also amplified by the PMT's multiplication factor). As an example, a PMT with low dark current such as Electron Tubes' model 9807B has cathode thermal emission that is in the order of 1000 electrons per second at room temperature [7], whereas the signal from a 662keV gamma photon in NaI(Tl) scintillator is in the order of 7500phe in 230ns. This corresponds to SNR of $2.8 \cdot 10^8$. Also, thanks to the enormous gain, the noise contribution of any well designed electronics following the PMT is negligible.

There is uncertainty at each stage of electron multiplication, which adds to the signal fluctuation very similarly to the effect of noise (in fact, this property of the PMT is called excessive noise factor [1]). However, this effect can be mitigated only by improving the design of the PMT, and not with noise reduction techniques in the electronics.

Another parameter of the PMT is its impulse response. It can be approximated as a Gaussian pulse with a variance $\sigma$ of few nano-seconds [7,8]. This is small compared to the NaI(Tl) pulse fast and slow components 230ns, and 2.2us [1,4,5], but can be of greater importance for processing pulses from faster scintillators such as LaBr(Ce) crystal that has rise time of 30ns [9]. These details can have implications in determining the right type of signal processing.

1.6 Connecting Detectors to Subsequent Electronics

There are two ways to connect a detector to the electronics and a high voltage power supply. One option is when a positive high voltage supply is connected to the charge
collecting electrode (anode). Since the signal charge is produced by the anode, the electronics has to be connected through a coupling capacitor which is charged to the full high voltage potential (Fig 1.5a). The other way is to connect the anode directly to the electronics and be at near zero potential, and connect positive high voltage to the detector's body or cathode (Fig 1.5a).

![Fig 1.5a](image1.png) Detector connected via a coupling capacitor $C_D$. Positive high voltage applied to the anode.

![Fig 1.5b](image2.png) Detector connected via a coupling directly to the electronics. Negative high voltage applied to the body.

The first way is usually more convenient for construction pursuits, but has a significant disadvantage. The signal from the detector is AC coupled, which means that whenever there is a pulse that charges the coupling capacitor, there will be an exponential discharge process with time constant determined by $C_D$ and $R_L$. This creates signal processing complications that are described in the following chapter.

If the anode is connected directly to the amplifier, then the received signal is only one directional (negative current generated by the detector). This allows the use of an integrator that would have monotonically rising output signal. In the next chapters it will be shown why this is a useful feature.
Chapter 2

Standard Electronics for Counting and Spectroscopy

2.1 Introduction

There is a variety of radiation measurements for many kinds of scientific experiments and applications where the properties of different types of radiation are being used or explored. We will focus on what is called a pulse height analysis, which is measurement of the amplitude of charge pulses produced by a detector when some kind of radiation interaction occurs. This is usually done with the intent to observe part of the energy spectrum of the radiation that reaches the detector.

Figure 2.1 shows typical configuration for gamma spectroscopy measurement system using a NaI(Tl) scintillation detector. The system is called a Multichannel Analyzer (MCA). It produces a histogram of the distribution of amplitudes of measured detector pulses. This distribution corresponds to the energies of interaction inside the detector. To preserve the quality of the energy resolution it's important to minimize the degradation of the detected signal caused by the electronics. To arrive to a good quality pulse height measurement, each of the functional blocs plays an important role and has to be designed carefully.

![Diagram of gamma spectroscopy system](image)

**Fig 2.1** Typical gamma spectroscopy system. Its components are NaI(Tl) detector, preamplifier, filter/shaper, and a pulse height analyzer. The filtering can be analog followed by digitization, or digital preceded by digitization.
2.2 Preamplifier

Signals received from different types of detectors vary widely in terms of their strength. We designed our electronics to work with ones that provide internal multiplication of primary charge (like scintillator detector coupled to PMT, or proportional counter), which reduces the noise requirements on the electronics. Never the less, the detector output signal is quite weak (the precise value depends on the light output, and PMT multiplication). In order to perform filtering and digitization, it has to undergo initial amplification, which is performed by the preamplifier. When an event is detected, at the output, the detector produces certain amount of charge in the form of a current pulse with some time profile. One way to amplify the signal is to have this current produce a voltage drop across a resistor, and then use a voltage buffer to for power amplification as shown on Fig. 2.2a. Then \( V_i \approx I_i \cdot R_i \) if \( R_S \gg R_i \), which is generally true for PMTs and proportional counters.

![Fig 2.2a Voltage-sensitive preamplifier. Figure adapted from [10].](image1)

![Fig 2.2b Current-sensitive preamplifier. Figure adapted from [10].](image2)

However, there are some problems with this implementation: First, detectors are capacitive devices, and connecting a cable adds more parallel capacitance \( C_d \). The actual setup then will look like Fig2.2b instead. In this case, the charge impulse response will be an exponential decay \( h(t) = \frac{1}{C_d} e^{-t/RC_d} \) with \textit{amplitude} at \( t=0 \) that depends only on \( C_d \), which is not well defined, and could change when the connecting cable is changed for example. Also, the time response is sensitive to \( C_d \).

The second problem is that \( R_i \) is a source of thermal noise proportional to its resistance value which can degrade the signal to noise ration (SNR) if the input signal is weak.
2.2.1 Charge Sensitive Preamplifier

If we want to know what charge is produced by the detector, it makes more sense to measure it directly instead of its derivative (the current pulse). In analog circuits, taking the derivative amplifies high frequency noise, which may not be reversible with subsequent integration.

A better architecture is to have an integrating capacitor in negative feedback with a high gain amplifier as shown on Fig 2.3a. The feedback makes the input very low impedance (virtual ground), so a current pulse flows into the amplifier without charging $C_d$ thus removing the amplitude's dependance on detector and wire capacitance. If the gain $A$ is large, output voltage will be $v_o \approx -\frac{Q_i}{C_f}$. $R_i$ is eliminated and along with it the thermal noise that it contributes. $C_f$ is well controlled and stable.

This circuit is called a charge sensitive preamplifier, but it's just a current integrator, who's output is proportional to the charge produced by the detector in the integration period. The bandwidth requirements of the gain stage are relaxed, because integration suppresses high frequency components. The filters that follow don't have to work with fast current pulses, so they also can have smaller bandwidth.

After not too long, the integrator will reach saturation. To prevent that, there has to be a way to introduce current at the input with opposite direction to what the detector
produces. The easiest way, is to put a “bleeding” resistor parallel to \( C_f \) to discharge it. In that configuration, if there is an impulse current with integral \( Q_i \) at the input, the output would jump to value \( v_o = \frac{Q_i}{C_f} \) and decays exponentially with time constant \( \tau = C_f \cdot R_f \). The subsequent stages of the system will have to deconvolve this exponential characteristic. This deconvolution so called pole-zero cancellation.

This solution has the following problems: The noise advantage of the simple integrator is lost because \( R_f \) introduces thermal noise just as \( R_i \) did on Fig. 2.2b. To reduce its effect, \( R_i \) has to be as large as possible. This means slow discharge and possible saturation at high count rate. Big valued resistors are large; their body adds capacitance from input to ground which increases the contribution of amplifier-voltage-noise [10,11]. Big value resistors also higher than the fundamental thermal noise, and have poor frequency behavior that makes it hard to use pole-zero cancellation [12].

To avoid these drawbacks, the integrator can be left to accumulate charge freely until it approaches saturation. Then a relatively large current source can be connected to the input that would abruptly discharge the capacitor, and the integration can begin again. The sudden change at the output, during resets, measurement will be interrupted and can begin again only after all the electronics have recovered.

Well designed pulsed reset preamplifiers have the best noise performance. However, due to the discontinuity of the process and the magnitude of the discharge pulse, there are many difficulties associated with this approach. The MCA design, which is the focus of this work solves many of these, and makes it possible to operate as if the reset was transparent. The problems are summarized and the solutions are presented in Chapter 3.

2.3 Pulse Shaper

In order to perform measurements of charge contained in a detector pulses, the amplified integrated signal goes through analog or digital filtering. The goal is to
suppress noise and convert most of the charge into an easy to measure form. These filters are commonly known as *shapers* because the pulse shape that's produced by the convolution of its impulse response with the detector pulse determines the noise and detector charge collection characteristics. Shapes are usually linear filters, but they are not always time invariant.

