NEW DIGITISERS FOR POSITION SENSITIVE ^{3}HE PROPORTIONAL COUNTERS

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Abstract

The present paper reports on the development of a new digital front-end electronics for position sensitive ^{3}He neutron detectors. The classical analog approach for the charge readout, consisting of a shaping amplifier coupled with a peak sensing ADC, has been replaced by a 64 channels, 62.5 Msample/s and 12 bit digitiser. Excellent results have been obtained in terms of position resolution and signal to noise ratio when adopting a continuous digital filtering and gaussian shaping.

INTRODUCTION

^{3}He gas-filled detectors are a classical choice for the detection of thermal and cold neutrons. The incident neutrons are captured by the ^{3}He producing a tritium and an hydrogen which are sharing the 756 keV of energy generated in the reaction.

$$^3He + n \rightarrow {}^3H + {}^1H + 756keV \quad (1)$$

The electron avalanche initiated by the 2 ions generates the detector signal.

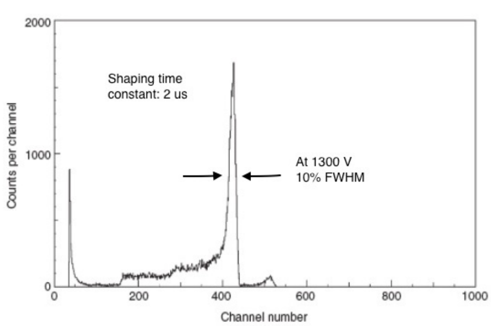

Figure 1: Pulse-height spectrum from ^{3}He tube.

Figure 1 shows a typical pulse-height spectrum from a ^{3}He tube at low gain. At higher gain, non-linear effects set in and the proportionality is lost. This limits the charge one can reasonably obtain from a neutron capture event to about 1 pC. The shape of the spectrum is due to the kinematic of the reaction but also to the choice of the amplifier time constant. The full peak is originated by the collection of the total energy of the two ions (765 keV). If one of the ions is absorbed by the tube walls then, the total collected energy is smaller. This results in the peak asymmetry visible at the left part of the peak in Fig. 1. A choice of a shaping time ranging from 0.5 to 2 μs is a good compromise between good resolution and high count-rate capability.

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The classical geometry of a charge-division neutron detector consists of a cylindrical volume housing a resistive anode (see Fig. 2).

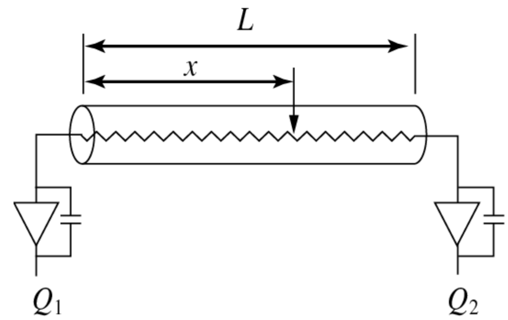

Figure 2: Principle of a charge-division neutron detector.

Electrical signals are extracted at both ends of the tube and the information about the interaction point along the tube can be derived by the ratio of the collected charged at both ends. The theoretical spacial resolution achievable for a detector of total length L is function of the temperature T of the resistive electrode, of the total detector capacity C and of the collected charge. This can be expressed by

$$\frac{FWHM}{L} = \frac{2.54 \cdot \sqrt{kTC}}{Q_1 + Q_2} \quad (2)$$

When using low-noise amplifiers, one can expect from the previous equation (2) a spacial resolution in the order of 0.1% of the tube length. However, Eq. (2) is not applicable in the case of ^{3}He neutron detectors because the charge collection time is much longer than the RC time constant of the detector, as shown in Fig. 3. This results in an actual spacial resolution of the order of 1% of the tube length. The limiting factor being dominated by the noise of the resistive wire and by the integration time.

FRONT-END ELECTRONICS

Our current front-end analog electronics for charge-division consists of a shaping amplifier containing a 4th order Gaussian filter, a baseline correction circuit and a single-shot 12 bit ADC [1], while the position is coded with a 8-bit word due to the limited resources available. The essential idea is to implement digitally an equivalent signal treatment using the CAEN type V1740 [2] digitiser board with an ad-hoc firmware. Since the pulse-shape of the signals depends on the interaction location and on the charges propagation direction (see Fig. 3), the conventional pulse-shape analysis is inappropriate.

