A "Cold" Discharge Mechanism for Low-Noise Fast Charge Amplifiers

Alberto Pullia, Roberto Bassini, Ciro Boiano, and Sergio Brambilla

***Abstract*—We present a new discharge mechanism for low-noise fast preamplifiers for γ-ray spectroscopy. Such circuital solution has been conceived in the framework of the INFN Mars experiment, which imposes stringent requirements for both the signal-to-noise ratio and the rise time of the front-end preamplifiers. Basically, a noninverting low-gain (G) stage is interposed between the output of a conventional charge amplifier and the high-value feedback resistor. It can be easily found that this adds no noise but the discharge time constant is reduced by a factor G, because the voltage drop across the resistor is G times as high, and such is the discharge current. A similar effect could also be obtained in the standard charge amplifier by diminishing the feedback resistor by a factor G. But this would yield an increase of a factor G of its thermal current noise. In this sense our equivalent discharge resistor is "cold," because it carries a current noise G times as low. This permits us to use a relatively large feedback capacitance which inherently yields a fast risetime. A rise time of 15 ns and an overall electronic noise of 1.03 keV FWHM at 3 μs shaping time have been obtained, with an input capacitance of 33 pF, and 1.5 pF feedback capacitor, 150 μs discharge time constant.**

***Index Terms*—Charge amplifier, low-noise amplifier, cold resistor, gamma-ray spectroscopy, pulse-shape analysis.**

I. Introduction

THE NEW generation of High Purity Germanium (HPGe) γ-ray detectors for large nuclear-physics experiments is characterized by segmentation of the electrode distributed on the outer surface of the germanium crystal, and an inherent increase of the number of read-out channels [1]. Such segmentation permits to observe from different viewpoints the mobile charges travelling across the detection crystal under the action of the present electric field. This is aimed at individuating by pulseshape analysis the interaction points of the γ photon inside the detector so as to reconstruct the cinematic of the Compton scattering sequence. This information, in turn, is used for estimating the angle between the trajectories of the impinging-photon and the radioactive source, which eventually permits to introduce the Doppler correction in the energy spectra [2]. In order to reconstruct the 3-D positions of the collision points, pulseshape analysis is used. In fact, depending on the time of flight of the electron and hole clouds generated in the collision, different structures of the charge signal induced at the electrodes are seen [3]. As an example in Fig. 1 such a signal as seen at the central electrode is reported in the case of a truly-coaxial germanium detector of realistic dimensions. It can be seen that the pulse shape is contained in a 400–ns time window. In order to resolve such structure, a fast rise time of the preamplifier is required, at least one order of magnitude shorter than such a time window. But this requirement is difficult to attain with these large-capacitance detectors while using standard charge preamplifiers for nuclear spectroscopy, unless an increase of the discharge time or of the noise is accepted, as will be shown in the next section. The circuital topology proposed in this paper realizes a low—noise discharge mechanism which in turn permits to obtain a fast enough risetime without increasing the discharge time and with a negligible sacrifice in terms of electronic noise and voltage/charge sensitivity.

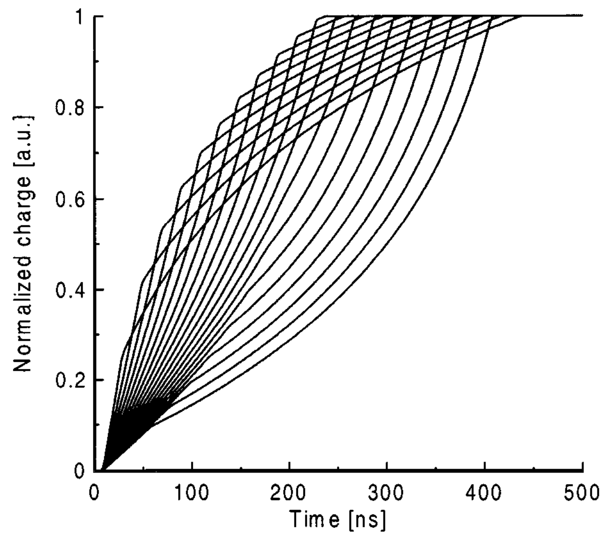

Fig. 1. Structure of the rising edge of the charge signal delivered by a cylindrical HPGe detector with an outer radius of 5 cm (obtained from computer simulation). The pulse shapes correspond to interaction points displaced by 2 mm throughout the radial coordinate. They are all contained in a time window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

