Measurement requirements and front-end design rules for gamma-ray tracking in large-volume germanium detectors through pulse-shape analysis

G. Ripamonti, A. Pullia, A. Geraci
Dipartimento di Elettronica e Informazione
Politecnico di Milano
Piazza L. Da Vinci 32, 20133 Milano, Italy
Phone (39) 2 2399-6178, Fax (39) 2 2367-604, E-Mail: ripamont@elet.polimi.it

Abstract* - In this paper we address the design of a mixed analogue-digital processor for pulse-shape analysis of signals delivered by large volume High-Purity-Germanium (HPGe) detectors for gamma-ray spectroscopy. The information to be extracted from pulse-shape analysis is the position of the interaction of the γ photon inside the detector. We discuss the structure of the measurement apparatus and the required characteristics of its components, with a special emphasis on the resolution and sampling frequency of the analogue-to-digital converter (ADC). We also address the problem of the occurrence time estimate of the events, which is strictly related to the pulse shape. A modular hardware solution for pulse-shape measurements is proposed.

I. INTRODUCTION

High-resolution nuclear-spectroscopy systems aim at measuring, with the maximum possible accuracy and precision, the characteristics of the radiation (X, γ) emanated in nuclear reactions, while simultaneously providing high throughput and stability. Mixed analogue-digital signal conditioning has proven to be useful to enhance the performances of the measurement setup. Analogue filters are used to preshape the signal in such a way as to minimise the requirements of the ADC (Analogue-to-Digital Converter). The cascaded digital section gets the required information from the digitised data more efficiently than in fully analogue setups [1].

In large-volume solid-state detectors (e.g. HPGe γ-ray detectors cooled to cryogenic temperatures) the characteristics of the events to be measured are: *i*) the energy of the impinging photons, which is proportional to the area of the delivered current signals; *ii*) the time of occurrence of the events, which permits to use coincidence-anticoincidence measurement techniques to establish correlations; *iii*) the three-dimension (3D) coordinate of interaction of the photon in the detector bulk.

On the one hand the digital measurement of the event energy has been already extensively addressed in the literature [2-7]. On the other hand the growing interest in digital methods for optimal filtering of the signals supplied by such detectors is now triggering the research on the other mentioned measurement tasks: determination of the arrival time and localization of the events. This paper deals with the topic of digital processing of the signals delivered by HPGe detectors, and is focused on the problem of the localization of the interaction point. It is worth pointing out that segmentation of the detectors has been proposed to guarantee that in most cases no double or triple event (e.g. multiple Compton scattering) occurs within each individual segment, thus simplifying the data analysis.

The shape of the signal generated at the detector electrodes is substantially dependent on the drift path of the charge cloud generated in the interaction between the γ photon and the detector. Thus some information on the interaction position must be achievable from such shape, namely from the structure of the pulse rising edge. The geometry of most HPGe detectors is almost cylindrical and so we focus on the estimate of the radial position of the interaction. In modern experiments of nuclear physics, with up to thousands of parallel channels, it is mandatory to perform the measurements with the least possible number of input data, while losing no important information and, possibly, operating on line in real time. We formulate here the problem of pulse-shape analysis in the frequency domain, by investigating the mutual spectral location of signal and noise (electronic and quantization noises). This approach permits to identify the requirements of the ADC, in terms of its sampling frequency and its resolution. The sampling frequency should be sufficient to cover the bandwidth where the spectral components of the signal dominate over the noise, but contextually as low as possible to minimize the data storage depth. The resolution, or number of bits, should guarantee that the quantization noise is lower than the electronic noise in the bandwidth of interest.

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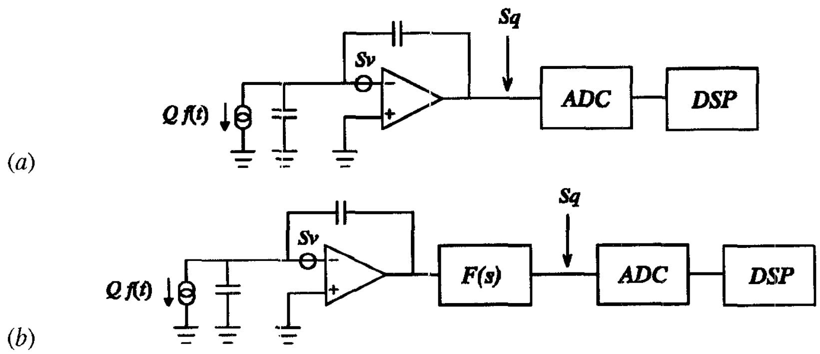


Figure 1. The detector delivers a current pulse *Qf(t)*. *Q* is the charge cloud, *f(t)* is a unit area function, describing the pulse shape. (*a*) The signal is sampled after integration. The electronic series noise (S_V) and the quantization noise (Sq) are highlighted. (*b*) Same as in Fig.1*a* but with an analog preshaper F(s). F(s) is its transfer function being "s" the independent variable in the Laplace domain.