The durations of the rising slope and “flat top” of the shaper's step response\(^1\) are parameters that affect the its noise performance and charge collection. It's been show that the optimal response for reducing parallel and series noise in the preamplifier, is the what's called and infinite cusp, which is the following: \( S(t) = e^{-\frac{|t|}{\tau}} \) shown on Fig 2.4a.

\[\text{Fig 2.4a} \quad \text{Non-causal infinite cusp time response of a theoretical optimal filter. Figure adapted from [10].} \]

\[\text{Fig 2.4b} \quad \text{Modified (finite response causal) cusp filter with a flat top for better immunity to pulse shape variations. Figure adapted from [10].} \]

The time constant \( \tau \) determines the contribution of the parallel and series preamplifier noise sources in the filtered signal. The minimum amount of total noise is achieved when the two sources have equal contribution. This result plus thorough discussion about noise and shaper response can be found in [13] and [11]. A finite duration version of the cusp filter has been implemented in a digital processing algorithm [14].

In cases when the detector pulses have some kind of variation in their shape as in proportional counters like \(^3\)He tubes shown in the previous sections, or the dependance of the rise time on temperature in NaI(Tl) scintillator, there must be a period of time when

\(^1\) The preamplifier is an integrator (charge sensitive), which means that to an impulse current at its input, it will produce a step function. This is why the shaper's step response is usually analyzed rather than its impulse response.
the response of the shaper is close to a constant (the so called flat top) in order to allow
full signal convergence in both the fastest and the slowest rise times. If the flat top doesn't
have sufficient duration to collect most of the charge (so called ballistic deficit effect),
pulses with the same amount of total charge but different shape, will have different
heights resulting in degraded resolution of the pulse height distribution. The cusp filter
has on flat top at all, this is why in most cases some variation such as the one shown on
Fig 2.4b has to be implemented. In many detector types, a good rule of thumb is that the
rise time should be less or equal to the flat top.

Practical shaper implementations have different time responses, such as triangular,
trapezoidal, approximations of Gaussian functions, simple RC network responses, etc.
These may have less than optimal noise performance and no real flat top portions in
some, but it's often an acceptable compromise when there is a need for reduced
complexity.

Increasing pulse duration by adding a flat top or optimizing for noise performance
increases the probability of pileup, which happens when two events arrive too close in
time to one another to have independent amplitudes.

The trade off for increasing pulse duration by adding a flat top or optimizing for noise
performance is increase of probability of pileup of pulses that arrive close to one another
in time.

2.4 Digitization and Pulse Height Measurement

Pulse height measurement is performed after the signal has been conditioned by the
shaper and several auxiliary processing operations (outlined in the next section). The
result has to be in digital form since every measurement is added to a histogram of the
pulse height distribution (a.k.a. pulse height spectrum) in a digital memory. Digitization
occurs either after an analog shaper, or directly after the preamplifier (sometimes
following minimal per-filtering) followed by a digital filter, which is the case in most
modern systems. Fig 2.1 shows both signal paths.
Some system use simpler processing because not every application requires optimal noise performance, high rate, or resolution. If the complexity of the filter, the rest of the signal processing, and the sampling requirements could be reduced, then significant reduction in power consumption and cost can be achieved. Part of the work here is focused on minimizing exactly these parameters (complexity and sampling rate) through a novel preamplifier concept and a very simple algorithm.

2.5 Additional Signal Processing Functions

There are some additional components in MCA systems that refine the measurement:

- A peak detector determines which part of the fully shaped pulse to use as a height measurement. This is not always a straightforward process, because fluctuations due to noise add uncertainty in determining where the slope is zero what is the maximum of a pulse.

- Pole zero compensation. This function takes away the very long RC time constant caused by a bleeding resistor, and/or discharge characteristic of a decoupling capacitor. It's a deconvolution process using the inverse transfer function of the exponential RC decay.

- A baseline restorer tries to accomplish more effectively what the pole-zero compensations does. It tries to keep the value of the signal base line despite the count rate and the offsets caused by the preamplifier and subsequent filters. This is accomplished with time varying techniques usually forcing the output to zero when no pulse is present.

- A pileup rejector relies on a fast shaper, not optimized for charge collection or noise. It's purpose is to detect pulses that are close enough in time to overlap in the spectroscopy channel, and to veto these events. It's an indispensable for high rate measurements.

Extended analysis of these functions and how they would be implemented can be found in [10].
3.1 Introduction

The focus of this work is the development of a universal Multichannel Analyzer (MCA) that can be used for X-ray and gamma-ray energy spectroscopy using a NaI(Tl) scintillation detector or $^{3}$He proportional counter neutron detectors. Its design incorporates a new idea for a pulse reset integrator that can emulate continuous integration without interruption due to resets, eliminating many problems in conventional reset integrators and substantially reducing complexity of subsequent electronics. In addition, conventional digital and analog filtering is replaced with a Correlated Double Sampling algorithm (described in section 3.6.2 ) that requires sampling and processing with frequency comparable to the charge collection time of the detector, which is about 100 times lower than conventional digital systems. This dramatically reduces power consumption from the ADC and the digital signal processor.

3.2 Specific Application-Uranium Enrichment Monitoring

The new design was developed under a larger project for uranium enrichment monitoring system because of the MCA's potential to offer several improvements over existing spectroscopy systems, such as low power consumption, lower cost, potential for customization, and others. This project allowed us not only to prove the concept, but also to implement a functioning well performing device with potential for deployment.

Even though the focus of this work is not on the uranium enrichment system, it's worth outlining its purpose and principle of operation for the sake of understanding the types of measurements, instruments, and the motivations for development of the MCA presented here.
3.2.1 Uranium Enrichment Monitoring Overview

A facility that is capable of processing uranium by one way or another altering the portion of the fissile isotope $^{235}$U relative to $^{238}$U is generally under strict scrutiny of national and international atomic regulatory agencies like IAEA (International Atomic Energy Agency). This is due to the fact that material with high concentration of $^{235}$U can be used to build nuclear weapons. Facilities that produce enriched uranium (uranium with concentration of $^{235}$U above the natural 0.7%) for peaceful purposes called Enrichment Plants, are of particular interest for monitoring. The most common type is the gas centrifuge enrichment plant for which the enrichment monitoring system is designed.

Regardless of the production method, verification of the declared activity is crucial for prevention of illicit nuclear material production, and for process control. The latter may seem less important than the need for verification, but is essential for preventing criticality accidents that can occur if the fissile material that's being processed has higher than expected enrichment, causing it to reach critical mass and result in nuclear reaction.

Many methods exist for uranium enrichment measurement. The one that is most accurate and reliable involves collecting a material sample from different stages of production and directly measuring the isotope content with a mass spectrometer. This method has an obvious disadvantage because it does not provide continuous data and because a genuine sample may not be provided to inspectors by unfriendly facilities.

In order to address these shortcomings, our team, is developing a system for uranium enrichment measurement based on transmission of X-Rays through the pipes where uranium hexa-flouride ($\text{UF}_6$) flows. [15].

3.2.2 System Description and Motivation

Although the enrichment monitoring system is very complex, and has several different implementation versions, the method uses to two types of radiation measurements:

Measurement of the energy distribution of the transmitted X-ray photons (X-rays shone at
the pipe that are not absorbed by it and material inside), and measurement of the spectrum of the uranium (characteristic gamma ray energy distribution produced by the uranium radioactivity) inside the pipe. In the two cases the energies are quite different, but the detectors and electronics used are the same. Only the gain that they operate at is adjusted to match the type of radiation.

Fig 3.1 Passive Enrichment Monitoring System. MCAs and detectors are labeled. So are other important components like the shielding and collimator. Figure from [16]

Fig. 3.1 shows the enrichment monitor with NaI(Tl) detectors connected to DigiDart MCAs manufactured by Ortec and used during the system's development. The problems with using off the shelf MCAs and the motivation for adapting our MCA design for this application are summarized in the next section.

3.2.3 Requirements

There are several necessary conditions for the electronics in the Enrichment System that
are dictated mainly by the fact that it has to operate autonomously and be unattended. The goals of our design are summarized below:

- The MCA should be able to collect large number of spectra and store them in non-volatile memory with a time and date stamp for retrieval in case of communication failure.
- Immunity to power outages. Very important requirement, and one that is the hardest to find among commercial MCAs, is power consumption that is low enough to allow battery operation for extended periods. A goal was set to achieve less than 100mW at supply voltage range of 3V (nominal voltage for a Li-Ion battery) up to 15V. That power consumption, for example, would allow 20 day operation on a standard 12Ah 12V UPS battery.
- Diversity of power sources. The wide supply voltage range of the electronics helps when there is a need be able to switch from sources like 12V main, power-over-Ethernet, and different battery types.
- Tampering resistance. This is why compactness is important. If the electronics is included in the body of the scintillation detector, there will be less chance for failure due to unreliable interconnects, cabling, or tempering with separate components of the system. A good implementation in this sense is the device shown on Fig 3.3a. This is an MCA produced by Ortec that includes front end analog electronics (preamplifier and necessary analog per-filters), digital signal processing, communication with a computer (USB or Ethernet option), and a high voltage power supply for the PMT. All of this is contained inside a cylindrical module the size of a normal 14pin PMT base.
- Data and Communication security. There has to be a way for the data and its transfer between the system and the user (for example a facility inspector) to be securable against tampering and eavesdropping. Encryption and transfer protocols may need to be user specific.