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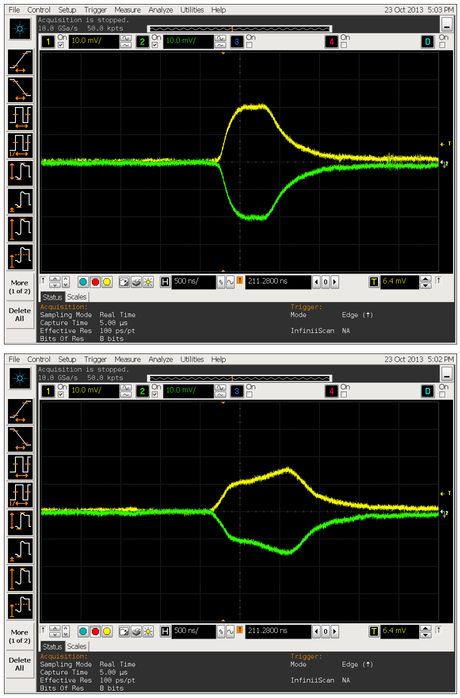

Figure 3: Examples of differential pulse-shapes from ^{3}He tube.

Moreover, in order to avoid errors in the position determination, each couple of digitiser's channels must have a common trigger based on the sum of the detected charges.

DIGITAL SIGNAL PROCESSING

Gaussian Filter

A 4$^{\text{th}}$ order approximation of a true Gaussian profile can be expressed by the following normalised transfer function, which has 4 poles in the complex plane:

$$H(s) = \frac{4.899}{4.899 + 11.42 \cdot s + 10.87 \cdot s^2 + 5.073 \cdot s^3 + s^4} \quad (3)$$

If we apply the input invariance method [3] that preserves the discrete samples of the impulse response of the continuous-time domain as the samples of the discrete-time impulse response, we can only transform exponential components of the form:

$$H(s) = \frac{1}{s - s_0} \quad (4)$$

which have continuous-time impulse response

$$h(t) = e^{s_0 t} \quad (5)$$

By sampling the response 5 at a period T$_s$ we can represent the resulting geometric series in the Z-domain as:

$$H_z(z) = \frac{T_s}{1 - e^{s_0 T_s} z^{-1}} \quad (6)$$

If our continuous-time transfer function can be written as a linear combination of this kind of single-pole functions, the technique generalises and the corresponding Z-transform results:

$$H_z(z) = T_s \cdot \Sigma_i \frac{A_i}{1 - e^{s_i T_s} z^{-1}} \quad (7)$$

Therefore, if we can write the original continuous-time domain transfer function as a partial-fraction expansion with simple terms (which means that we don't have double poles), we can obtain the partial-fraction expansion of the corresponding Z-transform with the same coefficients (A_i in Eq. (8), and with the poles following a $s_i \rightarrow exp(s_i T_s)$ mapping. However, in principle the recombination of all these terms into a single rational function implies the possible appearance of finite zeros. We neglect these zeros assuming they remain at the origin to avoid the increase of the number of multiplications in the filter implementation since we are limited in computing resources in the FPGA. This approximation is reasonable since the zeros are less important the the poles. They apply a FIR filter to the input of a length equal to the order of the filter (4 in our case). This will only have an impact if all the coefficients in the numerator of Eq. (6) were 0. In reality those are all positive, in which case the FIR filter will only calculate a kind of weighted average over 5 successive samples. In addition, it turns out that only 3 coefficients have a significant value and therefore, this term can be neglected as far as the time constant of the filter is much more that 3 samples. The denominator of Eq. (8), corresponding to the IIR filter, has the dominant effect on the impulse response. With the previous considerations, we can reformulate Eq. (8) as:

$$H_z(z) = \Pi_i \frac{constant}{1 - e^{s_i T_s / 2\pi\tau} z^{-1}} \quad (8)$$

where *i* runs over the 4 s-plane poles mentioned earlier. This will result in 2 second-order factors for the denominator

$$(1 - a_1 z^{-1} + b_1 z^{-2})(1 - a_2 z^{-1} + b_2 z^{-2}) \quad (9)$$

which can easily be implemented as a succession of two second-order IIR filters. The fixed-point implementation will consist in multiplying the constants a1 and b1 of Eq. (9) with a power of 2, applying the integer calculation and dividing the result by the same power of 2. We opted for the 14$^{\text{th}}$ power of 2. It can also be interesting to multiply input and output with a small power of 2 to get some *digits after the comma*. These can be removed at the end of the calculation, to take care of part of the rounding errors.