II. SYSTEM CHARACTERISTICS

We have taken into account two possible configurations for pulse-shape measurement.

In the first one (Fig.1*a*), the analogue section of the processor is a single-pole low-pass filter (the *"preamplifier"*), which - roughly speaking - "spreads and flattens" the time-domain shape of the input current impulse. It adds a white electronic noise at its output and is cascaded by the ADC, which is supposed to introduce a white quantization-noise contribution in the Nyquist bandwidth. In this way, the total noise is white. Note that the whiteness of the noises is considered a good approximation in the system bandwidth. Concerning the signal, the drawback of this solution is the superposition of successive events (the *"pile-up"*), which can saturate the input dynamics of the ADC, if it is not properly dimensioned on the basis of the expected occurrence rate of detected photons.

In the second configuration an analogue stage is added between the preamplifier and the ADC (Fig.1*b*). In this case, sampling the output of this suitable filter means that: *i*) the electronic noise is shaped by the transfer function F(s) of this additional block and compares itself with the spectrally white quantization noise; *ii*) pulse pileup can be practically avoided, provided that a cut of the dc and low frequencies is introduced in F(s); *iii*) quantization noise has to be much less than electronic-noise power density only where the signal is dominating the electronic noise itself. The most reasonable choice for F(s) is differentiation of the preamplifier output.

In this way the electronic noise, as referred to the input of the ADC, increases as f^2 (f is frequency) and can be kept larger than quantization noise for frequencies inside the signal spectral window. Moreover, the signals superposition is practically eliminated. A quantitative analysis of its effect is carried out in the next Section.

The current signals delivered at the detector electrode vs the radial coordinate of interaction may be derived under some simplifying hypotheses, in closed form in the case of indefinite cylindrical geometry. The details of the analytical computation used to derive such shapes, which is based on the weight-field method [8], are not reported here for the sake of clarity. The signal profiles, expressed as the time integral of a current, or an electronic charge, are reported in Fig.2 for a detector with an outer radius of 5cm, biased with a reverse voltage of 1000V. It can be noted that each profile shows two points of discontinuity in its time derivative. They occur at the times of collection of the electron and hole clouds, which respectively reach the front and back electrodes at different times. In the considered example the time width of all shapes is less than 500ns.

III. IMPACT OF QUANTIZATION NOISE

Let us consider first the setup depicted in Fig.1*a*. It can be demonstrated that in order to get quantization-noise power

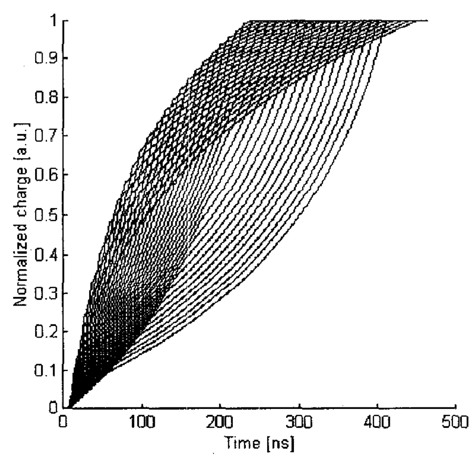
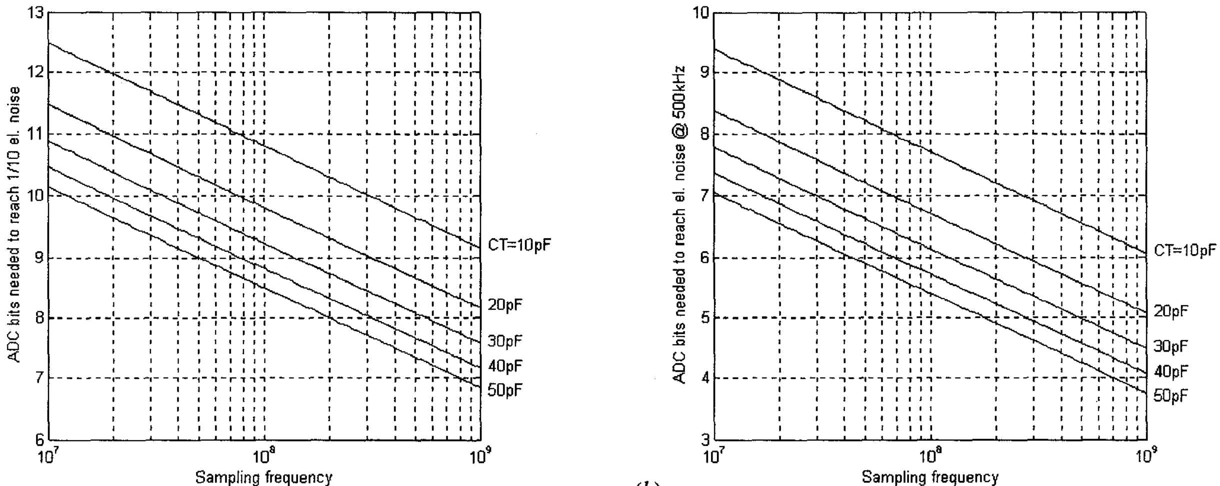