A global view of the final version of the system as it is intended to operate and communicate is shown on Fig 3.2.
Fig 3.2 Plan for deployable unattended enrichment monitoring system.

3.2.4 Commercial Alternatives

Many of these requirements are fulfilled separately in some commercial MCAs, but it's hard to find one that satisfies them all. The Ortec DigiBase, for example has a power requirement comparable to a USB port maximum power (<500mW). Despite the low price of flash memory, many MCAs offers only few spectrum buffers (like 4Kbyte single spectrum memory in DigiBase) [17].

The biggest hurdle is proprietary design that doesn’t allow modifications. Fig 3.3a,b,c show some MCAs used in the development of the Enrichment Monitor. The Ortec's DigiDart, and Canberra's Lynx require external preamplifier. The DigiBase is almost suitable, but has high power consumption and not enough memory.
3.4. Detector

Based on the physical requirements of the Enrichment Monitoring System standard NaI(Tl) detectors were deemed suitable.

The dependance on temperature of the light pulse duration and shape of NaI(Tl) material has to be taken into account when designing the new MCA’s signal processing, in particular its charge collection times. Having a PMT amplify the signal means that contribution of noise from the preamplifier will be very small, and optimization of rise time in the shaper is not critical. These considerations are relevant to the choice of signal processing and are explained in more detail later.

We use a PMT that is connected to negative high voltage. As mentioned in the
introduction, when using this type of connection, the charge from the anode is collected directly, unlike a positive high voltage connection where there is a coupling capacitor between the anode and the amplifier. Choosing the latter would have added a complication to the signal processing because, the average current through the capacitor has to be zero, so there is a long discharge current tail following each signal pulse, which introduces offers and signal shape variations.

3.5 Preamplifier

Our preamplifier is a version of a charge sensitive preamplifier with pulsed reset that is transparent to the signal processing part of the electronics. However, this design is an important advancement in that technology because it solves a myriad of problems associated with conventional circuits like this.

3.5.1 Problems with conventional pulse reset

Despite its superior performance, pulsed feedback is considered a higher complexity architecture because of several complications and disadvantages created by the need for an abrupt and full discharge of the integrating capacitor. A traditional reset amplifier was described in sec 2.2.1 and illustrated in Fig. 2.3c.

The average number of pulses after which a reset occurs, is a decision left to the designer. In general, it's a tradeoff between having smaller sensitivity/amplification in the first stage with large number of step pulses per resets, or having a reset after each pulse or only few pulses. The former requires lower noise electronics and more bits on the ADC to measure the smaller pulses. The latter increases the number of resets and the detrimental effects due to them. Each reset injects charge in the feedback capacitor $C_f$ (Fig. 2.3c) equal to all the signal charge accumulated since the last reset. The time that a reset takes has to be minimized if dead time is important. As a result, the pulse that occurs at the output of the preamplifier typically last several micro seconds, has polarity opposite to detector pulses, and amplitude that is much larger than an average detector pulse[18].
Several consequences occur due to the intensity, speed, and polarity of a reset. Most system perform digital and/or analog linear signal processing after the preamplifier stage. Even though the preamplifier is kept in linear operation except for possible “slewing” during the retest, such large signal pulse often saturates the next stages causing non-linear operation and dead time due a period of recovery of the electronics. Beside that, there is also an issue with long tails of large signal pulses. Most shapes have an infinite time response (with the exception of the ones implemented exclusively with analog delay line, or FIR digital filters). The duration of a pulse is determined by how long its amplitude is above a certain threshold. When signal with normal amplitude arrives, the response of the shaper becomes negligible after a certain period. However, when a reset occurs with an amplitude that could be many times larger than an average pulse, the the duration of the filtered pulse becomes significant, and measurement of consecutive pulses is affected.

Every pulse that occurs during a reset is lost in the dead time of the system. This however, is not the only problem. As explained above, the entire signal processing chain following the preamplifier is disabled due to overload, or because parts are intentionally stopped to prevent overload. A detector pulse that comes during this time is not registered creating a so called dead time (DT). Not only will this pulse not be counted, but it also will have effect on the following pulse particularly if it has significant slow signal component (as is the case for NaI(Tl) scintillators at low temperature).

In addition, the noise history before the reset will be lost. This means that the shaper, which is a causal linear filter that has states with memory of past noise, will have all these reset, and any noise improvement due to that memory will be lost.

When a reset occurs immediately after a threshold is passed, then the subsequent electronics must ignore the event that caused the overflow because a discharge of the feedback capacitor is initiated that doesn't allow measurement of the charge pulse. If this loss occurred with equal probability for every pulse, then it simply would contribute

---

2 Slewing in amplifiers occurs when the input changes faster than the device can drive its output resulting in non-linear performance during that condition.
equally to the dead time throughout the entire energy spectrum. Unfortunately the probability that a pulse will go above the threshold increases with its amplitude (proportional to energy). This leads to significant efficiency loss for high energy interactions, and spectrum distortion due to energy dependent dead time losses. Landis, Madden, and Goulding describe a solution to this problem in [19]: If after the reset threshold is crossed, a “wait time” is added enough for the signal processing to take place, then the reset will not cause loss of the pulse that triggered it. One caveat is that the range of linear operation of the preamp has to be extended above the reset threshold by at least the maximum value of measured pulse.

In our preamplifier design, which will be presented next, we have addressed many of these issues in a very elegant and low-complexity manner. More importantly, we are presenting a solution for the discontinuous nature of the reset process.

3.5.2 Free Running Integrator-Idea and Implementation

It is possible to emulate continuous integration with a pulse reset integrator if several conditions are met: A reset can be ignored successfully if, when finished, the preamplifier output returns to known (zero) value. The preamp has to be followed by an ADC and a digital filter that can completely disregard the transition and assume that the zero-value is just a continuation of the signal. Any analog time invariant per-filtering that settles slower than the sampling period of the ADC, will cause the reset transition to appear as signal. The return to zero-value has to begin immediately after an ADC sample and be complete before the next sample. Any actual signal charge that is produced by the detector during the transition has to be conserved and available at the output (as a voltage above the zero-value) for the ADC to sample. This includes actual pulses that occurred near the time of a resets, leakage current, and most often-long tails from high energy gamma-rays that shift the base line for a significant amount of time.

Signal processing in conventional systems remains discontinuous because some of the problems mentioned above are still unsolved. Charge accumulated during reset duration is lost because current is injected at the input until the lower discriminator value is
reached regardless of what kind of signal occurs during this time. Low noise techniques have to be implemented if the output is to return to a zero-value that doesn't fluctuate. This is not trivial because the reset current source is many orders of magnitude stronger than the detector leakage current [18], from which follows that the shot noise spectral density \( S(f) = 2q I \) will also be proportionally greater. This, and the fact that comparator threshold fluctuation has to be taken into account, means that the value that the output is reset to will be very “noisy”.

The idea for what we have named a Free Running Integrator (a charge integrating amplifier that allows continuous signal processing despite preamp resets) is based on continuously tracking of the preamp output during normal integration, and injecting the exact amount of charge necessary to return it to its zero-value during reset. This is achieved with a tracking capacitor matched in value with the integrating capacitor at preamp feedback. Fig. 3.4 shows the circuit diagram of the free-running-integrator used in our MCA system.

![Fig 3.4 Conceptual drawing of the “Free Running Integrator” idea. The signal charge is collected by two equal capacitors: \( C_f \) and \( C_s \). When \( C_s \) is reversed by the switches, the charge sharing process results in an output voltage equal to zero.](image)

The tracking capacitor \( C_s \) is connected in parallel with the feedback capacitor \( C_f \). Since their capacitance is matched, both contain the same amount of charge. When the reset logic triggers, the switches **SW1** and **SW2** reverse the polarity of the tracking
capacitor. This leads to charge sharing process between two equally charged capacitors with reverse polarity. When the process is finished, the charge in both will be zero without the need of a comparator to stop the discharge process. Assuming the switches have negligible leakage, no other charge is gain or lost except any signal charge generated by the detector. This is a very important advantage of our design because it allows signal to be collected continuously without loss, unlike any standard reset strategy. Figures 5.5a and b show events that occurred during a reset being successfully captured. This feature will be described in detail in section Chapter 5, which presents experimental results and performance of our device.