II. Background

Let us first consider a standard charge preamplifier (Fig. 2) constituted by a field effect transistor (FET) as its input transistor, a cascode-configured amplifying stage, a feedback capacitance with a resistor connected in parallel for continuous reset of the charge stored in the capacitor. Obviously enough, the bandwidth (or risetime) requirement is difficult to keep as

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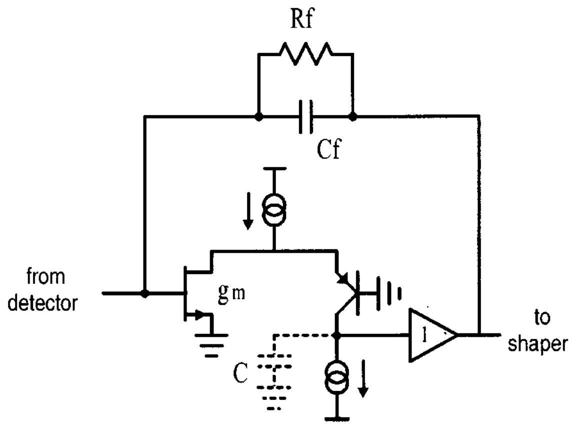

Fig. 2. Simplified schematic of a standard charge preamplifier. The detector adds a capacitance (not shown) at the input of the order of tens of pF for segmented HPGe detectors.

the detector capacitance is increased. In fact the risetime T_r of the preamplifier is given by [4]

$$T_r = C_t \left(\frac{C}{C_f \times g_m} \right) \tag{1}$$

where

- C_t sum of detector, preamplifier-input, and feedback capacitances;
- C_f feedback capacitance;
- C internal capacitance of the amplifying node of the preamplifier;
- g_m transconductance of the preamplifier input transistor.

Equation (1) is usually interpreted as the input CR time constant, given by the product of the total input capacitance C_t and the so-called cold resistor (term in parentheses), i.e., the resistive part of the impedance of the charge–sensitive loop.

Note also that the voltage-to-charge sensitivity S_{vq} of the preamplifier is given, apart from sign, by the inverse of the feedback capacitance, or

$$S_{vq} = -\frac{1}{C_f}. \tag{2}$$

Let us assume that the term (C/g_m) be a design constant. It can be seen from (1) that in order to keep T_r constant as the detector capacitance (and, in consequence, also C_t) is increased, the feedback capacitance C_f must also be increased. This both reduces the voltage-to-charge sensitivity (2) of the preamplifier and increases the discharge time-constant $R_f \times C_f$, R_f being the high–value feedback resistor (typically of the order of 1 GΩ). To enhance the voltage-to-charge sensitivity a low-noise gain stage is usually used, as shown in Fig. 3. A Pole-Zero (P-Z) cancellation network is also normally inserted to reduce the decay time constant (typically from 1 ms to 100 μs) before amplification, because the random pileup of pulses would otherwise easily saturate the output voltage swing at high count rates. Rarely is it preferred to reduce R_f and omit the P-Z network, because the enhanced current noise contributed by lowervalue R_fs would seriously limit the performance of the preamplifier, especially at long shaping times.

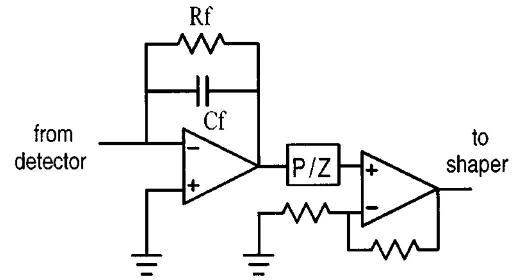

Fig. 3. A standard approach to increase charge sensitivity while reducing the decay time. Note that a P-Z network is required.

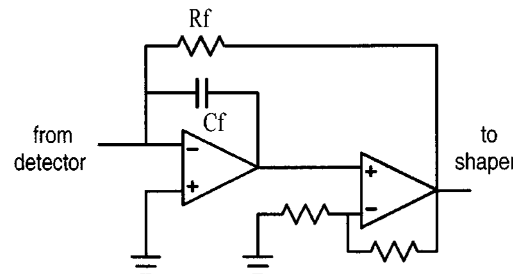

Fig. 4. Proposed approach. The discharge current is increased due to the increased voltage drop across R_f. The decay time constant is thus reduced with no need of a P/Z network.