Pole-zero Compensation

The continuous-time domain pole-zero compensation comes down in accepting a signal that went through a first order system with a long time constant, and modifying this such that it looks as if the signal went trough a first order system with a much shorter time constant. This can simply be implemented with the following transfer function:

$$H(s) = \frac{s + 1/\tau_l}{s + 1/\tau_s} \quad (10)$$

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where τ_s is the short time constant and τ_l is the long one for which we want to compensate. In this case, τ_s is our own choice while τ_l must match precisely the time constant of the system. Transforming Eq. (10) in the discrete domain we can write

$$Z(z) = 1 + (T_s/\tau_l - T_s/\tau_s) \cdot \frac{1}{1 - e^{-T_s/\tau_s} z^{-1}} \tag{11}$$

If τ_s is much bigger that the sample period one can approximate $e^{-T_s/\tau_s} \approx 1 - T_s/\tau_s$. This can, then, be implemented as a linear combination of the direct input and a first order IIR filter:

$$w[n] = i[n] + (1 - \frac{T_s}{\tau_s}) \cdot w[n-1] \tag{12}$$

$$u[n] = i[n] + (T_s/\tau_l - T_s/\tau_s) \cdot w[n] \tag{13}$$

where we can calculate in fixed point integer and shift the result with a certain power of 2 to allow integer multiplication, followed by a division and truncation by the same power of 2.

Baseline Correction

This correction neutralises offset in the signal input and it adjusts the signal level in between pulses to zero. This means that the average of the signal is not zero (given the pulses are positive), and hence that the AC coupling is undone. The base line correction is a non-linear operation. Our implementation consists of chopping the samples up in chunks of length N. In every of these chunks we determine the minimal sample value and we fed this value to a first order low-pass digital filter:

$$w[n] = \frac{1}{k} \cdot i[n] + (1 - \frac{1}{k}) \cdot w[n-1] \tag{14}$$

If k is a power of 2 then, this can easily be implemented with just shifting binary words. In Eq. (14) $i[n]$ is the minimum of the last chunk and n counts the number of chunks, while $w[n]$ is the estimated baseline of the input signal. This method implies of course a small error since the minimal sample will be the minimum of the noise excursion and not the average baseline value. Therefore, on a pure noise signal, the average level after correction will not be 0, but it will be the negative of the average negative excursion (due to noise) on N samples. This amounts to a few sigma of the noise level if the noise has a Gaussian distribution.

Implementation

The signal treatment per channel consists of 5 blocs that can be independently activated or deactivated:

- first baseline correction applied to the incoming signal
- pole-zero compensation
- first second-order Gaussian filter
- second second-order Gaussian filter
- final baseline correction

The input is a 12-bit unsigned data and the output is a 16-bit unsigned data while between blocs the signals are of signed type to be able to treat positive as well as negative samples. All blocs use fixed-point arithmetic with a certain binary fraction to avoid too many significant rounding errors while saving resources.

DATA TREATMENT

Trigger

To avoid having a different trigger efficiency along the tube, the channels of the digitiser have been paired. Indeed, a neutron interacting close to one edge of the tube will generate a large signal on that end and almost no signal on the other end. To avoid missing triggers when the signal amplitude is very low, each channel independently makes the sum of his proper output data stream, and the data stream of its paired channel. A local trigger is fired if that sum crosses the set threshold for the channel. As such, if the paired channel (who is treating the same sum of course) has the same threshold setting, both independent triggers will fire at exactly the same clock pulse even though the trigger signals themselves are not coupled.

Maximum

When the trigger is fired a maximum finder on the sum signal will find that sample which is the largest since the trigger was fired. The corresponding local channel sample value is stored until a larger value on the sum is found or a new trigger is issued. As all this happens in real time, the value should be read out after the maximum has been reached on the output data stream, and before a next trigger is fired. The local output sample value corresponding to the time tick when the sum of the two channel outputs reached its global maximum since the threshold crossing, is what is sent to the output stream on an unsigned, 16-bit word.