The speed of the charge sharing is limited only by the resistance of the switches and the step response of the amplifier without loss of signal charge. In this version of the design, the discharge path is closed between the two capacitors and the switches. This relaxes the slew rate requirements of the op-amp because it only has to charge/ discharge its internal capacitance. As long as the overall response is fast enough for the system to settle to a desirable value before the ADC samples, it doesn’t matter whether the switches have non-linear resistance, if the op-amp response is dominated by slewing, or if it has some ringing due to the transition.

The full noise advantage of a pulse reset amplifier is realized in this architecture because there is no current source or resistor delivering charge at the input of the preamp. The capacitor is a noiseless component which means that after system has settled, no fluctuations are added of the zero-value due to thermal or shot noise. As far as noise history conservation is concerned, the reset neither adds nor subtracts from the amplifier and detector noise; the ADC will see all of it before and after the reset.

The features listed above can be used to eliminate dead time due to resets and allow virtual uninterrupted charge integration. Further more, this is a very low complexity circuit. Resets can be initiated by the digital signal processor at the most opportune time, which is a short time after a pulse has crossed the threshold. The wait period is not necessary because charge won't be lost even during the rise time of the pulse, but a short
wait would reduce error due to nonidealities. Because the pulse which triggers a reset is not lost, energy dependent dead time losses are eliminated.

This is a switched capacitor circuit, and as such, it has some shortcomings associated with that technology. Switched capacitor circuits were made practical with the evolution of MOSFET transistors. Thanks to the fully insulated gate, no charge from the control signal is transferred during switching except via the parasitic capacitors. The channel has resistive properties when the device is on (there are no junction potentials, or significant leakage like there is through the base of a BJT). These properties allow lossless transfer and measurement of charge, which has enabled the implementation of excellent sample and hold circuits, clocked analog filters, charge buffers, Analog to Digital Convertors (ADCs), Digital to Analog Converters (DACs), and many other circuits. Switched capacitors are best implemented in integrated circuits, because the technology allows very precise matching of transistor and capacitor parameters. The precision of these circuits hinges on matching. Good overview of switched capacitor design can be found in [19].

Unfortunately our discrete component implementation can't take advantage of all that integrated circuit technology can offer. Even though this is a very mature technology, and many of its problems have been overcome, we had to deal with them in more conventional manner. Two issues in particular, had to be resolved:

- Mismatch between $C_f$ and $C_s$. This leads to the reset not returning the output to zero. Because the amount of charge in mismatched capacitors that are charged to the same voltage is not equal,

$$V_f = \frac{V_i(C_f - C_s)}{(C_f + C_s)} \approx \frac{V_i C_{err}}{2}$$

Where $V_f$ is the final voltage, $V_i$ is the initial voltage at the output before the reset occurred, and $C_{err}$ is the relative mismatch of $C_f$ and $C_s$. If the mismatch is 2%, the residual value after reset will be about 1% of the output before the reset. It's good that at least this dependance is linear, which is not the case in the next problem.
• Charge Injection from the switch. We used a commercially available analog switch ADG633 manufactured by Analog Devices Inc. It has an average charge injection of 1pC. This results in couple percent shift in the zero-value after reset ($C_f$ and $C_s$ were in the order of 20pF). Furthermore, the actual value depends on the amplitude at the output before the reset begins. And unlike the capacitor mismatch problem, the dependence is not linear [20].

These problems were resolved using calibration. The MCA had a self-calibrating procedure that could be run periodically. It will be described in detail in Chapter 5.

Careful design decisions had to be made in several occasions in order to avoid pitfalls particularly charge leakage due to poor component choice. For example, the op-amp had to have very little input leakage current in order to avoid excessive positive or negative ramp (there is a natural ramp at the output due to integration of detector dark current). Op-amp with MOSFET inputs was chosen (ADG8615 from Analog Devices) because of its negligible input current (1pA) [21]. The second component that could introduce charge loss is the switch. We chose ADG633, a MOSFET switch that has the so called “break before make” property. This means that the capacitors' terminals will be completely disconnected by the switches before reconnecting them with opposite polarity. The reason for this is to ensure that the capacitors won’t discharge due to short connection at any time in the process.

3.6 Multichannel Analyzer / Digital signal processing

3.6.1 Introduction

Besides having straightforward, low complexity design, the preamplifier described in the previous section provides continuously integrated charge signal, which allows a simple and effective signal processing strategy. As it will be described soon, one of its merits is operation at low sampling rate and minimal number of arithmetic operations needed per sample. This is important because in an MCA systems, the highest contribution to power
consumption comes from the signal processing and analog to digital conversion. The DSP portion of the power depends heavily on the complexity and optimization of the algorithm. Generally, as long as the specifications are fulfilled, reducing complexity will reduce cost and power consumption. Also, higher than necessary sampling rate will require proportionally higher number of digital operations respectively increasing the DSP power.

While power consumption increases linearly with frequency in the digital signal processing, it increases quadratically in the ADC electronics. Fig. 3.5 shows couple of data points for different modern ADC chips produced by National Semiconductor Inc.

<table>
<thead>
<tr>
<th>ADC Type</th>
<th>Number of Bits</th>
<th>Speed (Msps)</th>
<th>Pwr. (mW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>ADC1173</td>
<td>8</td>
<td>15</td>
<td>36</td>
</tr>
<tr>
<td>ADC1175</td>
<td>8</td>
<td>20</td>
<td>60</td>
</tr>
<tr>
<td>ADC1175-50</td>
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<td>125</td>
</tr>
<tr>
<td>ADC10321</td>
<td>10</td>
<td>20</td>
<td>98</td>
</tr>
<tr>
<td>ADC10030</td>
<td>10</td>
<td>30</td>
<td>125</td>
</tr>
<tr>
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<td>1.3</td>
</tr>
<tr>
<td>ADC121S051</td>
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<td>1.6</td>
</tr>
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</tr>
<tr>
<td>ADC12040</td>
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<td>340</td>
</tr>
</tbody>
</table>

Fig 3.5 Table of several ADC parameters. Note how sampling rate and number of bits affect the power consumption. Source of data: National Semiconductor Inc.

Commercial MCAs operate at sampling rates between 10 and 100 mega samples per second (Msps), and have FPGAs to perform digital filtering.

3.6.2 Charge Pulse Processing

The preamplifier described in the previous section is a pure integrator. There is no need for compensation of decay constants, or pole-zero cancellation. Fig. 3.6 illustrates the algorithm for pulse height measurement of an integrated detector pulse.
Fig 3.6 Illustration of pulse height measurement algorithm of an integrated charge pulse using Correlated Double Sampling method.

- The DSP continuously subtracts the previous sample from the current in order to detect a rise in the output voltage above what's caused by noise or positive slope due to the PMT's dark current. If $V_{S2} - V_{S1} > V_{LLD}$ then a pulse was detected and it would have to be processed further. $V_{LLD}$ refers to a Low Level Discriminator value, which is the minimum rise that would cause a pulse detection. Alternatively, if $V_{S2} - V_{S1} < V_{LLD}$, no further processing is needed, and the algorithm goes on as normal with $S2$ taking the role of $S1$. $S2 \rightarrow S1$.

- If a rising edge is detected, the DSP waits for the next sample, $S3$. Then, a pulse height calculation is performed as follows: $V_E = V_{S3} - V_{S1}$, where $V_E$ is the pulse height, which is proportional to the detected energy.

- After the calculation is performed, the controller increments the value in the corresponding channel of the energy spectrum histogram and returns to the starting point in the algorithm (rising edge detection). Now $S3$ takes the place of $S1$. $S3 \rightarrow S1$.

This type of measurement is not new. It's called correlated double sampling. The noise performance is determined by the rise time of the integrator's step response [10], similar
to Gated Integrator time variant shaping. Also, the charge collection time is independently controlled by the integration period [22], which in our case is the sampling period of the ADC.

The processing algorithm on average requires only one subtraction operation per ADC cycle. This is probably the lowest digital signal processing complexity achievable. In addition, our strategy eliminates the need for several complicated functions that are required in standard MCAs: These are Base Line Restorer (BLR), shaper (even though it's implicit in the way we perform the measurement), peak detector, and various pulse shape compensations such as pole-zero cancellation and deconvolution of RC response of the preamp. Combined with low sampling frequency, it is possible to implement the entire DSP on a single micro-controller with a built in 12 bit ADC that samples at maximum of 200kHz (the micro-controller we used is C8051F140 produced by Silicon Labs Inc.). This allowed us to achieve less than 100mW power consumption for the entire MCA (including the analog front end, and communication with a computer).