III. METHOD

Our solution is proposed instead in Fig. 4. We insert the gain stage (say G) in between the charge amplifier output and the feedback resistor R_f, with no use of a P-Z network. Such a solution 1) enhances the voltage-to-charge sensitivity, 2) provides a faster discharge time with no need to trim a P-Z network,and 3) adds no significant additional noise, because the feedback resistor value is kept high with an inherently low current noise. The mechanism which reduces the discharge time in this circuit is rather intuitive: the voltage drop across R_f is G times as high as that across C_f, because the gain stage is interposed between the right ends of C_f and R_f. Hence the discharge current is also G times as high and the discharge time constant T_d is G times as low, or

$$T_d = \frac{R_f \times C_f}{G}. \tag{3}$$

The discharge time (3) of C_f is in practice seen as the decay time constant of its right-end voltage, and such an exponential decay is also seen after the gain stage, or at the preamplifier output. So, on the one hand the decay time constant seen at the preamplifier output is G times shorter than the time constant $R_f \times C_f$. On the other hand the equivalent input current noise density S_I contributed by R_f is still

$$S_I = \frac{4KT}{R_f}. \tag{4}$$

From (3) and (4), it can be seen that as far as the decay time constant is concerned the equivalent discharge resistor is R_f/G, but as far as noise is concerned the resistor to be considered is R_f. Therefore, it can be concluded that the equivalent discharge resistor R_f/G is less noisy of a physical resistor of the same value. In this sense it is a “cold resistor.”

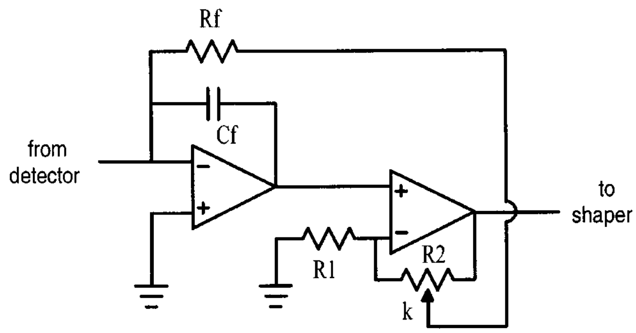

Fig. 5. Circuit used to dose the discharge mechanism by trimming $R2$.

The voltage-to-charge sensitivity S_{vq} of the preamplifier is obviously given by

$$S_{vq} = -\frac{1}{C_f} \times G. \tag{5}$$

Eventually it must be observed that (1) still holds for the risetime of the first stage of the circuit of Fig. 4. The very output of the preamplifier is slowed down by the finite bandwidth of the gain stage, but this strictly depends on the gain-bandwidth product of the used operational amplifier and poses no conceptual limitation. So (1), (3)–(5) characterize the preamplifier of Fig. 4. To summarize, it can be seen that by increasing C_f; 1) the risetime (1) is made faster; 2) by choosing a proper value for G in (3) the increase of the time constant $C_f \times R_f$ may be compensated with no need of a P-Z network and no change of R_f; 3) there is no impact on the noise (4) because R_f is not changed; and 4) the voltage-to-charge sensitivity (5) is enhanced by factor G.

The absence of a P-Z network guarantees a virtually perfect single-time-constant exponential decay at the preamplifier output with no need of trimming, and no risk to have residual uncompensated singularities in its transfer function, which is a particularly important feature if a digital processing of the signal is used. In fact it has been found that the synthesis of quasioptimal digital filters in nuclear spectroscopy may be seriously compromised by the presence of unwanted singularities in the analog waveform supplied to the ADC (analog to digital converter), such those arising from imperfect P-Z compensation. To exploit the potentialities of digital filtering a virtually perfect P-Z compensation is really required [5].

IV. Experimental Tests

Tests of the circuit have been done by using a standard discrete charge-amplifier properly modified. In order to investigate experimentally the effectiveness of the discharge mechanism of the structure of Fig. 4 we have added a trimmer which permits to dose the amount of signal circulating in the second feedback loop, as shown in Fig. 5. By trimming $R2$ through parameter k, it is possible to dose the amount of the amplified signal carried back to R_f. When $k = 0$ the right ends of R_f and C_f are virtually connected and no effect, other than amplification is brought about by second stage of the circuit of Fig. 5. When $k = 1$, instead, the second loop is fully active and the decay time constant is expected to be shrunk by a factor equal to the amplification of the non inverting gain stage. Intermediate situations are expected for $0 < k < 1$. The results are reported in Fig. 6 where the observed output waveforms are reported for some values of parameter k. In the setup used $R_f = 1$ GΩ, $C_f = 1.5$ pF, $R1 = 470\ \Omega$, and $R2 = 5$ kΩ. It can be seen that the experimental data confirm qualitatively the theory. But an unexpected effect is also seen: as k is increased a loss in signal amplitude arises, as shown in Fig. 7. Furthermore the shrunk decay time observed for $k = 1$ is not as short as predicted by (3). These effects can be explained by taking into consideration the stray capacitance of R_f, which introduces the "built-in" time constant τ_s of the resistor R_f, evidenced in Fig. 8. Such a stray capacitance is expected to be in the order of a fraction of pF, but luckily enough it does not introduce an extra pole in the circuit transfer function. In fact its voltage drop is proportional to the voltage drop across the capacitance C_f, because the right ends of C_f and R_f do coincide with the input and output of the shown noninverting gain stage. Thus, the voltage across the stray capacitance is $(1 + k \times R2/R1)$ as high as that across C_f. Such a constraint takes one degree of freedom away and is such that the two capacitances introduce one pole only in the transfer function of the circuit. By making the necessary calculations, one obtains the relationship