RESULTS

To test the quality of the digital signal treatment and the achievable resolution we have compared the results obtained with our standard analog system versus those from the digitiser. To decouple possible problem in the firmware implementation from those of the algorithm we have first treated offline the samples from the digitiser with a c++ code implementing the digital filter. Signals at 1 kHz rate from an Agilent waveform generator have been injected into a PAD02 [1] shaping amplifier implementing in an analog way the Gaussian filter with a time constant of about 1.4 μs. A resistance of 3 kΩ was used to simulate the noise from the resistive wire of the ^{3}He gas detector. The same PAD02 amplifier but without the Gaussian shaping part has been used to amplify the signals before sending them to the digitiser. The shaping time of the digital filter has been set to 1.6 μs.

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Table 1: Position Resolution Obtained With Analog and Digital Front-End Electronics

	Analog	Charge Integ.	C++ code	Firmware
Injected charge (pC)	0.2	0.2	0.2	0.2
	0.6	0.6	0.6	0.6
	1.0	1.0	1.0	1.0
Dynamic	256	4096	4096	65536
FWHM (measured)	4	65	124	553
	2	20	43	187
	1	15	25	113
FWHM (theory)	3.9	63.5	119.8	507.9
	1.3	21.1	39.9	169.3
	0.8	12.7	23.9	101.6
Resolution (%)	1.56	3.03	1.59	1.69
	0.78	1.05	0.49	0.57
	0.39	0.61	0.37	0.34

Table 1 summarises the obtained results. The measured FWHM of the position distribution has been compared with the theoretical one. This last include only the noise from the amplifier and the resistance. Therefore, a FWHM value close to the theory implies that the digitalisation and the Gaussian filtering do not contribute significantly to the noise balance. One can notice that, as expected, no substantial difference exists between the off-line (C++ code) and the FPGA implementation of the algorithm. The same measurements have been repeated using the standard firmware provided with the CAEN digitiser. In this case (column 3 in Table 1), a simple charge integration is performed. The obtained resolution is about a factor of 2 worst than for the Gaussian filter. A second set of measurements was performed using a 1 kΩ to simulate a different resistive wire and the obtained results are consistent with those reported in Table 1. For the final firmware validation we have acquired data using a large ^{3}He detector counting 128 tubes each of 1 m hight and 8 mm diameter and containing 12 bar of ^{3}He. This detector, installed at the D22 [4] beamline of Institut Laue-Langeving, provides a detection efficiency of about 70% at a wavelength of 6 A. Due to the lack of channels available in the prototype digitiser we could only use 8 out of the 128 tubes.

Figure 4 depicts the obtained results when exposing the detector to a AmBe neutron source. A Boral grid with horizontal linear gaps of 2 mm hight was placed right in front of the detector. Left image has been obtained with our standard

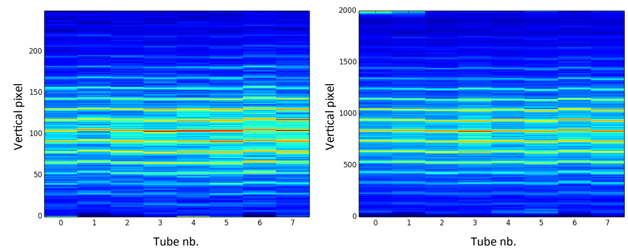

Figure 4: Comparison of D22 detector images acquired with our standard front-end electronics (left) and with the new digital system (right).

front-end electronics while the right side is the result from the digitiser running our new charge-division firmware. One can notice the excellent spacial resolution and, as expected from Table 1, no significant differences with respect to the analog approach.

CONCLUSIONS

Using as base hardware a commercial digitiser board from CAEN [2] we have implemented in the existing FPGA of the board a new charge-division firmware especially designed for position sensitive ^{3}He proportional counters. A 4^{th} order digital Gaussian filter for the pulse shaping, as well as pole-zero and baseline correction have been included in the firmware. Obtained results both on laboratory test and on real detector show excellent performances of the digital electronics that will be soon ready for commissioning on various instruments at the Insitut Laue-Langevin.

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