Since the pulse can arrive at any time between samples $S_1$ and $S_2$, it means that at one extreme it could begin right after sample $S_1$, and will have two full periods to reach its maximum. In the other extreme, the pulse can start to rise right before sample $S_2$. Then it will have only one ADC period to converge. The maximum difference between pulses with the same amplitude that occur at these two extremes depends on how close they converge in one ADC period. For example, if a pulse reaches 99% of its maximum in $T_{ADC}$ and close to 100% in $2 \times T_{ADC}$, then the discrepancy between the two extremes will be 1%. This is acceptable for NaI(Tl) scintillators because intrinsic resolution (FWHM) for 662keV photons is more than 7%, and worse at lower energies. However 1% degradation may prove too much for better detector materials.

This is why it's important for the sampling rate to be long enough to allow close to full charge collection. The signal processing method and the charge integrating amplifier allow full control over the time for collection. In the NaI(Tl) scintillator, the light decays roughly exponentially with dominant time constant of 230ns [1]. There is another, $\sim 2.2\mu$s
long time constant that is negligible at room temperature, but was measured to contribute about 40% of the light at -20°C. This slow component was shown to deviate about 10% from true exponential decay [4]. Fig. 3.7 shows this significant temperature dependence and pulse convergence rates.

![Fig 3.7](image)

**Fig 3.7** Temperature dependence of pulse shape of a NaI(Tl) scintillator. The upper graph shows the integrated charge. It can be used to estimate how long it takes to converge to an acceptable value. Figure from [4]

If the detector current is approximated as a sum of two decaying exponents, then the normalized charge pulse can be described by the following expression:

$$Q_{\text{norm}}(t) \approx 1 - A_1 e^{-\frac{t}{\tau_1}} - A_2 e^{-\frac{t}{\tau_2}}$$

Where $A_1$ and $A_2$ are the proportion of the charge contributed by the corresponding exponential components. $A_1 + A_2$ must add to 1 in the above expression. This expression can be used to estimate the amount of charge collected after certain integration time.

In our design we chose ADC sampling period of 5μs (200kHz frequency). At room
temperature, the fast decay constant \( \tau_1 = 230 \text{ns} \) dominates. After 5\( \mu \text{s} \) less than \( 10^{-9} \) of the pulse remains. Practically all the light is collected. This is not the case at -20°C when after the same period, where the slow component contributes 40% of the charge, and dominates the time response after few micro seconds.

\[
Q_t \approx 1 - 0.4 e^{\left(\frac{t}{\tau_2}\right)}
\]

where \( \tau_2 = 2.2 \mu \text{s} \). The error then is about 4%. Convergence for other temperatures can be estimated from Fig. 3.7. If the detector has to be operated at temperature below 0°C, the sampling period will have to be increased, or another solution implemented. A very interesting solution is demonstrated in [4,5,6] and is discussed in section 6.2. It is based on a method of compensation that could cancel an exponential rise time and make it approach a step function.

Fortunately, the ADC frequency in our design is easily modifiable. All our measurements are done at 200kHz because it's anticipated that the operating conditions will be close to room temperature, but this is not necessary, and the method is valid for lower sampling rates as well.

As in all spectroscopy systems, a longer duration of the energy pulse means higher probability of pileups and longer dead time. Just as in our signal processing strategy, longer impulse response of the shaper is justified when there is a need to collect the charge from a slowly rising charge pulse. This is particularly important when the signal rise time varies (as it does with temperature in NaI(Tl) scintillators). One benefit that comes out of the necessary reduction in sampling rate is reduced power consumption.

### 3.6.3 Reset Processing

In order to complete the MCA's signal processing mechanism, a strategy for how the device will deal with integrator resets has to be chosen. First, as was presented in detail earlier, the novel reset mechanism allows the emulation of continuously integrated input signal. The algorithm has to take advantage of the benefits it offers, namely return to a known value, conservation of signal charge during reset, and settling time that is shorter
than the sampling period.

Let's first take the case where the reset is necessary because a pulse pushed the integrator above its threshold. All resets are performed when the micro-controller makes the decision to do so. Fig. 3.8 shows one possible way to perform a reset and do a pulse measurement.

At the time of ADC sample $S_2$, the processor sees that the threshold was exceed, and initiates a reset. The signal charge is conserved during the event, and at the time of $S_3$ it can be measured.

Besides the need for reset at $S_2$, the processor also detects a rising edge of the pulse, which means that a pulse height measurement is needed. Unlike the normal procedure where $V_E = V_{S3} - V_{S1}$, now we have to take into account the amount of signal collected after the reset had canceled all the previously integrated charge. Because we can rely on the fact that the preamp voltage will return to a known value, and everything above it is due to collected signal charge, we can say that the pulse height is...
\[ V_E = (V_{S2} - V_{S1}) + (V_{S3} - V_{RST}) = \Delta V_1 + \Delta V_2 \]
where \( V_{RST} \) is the known reset level.

The algorithm returns to its original state the same way as before, by making \( S3 \rightarrow S1 \) as the integration begins anew.

There is one caveat that complicates processing a bit. After a sample is taken by the ADC, any charge integrated on the feedback capacitor before the reset begins will not be taken into account because the reset will bring everything to zero. This is why the ADC sample and hold operation must occur almost at the same time as as the reset. In this case, it's not a good idea for the processor to be doing calculations (and to wait for the ADC to converge) so it could determine whether the threshold was crossed or not. Therefore, the processor will have to rely on the previous measurement, then take a sample and immediately do a reset (in our case we signal the ADC to sample, and at the next operation 20ns later send a RST flag). The process now looks like Fig. 3.9.

**Fig 3.9** Improved Sampling and measurement strategy during a reset (explained in the text).

In this case S2 still is the threshold overrun determination point, but this time the processor initiates the reset on the next ADC sample, S3. The pulse height measurement
is done as in a non-reset case $V_E = \Delta V_1$. However, to return the algorithm to initial state, instead of making $S1$ be $S3$, we have to take into account the reset value. The processor begins new rising edge detection cycle, but instead of using the value measured at $S3$ as a first measurement, it uses $V_{RST}$. The next determination whether a pulse began is made based on $V_{TRANSIENT} = V_{RST} - V_{S4} = \Delta V_2$. Since most times threshold overruns are due to rising edges of pulses, this strategy has the added benefit of not resetting in the middle of a pulse, which reduces energy dependent resolution deterioration due to reset nonidealities. This is similar to what Landis, Madden, and Goulding did in [18] to address energy dependent dead time losses, except here we don’t have dead time only possible error in measurement.
Chapter 4

Device Construction

4.1 Design Overview

Several experimental prototypes were built to test and refine the analog and digital aspects of the Free Running Integrator's reset strategy, and signal processing. Some initial designs included reset controls that were independent from the DSP. The design eventually evolved towards the use of a family of Micro Controller Units (MCUs) produced by Silicon Labs Inc. that included a 8051 family processor and an ADC on a single chip. With some code optimization we were able to perform all functions in a 50MhZ device that had a 12bit ADC, with maximum sampling frequency of 200kHz and several other useful peripheries. The analog portion consists of the charge sensitive preamplifier with analog CMOS switch (AD633). One of the switches on the same IC also multiplexes the input of the preamp between detector signal, and a ramp generator (simple current source) that is used for self calibration.

4.2 Free Running Integrator Electronic Design

Fig. 4.1 shows the implementation of the Free Running Integrator with lossless reset that was described in the previous chapter. This schematic shows physical circuit that implements Fig. 3.4 with some additional features like a calibration current source.
The op-amp (AD8615) was chosen to have MOSFET inputs that have extremely low leakage current (1pA), relatively low voltage noise (8 nV/$\sqrt{\text{Hz}}$), good bandwidth and slew rate (20MHz and 12V/$\mu\text{s}$ respectively), operating supply voltage down to 2.5V, and very low supply current for that performance (2mA). These parameters fulfill the free-running integrator requirements (low input leakage into the integrating capacitor, and speed for adequate settling time), and the enrichment monitoring system requirements for high power efficiency, voltage that can be supplied by a single Li-Ion battery, and. Noise is less important for scintillation detectors, but AD8615's good noise characteristics became useful in experiments performed with $^3$He neutron detectors.

The entire circuit is supplied by positive single polarity voltage (from ground to VA). All components (including digital) are capable of operating down to 3V. Even though, the input voltage range of the op-amp includes 0V, we have created a positive offset via the resistive divider R21 and R22, in order to avoid any degradation when operating at the minimum input range. This also allowed a simple calibration current source to be implemented (discussed below).
The ADG633 integrated circuit contains three analog CMOS switches. Two of them, the first connected to pins 1, 2, 15, and the second to pins 3, 4, and 5, make up the reset circuit described in Chapter 3 and shown on Fig. 3.4. The switch has the following characteristics [20]:

- Minimum single supply voltage of 3V (with guaranteed and fully specified performance).
- 0.2nA leakage current.
- 52 Ohm on resistance over full signal range.
- Typical charge injection 1pC at 3V supply.
- Power consumption <0.1μW
- The switch operates in “break before make” mode, which means that it breaks the connection on one side before connecting the other side of the switch. This is necessary to ensure that the capacitors won’t discharge due to short connection at any time during the switching process.