Figure 2. Structure of the rising edge of the charge signal delivered by a cylindrical detector with an outer radius of 5cm. The pulse shapes correspond to interaction points displaced by 1mm throughout the radial coordinate. They are all contained in a time window of about 400ns.



(*a*) (*b*)

Figure 3. The ADC resolution needed to make the quantization noise negligible as compared to the electronic noise is plot vs. the ADC sampling frequency. (*a*) With the setup of Fig.1*a* a 12bit ADC is needed in a relatively large range of total input capacitances C_T and sampling frequencies. (*b*) Assuming that in the setup of Fig.1*b* F(*s*) is derivator stage, the ADC resolution needed to make the quantization noise negligible against the electronic noise is only 8bits in all considered conditions.

density to be much less than electronic-noise power density means:

$$\frac{N_{el}}{N_q} = \sqrt{12}\left(\frac{\sqrt{S_V}C_T}{kQ}\right)\left(2^N\sqrt{f_{Ny}}\right) >> 1 \qquad (1)$$

where N_{el} and N_q are the rms noises (electronic and quantization noises), S_V is the power-spectral-density of the input series noise (Fig.1*a*), C_T is the total input capacitance of the preamplifier, Q is the maximum charge delivered by the detector, k is a factor used to extend the charge dynamics so as to accomodate event pileup (typically k=4), N is the ADC resolution, f_{Ny} the Nyquist frequency. Note that the first term in parentheses depends on the front end, and that the second depends on the ADC. In Fig. 3*a* N is derived from (1) for typical values of Sv and Q, while putting N_{el}/N_q=10. It is seen that this solution implies to have an ADC with 11 bits in order to reduce the quantization noise to be one tenth of the electronic noise at 100 MHz of sampling frequency and with a total input capacitance of 10pF.

Consider now the setup of Fig.1*b*. The electronic noise can be compared to the quantization noise by referring both noises to the input (equivalent current noises). It is relatively easy to show that the electronic noise is proportional to f^2 whereas the quantization noise is white. Thus they meet at some angular cross frequency ω_c, and the electronic noise dominates over the quantization noise, beyond. From simple calculations it turns out that

$$\omega_c = \frac{1}{\sqrt{12}}\left(\frac{I_p}{C_T\sqrt{S_V}}\right)\left(\frac{1}{2^N\sqrt{f_{Ny}}}\right) \qquad (2)$$

where I_p is the maximum peak current of the input signals. Note again that the first term in parentheses depends on the front end, and that the second depends on the ADC. In Fig.3*b* N is derived from (2) for values of Sv and I_p con sistent with the previous case, while putting $\omega_C = 2\pi 500$kHz. It is seen that, at 100MHz of sampling frequency and with a total input capacitance of 10pF, 8 bit are needed in order to equalize the electronic noise to the quantization noise at 500kHz, which is by far outside the spectral region containing the signal (i.e. a rising edge lasting about 400ns, as shown in Fig. 2). This latter solution looks advantageous, because the resolution (and cost) of the ADC is more relaxed. As a drawback an extra analogue stage F(s) is to be used in the electronic chain.