$$v_o = -iR_f \left(\frac{1+R2/R1}{1+k \times R2/R1}\right) \frac{1}{1+s\left(\tau_S + \dfrac{R_f \times C_f}{1+k \times R2/R1}\right)} \tag{6}$$

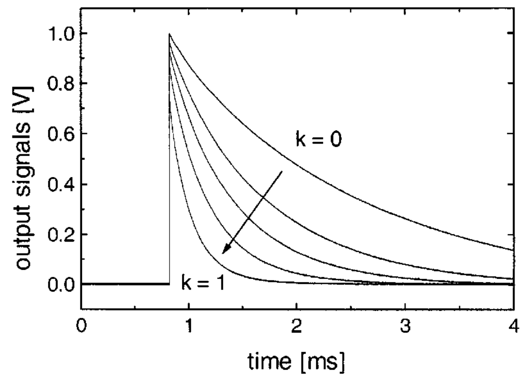

Fig. 6. Cold discharge mechanism action. The decay time constant is reduced by the second feedback loop as parameter k is increased while no change is done for R_f and C_f.

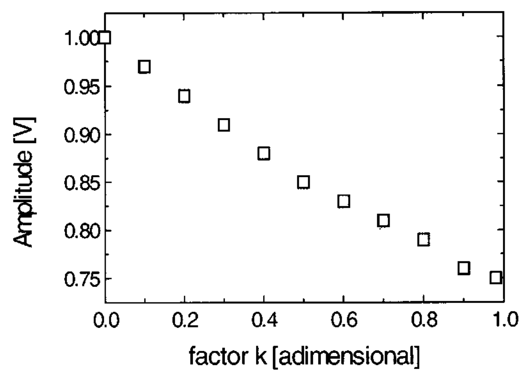

Fig. 7. Loss in amplitude of the output waveform versus parameter k.

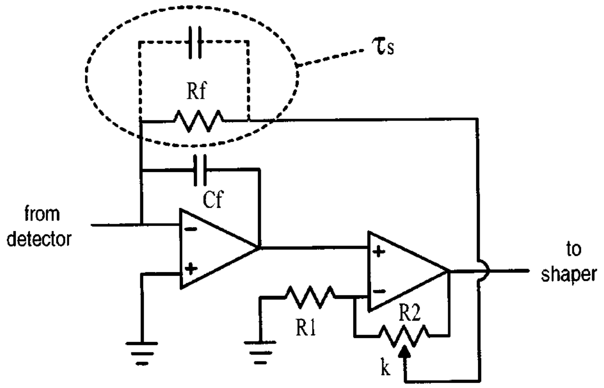

Fig. 8. The feedback resistor has a "built-in" time constant τ_s due to parallel capacitive parasitism.

where v_o is the output voltage of the circuit and i is the input current signal coming from the detector. From (6) it can be seen that the transfer function has actually a single pole. The pole time constant, which is the decay time constant of the preamplifier, has an inferior limit given by τ_s. In fact, the pole time constant tends to τ_s as k tends to infinity. Such an effect could not be seen from (3) and explains why the decay time constant is not reduced as much as expected from (3) when $k = 1$. From (6), it can also be seen that, while injecting a charge Q at the input through a current delta-like pulse, the "high-frequency" output signal amplitude is

$$V_O = \frac{Q \times R_f \left(\dfrac{1 + R2/R1}{1 + k \times R2/R1} \right)}{\left(\tau_S + \dfrac{R_f \times C_f}{1 + k \times R2/R1} \right)}. \tag{7}$$

It can be easily verified that, for $\tau_s \rightarrow 0$ (7) yields $V_O = Q/C_f \times (1 + R2/R1)$ as is expected. From (7), the loss of gain observed in Fig. 7 is thus explained. Again, it is due to stray time constant τ_s. By fitting (7) to the data of Fig. 7, an estimate of τ_s can be derived, or $\tau_s = 57$ μs. Because $R_f = 1$ GΩ, this yields $C_S = 0.057$ pF. In order to reduce the effect of such stray capacitance, R_f could be physically realized by using a series of lower value resistors. In fact, the stray capacitance of a discrete resistor is mostly due to the geometry of its pads and has a modest dependency on its value.