The third switch is connected to pins 12, 13, and 14. It's used to multiplex the input of the reset integrator between the detector, and a calibration current source. For the MCA's self-calibration procedure the micro-controller needs a gradual positive ramp at the integrator output. The exact slope of the ramp is not important as long as it doesn't contribute significant charge during a reset. The MCU uses that signal to ramp up to different integrator voltages then measure and record the corresponding reset levels. The calibration source is simply a large resistor connected between the input and ground. Since the op-amp input is at the slight positive offset potential, the resistor provides negative leakage current that corresponds to positive ramp at the integrator because the amplifier is inverting.

As it is drawn on the schematic, if the supply voltage is 3V, the offset potential will be around 100mV. The calibration resistor is 100 MΩ, which means that the calibration current will be around 10nA. This corresponds to a slope at the integrator output of
\[
\frac{dV_{\text{INT}}}{dt} = \frac{I_{\text{CAL}}}{C_F + C_S} = \frac{1\text{nA}}{20\text{pF}} = 50\mu V/\mu s.
\]

If a reset last around 60ns (ADG633 switching time [20]), the error due to charge deposited during a reset will be only ~3\(\mu\)V.

In normal spectroscopy mode, the current source is disconnected from the input, and the detector is connected.

### 4.3 Digital Signal Processor Hardware Design

Fig. 4.2 shows the schematic diagram of the micro-controller with its connections to different peripheries, communication ports, and the analog switch for controlling integrator resets and calibration.

![Digital signal processor schematic](image)

**Fig 4.2** Digital signal processor and control unit for the MCA.

The C8051F410 MCU has the following features relevant to our design:

- Built-in12-Bit ADC with programmable frequency up to 200kHz. It's used to
sample directly the output of the free-running integrator. It's input is at pin 16 on the MCU chip, configured as an analog input.

- Up to 50MHz system clock with 70% of instructions complete in one clock cycle.
- Serial communication ports that support several standards: UART that we use to communicate with a computer, SPI port that we use to control several peripheries on the MCA circuit boards, and an SMBus (that supports I2C protocol).
- General purpose register-addressable ports that are used for the following functions:
  - Controlling the switch that resets the free-running integrator(port P1.4, pin 14).
  - Controlling the switch that connects the preamplifier input to either the detector, or the calibration current source when the self-calibration routine is running.
  - Pins 25-32 are general purpose ports used for debugging purposes, or to communicate with an optional second MCU on the same board.
- There are several Timer/Counter modules that are used in the software as periodic interrupt sources that signal the ADC to take samples, and for other purposes.
- Various types of memory used for code storage, working memory, and spectrum data storage and buffering.

These and other features of the micro-controller are summarized on Fig. 4.3 taken from [23].
The primary MCU performs all signal processing functions, controls integrator resets, runs the self-calibration routine, and communicates with peripheral devices including a computer via RS232 serial port.

4.4 Additional Peripheries

Additional components were included in the final design in order to make the MCA useful for the uranium enrichment monitoring system, in other words to be a power efficient, customizable, and autonomous. These components are as follows:

- Switching voltage regulator with 96% efficiency to stabilize supply voltage for the electronics, and to maximize battery power utilization. The integrated circuit used is TPS62122 by Texas Instruments Inc.
- Two Digital to Analog Convertors (DACs) that are controlled by the MCU. They are there to provide variable reference voltage for controlling a high voltage power supply module. The data transfer between the DACs and the MCU is done via an SPI serial buss. IC used is AD7390 by Analog Devices Inc.
- An RS232 serial port line driver (MAX3232 by MAXIM Inc.). This is a convertor from low-voltage UART serial bus (around 2V driven by the MCU) to standard
RS232 voltage levels. This driver is used for connecting the MCA to a computer with serial port.

- 2x1 Mbit non-volatile FRAM chips (FM25V10). FRAM is a type of memory that uses ferromagnetic properties of materials to store data. It has better radiation immunity than flash memory, but we chose this type mainly because of low power requirements, and the fact that each IC has only 8 pins, four of them used for read and write operations via SPI interface.

- An external real time clock (RTC) to keep date and time information even if the main power and auxiliary battery power are not available. It needs a small backup battery. Its power consumption while running is only 100nA. Circuit used PFC2123TS by NPX.

- On the board, there is space for a second MCU of the same type (C8051F410). It was put there to relieve the main one in operations specific to communication protocols, security (encryption etc.), control of high voltage, gain stabilization, etc. It can communicate with the primary MCU via the I2C bus and general purpose ports P2.0 through P2.6. There was no need to mount and use this microcontroller at this phase of development.

Fig. 4.4a,b,c, and d shows the circuit diagrams of these peripheries and via which ports they connect to the primary MCU.
4.5 Physical Detector System and Testbed

The design was fit on three round 2 inch diameter printed circuit boards (PCBs) that stack vertically. The bottom board is very sparsely populated because it contains only the preamplifier and the analog reset switch. This was done intentionally in order to separate the “noisy” digital components from the high gain front-end analog electronics. The layout of the bottom (analog) and the middle boards was done in such a way that when they are connected, only their ground planes face each other. This is an additional measure to shield the sensitive amplifier from digital signals.

The middle board contains the primary micro-controller, the secondary micro-controller, the two FRAM non-volatile memory chips, and the Real Time Clock. Even though the
micro-controller contains an analog part (the ADC), it does not have to be shielded from digital components as carefully, because it receives an amplified analog signal from the bottom board.

The top board contains the serial port driver, the two DAC chips, and the power supply (switched voltage regulator). Fig. 4.5b show the three MCA boards unconnected, in the following order: left most is the top board, middle on the picture also corresponds to the middle in the assembly, and the right one (which is attached to an actual detector) is the bottom analog board. Only a ground plane and few passive components can be seen on the analog board because all sensitive components are on the other side shielded by the ground plane. The cable coming out of the top board is a RS232 communication cable that connects to a serial port of a computer. The cable that comes out of the middle (signal processing) board is connected to the amplified integrator signal that the ADC sees. We put it for diagnostic purposes in order to see detector pulse shapes and to evaluate integrator reset performance.
Fig. 4.5a shows the assembled MCA, connected to a NaI(Tl) scintillation detector. The detector has a 2 inch diameter, ½ inch thick NaI(Tl) crystal (housed at the bottom) attached to a Photo Multiplier Tube (PMT); that's the several-inches long tube that extends up to the electronics section. The shape and size of the MCA boards was chosen to fit the PMT base. There are many sizes of NaI(Tl) crystals that fit different sizes of PMTs, but most of these detectors have a 2 inch diameter base of the PMT. The cables visible on this figure are: a grounding copper strip, RS232 serial cable, analog diagnostics cable, and a high voltage cable that supplies the the PMT HV-divider. Our design is compact and fits a wide range of detectors.
Chapter 5  
Experimental Results and Performance

The MCA's operating behavior was first examined. The first step was to establish that communication with the computer worked and that the device could be controlled in terms of starting, stopping data transfer, switching between modes (like calibration mode, measurement mode, debug, reset-test mode, etc.). Then the detector was put in normal operation mode measuring gamma radiation from a radioactive source as shown on Fig 5.1.

![Experimental setup for MCA electronics evaluation.](image)

**Fig 5.1** Experimental setup for MCA electronics evaluation. In addition to collecting spectra, we observed the analog signal coming for the Free Running Integrator during normal operation and resets.

In this mode, the preamplifier's analog output signal was examined for noise and response to detector pulses. Fig. 5.2 shows the amplified integrated signal produced by the NaI(Tl) detector shown on Fig 4.5. There is no excessive noise, and the rise time looks typical for this scintillator at room temperature. The system performed measurements, resets, and self-calibration successfully.
Fig 5.2 Oscilloscope picture of the output of the integrator showing a pulse from the NaI(Tl) detector. This was used to make sure there was no noise or interference, and to examine the rise time of detector pulses.

As was mentioned before, the power consumption at 3V supply is 18-19mA, which comes to less than 60mW (there are ways, discussed in the next chapter, to reduce this further).