For both setups, Figure 4 shows the FFT of the current signals (viz. the time derivatives of the curves shown in Fig.2) for various interaction radii inside the volume of the detector, compared to the present noise spectral densities. As we are interested in discriminating between contiguous radial positions, the information is stored in the difference of the signals coming from different interaction locations. Figure 5 is the same of Fig.4 but in terms of these differences. It is clear from the crossing point of signal and noise spectra (i.e. signal-to-noise ratio equal to one) that a sampling frequency of 100MHz and a resolution of 8bit are

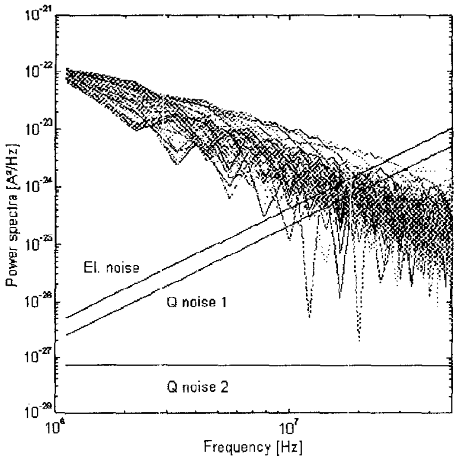


Figure 4. Signals and noise in the frequency domain. A sampling frequency of 100MHz is used. Q noise 1 is the quantization noise referred to the input in the system of Fig.1*a*. Q noise 2 is the quantization noise referred to the input in the system of Fig.1*b*.

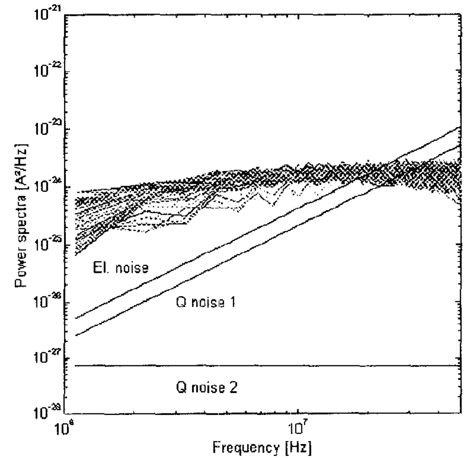


Figure 5. Similar to Fig.4, but the differences between adjacent pulse shapes (see Fig.2) are considered.

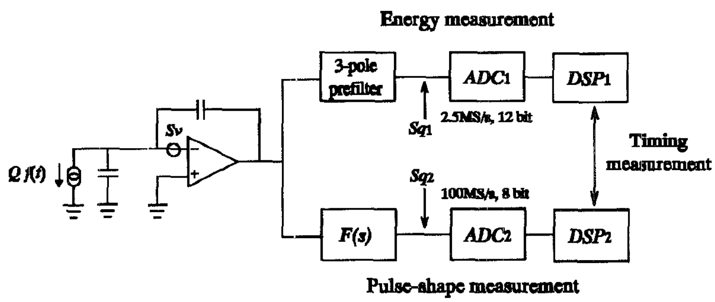


Fig. 6. Schematic of the proposed processor architecture. Besides a slow 2.5Ms/s 12bit ADC used for the estimate of the charge Q (proportional to the energy of the photon), a fast 100Ms/s 8bit ADC is used for pulse-shape analysis. A sinergy between the chains is possible for time measurements.

sufficient if a differentiating stage between the preamplifier and the ADC is used.

As the signal arrival time is not necessarily synchronous with the sampling comb, the problem arises of the jitter of the shape to be analysed. We have addressed this problem and found that the best fit of the interaction position implies a knowledge of the arrival time with much less than 10ns resolution at 100MHz of sampling frequency. This piece of information could be derived as a global time estimation, that is, an average time estimate from many events occurring in coincidence (if available) on many detectors and including also the actual signal shape.

In Fig.6, we propose the architecture of a global modular hardware solution for energy and pulse shape measurements. It comprises a standard channel for energy measurements [3] with a very slow 12-bit ADC and a channel for pulse-shape measurements with a faster 8-bit ADC. A sinergy between the two channels may be considered for the estimation of the event arrival time, if an external timing is not available.

IV. CONCLUSIONS

The design rules of a mixed analogue-digital setup to be used for pulse-shape measurements of nuclear signals have been analyzed with a frequency-domain approach. The quantization-noise impact has been evaluated to derive the features of the ADC. It has been shown that a suitable analogue preprocessor may be used to relax the requirements of the ADC, particularly its resolution. No detail has been given about the alghoritm which could be used to derive the interaction position vs the pulse shape. Such alghoritm will be considered and described in a future work. However the estimation procedure which is under study has a two-step structure. In the first step a rough non-linear estimation method is used, so as to determine a "bias point" for the subsequent step. Then a more precise estimation, obtained through a linearization of the relationships in the bias point, is used to get the final radial-position measurement.

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