5.1 Spectroscopy Measurements

With the setup on Fig. 5.2 we took spectra of Cesium 137, which is a typical calibration radioactive source. The resolution of a detector measuring the 662keV gamma energy line produced by $^{137}$Cs source, is often cited and used for detector performance measure. The spectrum from out detector is shown on Fig. 5.3a and b.
Our system had resolution (FWHM) of 7.2% at 1000 counts per second which is the same as measured with a DigiDart MCA with the same NaI(Tl) detector. However, on the figure it can be seen how for high count rate the spectrum deteriorates, and the resolution goes to 7.4%. This is because of pileups, and the lack of pileup rejector (a.k.a. PUR). Even though we don’t need a PUR for the enrichment monitoring, or to test the concept, it may prove to be useful for feature application. This possibility is discussed in Chapter 6.

We also connected our MCA to a detector with a large plastic scintillator (4”X4”X4” cube). Plastic scintillators are widely used in radiation portal monitors where the material’s low cost allows fabrication of large scintillator volume. In these detectors, resolution is not as important as in spectroscopy, because portal monitor detectors only measure changes in gross background radiation intensity and trigger alarms when there is an increase. They only need sufficient resolution to establish a counting threshold. Fig. 5.4 shows the spectrum collected with the plastic scintillation detector.
Fig 5.4 Spectrum taken with a 4 inch cube plastic scintillator with $^{137}$Cs source. This measurement is intended to show the MCA applicability as a portal monitor system.

This measurement was done not so much to show the MCA's performance, as to demonstrate its capability as a portal monitoring device. That's not an application that should be ignored because there are very large number of portal monitors and there is a need for low cost robust electronics.

5.2 Reset Performance

The most prominent feature that distinguishes the system described in this work from other spectroscopy systems is the innovative reset mechanism. The MCA that was built serves mainly to demonstrate the principle of a charge sensitive integrating preamplifier with transparent pulsed reset that allows an emulated continuous integration (what we call Free Running Integrator). As described in the previous sections, the Free Running Integrator performs resets without signal charge loss or added noise. This allows the digital signal processor to disregard the large transient and to process pulses that occur during a reset, without interruption.

The experimental setup described in the previous section was used for evaluation of performance during this test. The device was put in a reset-test mode, where we could change the time of sampling relative to the resets, and to observe reset waveforms. Fig. 5.5 shows actual preamplifier output voltages collected during resets, overlaid using a digital oscilloscope’s persistence mode. Clearly visible are examples of detector pulses that occurred at different phases of the resets.
As we pointed out before, because this device was not build using ASIC technology, there are many problems associated with the switched capacitors circuit that couldn’t be addressed the best way possible. Therefore, a calibration routine for reset level compensation had to be added. It's visible on Fig 5.5 for example that two discrete reset levels exist, corresponding to the two discrete satiates of the switches. This is most likely due to mismatch in charge-injection, because the analog switches (SW1 and SW2 on Fig. 3.4) are not absolutely symmetrical, and don't operate in identical conditions.

The self-calibration procedure performs resets at different integrator output voltages sweeping the dynamic range of the ADC, and recording the corresponding values that the integrator output returns to. This procedure compensates offsets due to Q-injection, feedback and reset capacitance mismatch, asymmetric switching (difference in reset levels for the two discrete states of the analog switches as illustrated on Fig. 5.5), and offset voltages in the op-amp. Fig. 5.6 shows the values of reset levels at different amplitudes measured during calibration.
Fig 5.6 Calibration curve produced by the MCU as it measured reset-values for different thresholds thus allowing compensation of errors due to resets occurring at different amplitudes.

By design, this MCA doesn't have dead time caused by integrator resets. However, this by itself is not enough to claim improvement over the conventional technology. Besides the eliminating the long recovery times and overloading of signal processing electronics, the new principle should not contribute to a significant degradation of the spectrum. To evaluate the performance of the Free Running Integrator, a $^{137}$Cs (with photo peak at 662 keV) spectrum was collected consisting of detector pulses that occurred only during resets. This was done by modifying the MCU firmware to save only events that registered as steps above the zero-level following a reset. The spectrum is shown on Fig. 5.7.

Fig 5.7 Spectrum taken with a $^{137}$Cs source and the physical detector shown on Figure 4.5. This spectrum consists of only events that happened during a reset.
It has low statistics, because pulses occurring during resets are rare at the intensity with which we performed the experiment (1000 counts per second in normal detection mode). We had to take the measurement at this low count rate in order to reduce the influence of pulse pileups and make a meaningful comparison with the spectra collected in normal operation.

It can be seen, that first of all, the events are very rare relative to the overall count rate (0.1cps, 10,000 times less than normal operation). This means that errors introduced because of resets wont have very significant weight. Second, the reset-only spectrum has a 662 keV resolution of 7.6% compared with 7.2% at normal operation. This shows that the lossless reset mechanism works fairly well. This is an important result because our system has a lot of room for improvement, that can make the underlying principle applicable for high resolution spectroscopy with low primary signal detectors such as High Purity Germanium that are much more sensitive to noise and charge injected. The third result is that energy dependent losses are not present showing the $^{137}$Cs peak not to be a smaller portion than the rest of the spectrum. This was expected from the principle of operation.
Chapter 6

Future Work

6.1 Technology Improvements

Some obvious improvements were already mentioned. In particular, taking advantage of techniques in modern ADC integrated circuit technology, which has relied on switched capacitors for many years. Some of the toughest problems have been solved to a remarkable extent, which has allowed precision to 24 bits or more (this corresponds to less than a micro volt resolution) [24]. To achieve that, the exact problems we are facing, namely mismatch in sampling/integrating capacitors and charge injection from the switches, were practically eliminated by ADC designers. This is mainly done using differential circuits and capacitor structures that are highly symmetrical. That kind of symmetry can only be achieved with integrated circuit fabrication technology. This is why, in order to maximize the potential performance of the switched reset integrator, it should be redesign and implemented as an Application Specific Integrated Circuit (ASIC).

Substantial reduction in power consumption is possible if the signal processing algorithm is offloaded to a low power FPGA instead of a 50MHz micro controller. The algorithm has to perform on average about one subtraction per ADC cycle. In our implementation this happens every 5μs. There is no need to have a 200 times faster processor operating all the time. There are some ultra-low power FPGAs developed for customization in mobile phones, that consume couple of miliamps at fool processing power at 300kHz [25].

6.2 Signal Processing Improvements

In our application the intensity of the radiation is fairly low. However, the main limitation for the general use of this MCA is degradation of performance at high count-rates due to
pulse pileups. This is due to two things:

- Detector pulse sampling that is too slow. Because we are trying to collect most of the charge from a slow rising detector signal, we are limited to sampling frequency of 200kHz. Since radiation pulses arrive at random times, there is a probability that two pulses will occur in a single sampling period. This probability increases with average frequency of events.

- Lack of pileup detection and rejection circuit (the so called Pileup Rej ector or PUR). The benefits of PUR were described in section 2.5. It's principle of operation is as follows: There is a fast shaper working in parallel with the spectroscopy channel (the digital processor in our case). The purpose of the fast channel is not to collect all of the signal charge, or to maximize signal to noise ratio, but instead to detect pulses that occur too close to each other in time. When these are detected, it sends a signal to disregard the current energy measurement because a pileup likely occurred. In effect, the PUR introduces dead time and removes most measurements done during pile up.

One way to add a PUR to the design is to include a fast analog shaper connected to the output of the preamplifier. A simple threshold comparator could detect the fast pulses and increment one of the MCU's counter registers. If the processor reads more than a value of 1 in that register during the detection and measurement of the integrator pulse, it could reject that pulse.

The problem of slow sampling time is based on need to wait for a detector signal pulse that has slow components to deposit most of its charge. There are scintillation materials like LaBr(Ce) that have negligible slow rising component and a dominant fast component of 30ns [9]. In this case it's easy to increase the sampling frequency and reduce the probability for pileups.

There is however, a clever way to compensate slow rise time components even in scintillators like NaI(Tl), if they can be reasonably approximated using with few exponents. If the current generated by a detector pulse can be approximated with the
following expression,

\[ i_D(t) = \frac{Q_1}{\tau_1} e^{-\frac{t}{\tau_1}} + \frac{Q_2}{\tau_2} e^{-\frac{t}{\tau_2}} + \frac{Q_3}{\tau_3} e^{-\frac{t}{\tau_3}} \cdots \tag{1} \]

then we could make the circuit on Fig. 6.1 generate a step function for each slow rising pulse by choosing appropriate values for the feedback resistors and capacitors. In the expression above, \( i_D(t) \) is the current produced when radiation is detected \( Q_1, \tau_1, Q_2, \tau_2, Q_3, \tau_3 \ldots \) represent the charge \( Q_n \) contained in the respective component with time constants \( \tau_n \).

Fig 6.1 Circuit of charge sensitive preamplifier with compensation for pulses with exponential rise time components.

If we take the Laplace transform of eq. (1), we get:

\[ I_D(s) = \frac{Q_1}{1 + s \tau_1} + \frac{Q_2}{1 + s \tau_2} + \frac{Q_3}{1 + s \tau_3} \cdots \]

The impedance of the feedback network on Fig. 6.1 in \( s \) domain is the combination of parallel RC impedance in each branch.

\[ Z(s) = \left( R1 + \frac{1}{s C_1} \right) \left( R2 + \frac{1}{s C_2} \right) \left( R3 + \frac{1}{s C_3} \right) \cdots \]
\[ Z(s) = \frac{1}{\left( \frac{1+s R_1 C_1}{s C_1} \right) + \left( \frac{1+s R_2 C_2}{s C_2} \right) + \left( \frac{1+s R_3 C_3}{s C_3} \right) + \ldots} \]

Then voltage at the op-amp output is \(-V_{\text{out}}(s) = I_D(s) \cdot Z(s)\). Therefore,

\[ -V_{\text{out}}(s) = I_D(s) \cdot Z(s) = \frac{Q_1}{(1+s \tau_1)} + \frac{Q_2}{(1+s \tau_2)} + \frac{Q_3}{(1+s \tau_3)} + \ldots \]

\[ = \left( \frac{s C_1}{1+s R_1 C_1} \right) + \left( \frac{s C_2}{1+s R_2 C_2} \right) + \left( \frac{s C_3}{1+s R_3 C_3} \right) + \ldots \]

If in the above expression, the capacitance and charge values are substituted with the following coefficients,

\[ C_2 = m_2 C_1, C_3 = m_3 C_1 \ldots \text{ and } Q_2 = m_2 Q_1, Q_3 = m_3 Q_1 \ldots \text{ where } m_x = \frac{C_x}{C_1} \ldots \text{ and } n_x = \frac{Q_x}{Q_1} \ldots \text{ then} \]

\[ -V_{\text{out}}(s) = \frac{Q_1}{s C_1} \cdot \left( \frac{1}{1+s R_1 C_1} \right) + \frac{n_2}{1+s R_2 C_2} + \frac{n_3}{1+s R_3 C_3} + \ldots \]

\[ = \left( \frac{1}{1+s R_1 C_1} \right) + \left( \frac{m_2}{1+s R_2 C_2} \right) + \left( \frac{m_3}{1+s R_3 C_3} \right) + \ldots \]

If we make \( \tau_1 = R_1 C_1, \tau_2 = R_2 C_2, \tau_3 = R_3 C_3 \ldots \text{ and } m_x = n_x \) we can cancel all terms except

\[ -V_{\text{out}}(s) = \frac{Q_1}{s C_1} \]

This means that if \( \frac{Q_2}{C_2} = \frac{Q_3}{C_3} = \frac{Q_4}{C_4} = \ldots = \frac{Q_1}{C_1} \) the output of the circuit will be a step function with amplitude \( \frac{Q_1}{C_1} \).
The most commonly used scintillator (and the one we use for enrichment monitoring) is NaI(Tl). It has two dominant signal components: one with rise time around 230 ns [1] that is very close to an exponent, and a slower component around 2.2 μs that can be approximated by an exponent with accuracy around 10%. This method was described and used in [4, 6, 26] to reduce NaI(Tl)'s slow component, and its temperature dependance.

Fig. 6.2 shows examples of compensated pulses from a plastic scintillator that has so-called pulse shape discrimination (PSD) property. (The signal has different duration depending on whether it was produced from gamma-ray or a fast neutron.)

![Graph showing compensated pulses](image)

**Fig 6.2** Example of actual plastic scintillator pulses (averaged) before and after applying exponential tail cancellation.

A plastic scintillator with PSD properties is a new development from Lawrence Livermore National Lab and with potential for commercialization by Eljen Technology. Fig 6.3 is an experiment we performed following a study of the PSD properties of the new material published in [27]. We used the compensation technique to condition the signal for easier PSD determination. Only the gamma response is shown because only the effectiveness of the compensation method is of interest in this paper.

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Another example is shown on Fig. 6.5 related to $^3$He proportional counter experiments.

### 6.3 Neutron Counting

In nuclear safeguards, a goal is to detect diversion of fissile material, such as Plutonium or Uranium, therefore an important part of the research is dedicated to the development of non-destructive assay methods (NDA) to detect fissile materials. A set of these methods, called coincidence and multiplicity counting, exploit a useful property of the neutrons emitted during fission, i.e. the number of neutrons produced by a single event can vary between one and eight, with a distribution $P(\nu)$ for the emission of $\nu$ neutrons.

The average $\nu = \sum \nu P(\nu)$ can vary from 2 to 8 for most of the events produce by spontaneous fission or thermal induced fission.[28] The neutrons emitted are correlated in time with the same event, which is a distinct signature of fission and fissionable material. In a neutron coincidence detector, neutrons from the same burst remain on average correlated in time in the detector, so the timing analysis of a pulse train generated by the detector can be used to extract the time correlation of the events. Neutrons of fissions have a kinetic energy of few MeV, and are generally indicated as fast neutrons.

A coincidence neutron detector is generally built using modules based on $^3$He proportional counter tubes embedded in polyethylene blocks, arranged to have a well structure, with the sample under test in the middle. Modules are wrapped with cadmium (which is a thermal neutron absorber) in order the signal to be only a function of the fast neutrons due to fissions in the sample under measurements. The polyethylene acts as a neutron moderator (material that slows down the fast neutrons), which increases the probability of the neutrons interacting with the $^3$He gas in the detector. This process was described in sections 1.2.2 and 1.5.1.

Algorithms have been developed to analyze the pulse train, and the two most common are Rossi-a and Feynman Variance-to-mean. A detail description of these methods is reported in [29].
Because the information is in the timing of the train pulse, it is evident that the design of
detector and electronics requires minimizing the dead time, i.e. the time during which the
neutron detector is not able to detect another event. Dead time limits the accuracy of a
NDA measurements.

This measurement requires detection of events that happened close in time, therefore the
dead time due to pileups has to be minimized even for low count-rates. Even though \(^3\)He
detectors are successfully used for this application, improving dead time will be
beneficial. These proportional counters have pulse shape that is highly non-uniform due
to diffident orientation of ionization tracks, and also they have long ionic components.
These properties were discussed in chapter 1. Fig 1.3 shows actual pulses from a
relatively fast \(^3\)He tube, where the nonconformity of both integrated, and filtered pulses is
shown. The ionic component visible in the integrated pulse contributes to slow decay in
the filtered one.

We are developing a method that uses pulse height analysis of an integrated charge signal
from the detector to eliminate the dead time in neutron counting even with slow
detectors. It's based on the observation that at the end of an integrated pulse that contains
pileups, the total charge will be simply the sum of all pulses put together, as illustrated on
Fig. 6.3.

![Fig 6.3 Examples of integrated pulses arriving at different times being added in amplitude in a pileup process.](image)
In a typical $^3$He pulse height distribution these pileups will be manifested as follows:

**Fig 6.4** $^3$He spectrum with indicated pileup region and their intensities in counts per second (cps). The dominant peak does not appear much larger than the rest of the spectrum because the vertical scale is logarithmic. Discriminator levels are indicated with blue bands.

One way to estimate the probability that an event was the result of a single pulse, a double pileup, a triple pileup, etc. is to simply put thresholds where the maximum single event can be, than do the same for the rest of possible pileups as is shown on the figure. This is not ideal because some low amplitude pileups will be counted in the no-pileup region, but it works if most events in the distribution lay in a dominant peak. This is roughly the case for a $^3$He spectrum as shown on Fig 1.2. It is also possible to deconvolve the distribution fully and find real probabilities of pileups.

Since, to perform this measurement we need an MCA that uses integrated signal with reset after each pulse, we can use the Free Running Integrator concept. The signal processing has to be modified to detect when a pulse ends and then perform a reset. A single detector system using this method will need tens of MCAs (one for each $^3$He tube). In this case, price and low power consumption become important. In addition, the removal of the very long ionic component is necessary in order not to extend the duration
of a single pulse too much. We have done this using the compensation method described in the previous section.

**Fig 6.5a** Timing diagrams of standard charge sensitive preamplifier (yellow); unipolar output (green) and amplifier BUSY signal (purple). The average duration about 3.2 μs was measured.

**Fig 6.5b** Timing diagrams of charge sensitive preamplifier with rise time correction (yellow); unipolar output (green) and amplifier BUSY signal (purple). The average duration about 1.7 μs

It should be noted that this measurement method has been tested (and the results obtained are shown above) with custom electronics that we developed, connected to standard gated integrator modules.

The use of the Free Running Integrator with modified signal processing is still under consideration.
References