The circuit has been also tested for risetime (see Fig. 9 for an example) and noise (Fig. 10) performances. In the measurements, no cooling system has been used to reduce the noise of the input FET and the feedback resistor, and the detector has been simulated with a capacitor. The noise measurements have been performed by using a calibrated test capacitance to inject a known amount of charge at the input and by simulating the detector with 33–pF and 66–pF capacitances connected from the input node to ground. A quasi-Gaussian shaper has been used for signal conditioning. The FWHM fluctuation seen at the shaper output has been evaluated by Gaussian fitting of the spectral line acquired with a pulse-height analyzer for nuclear spectroscopy. Eventually, such a measurement has been referred to the input by dividing by the chain gain and converted in eV FWHM (resolution in Fig. 10) by using the HPGe energy-to-charge conversion factor (2.98 eV/pair). In Fig. 10, the measurements, the

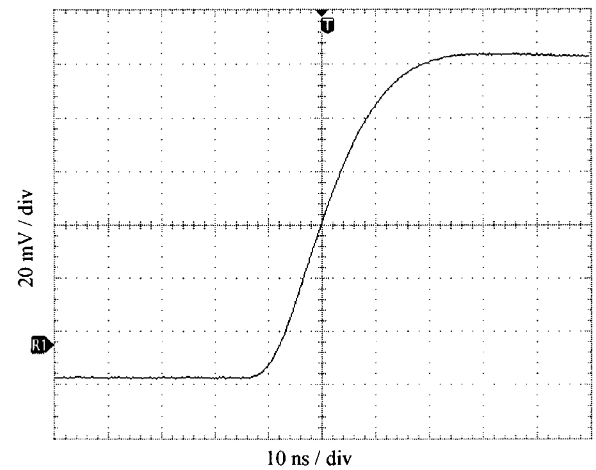

Fig. 9. Oscilloscope track of the signal seen at the output of the tested preamplifier. The detector capacitance is 66 pF and the rise time is 20.6 ns.

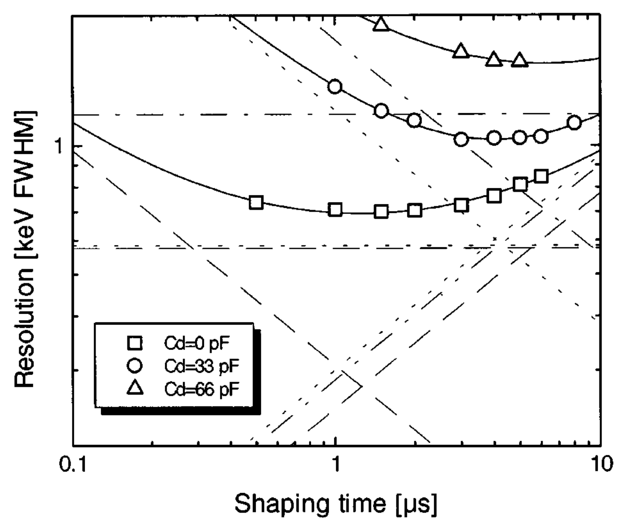

Fig. 10. Noise of the preamplifier. The preamplifier has been tested with 0–pF, 33–pF, and 66–pF detector-capacitance values. The single points are measurements, the solid curves are fittings. The noise components returned by the fitting procedure are also shown. To obtain the resolution in electrons r.m.s divide the eV FWHM by 7.02.

fitting curves, and the three noise components returned by the fitting algorithm [6] are shown. The descending component is series white noise, the flat one is series $1/f$ noise, the ascending one is parallel white noise. It can be seen that as the input capacitance is increased, the series-white-noise component grows substantially whereas the parallel-white-noise component does not, which confirms expectations. In fact, it is well known that the input capacitance influences the series noise contributions only [4]. Note that, in case of an amplifier without the additional feedback amplification stage, but with an identical discharge time, the lower-value feedback resistor would yield a G-times larger parallel-white-noise contribution ($G \approx 8$ here). The principal experimental results are summarized in Table I. It can be seen that an excellent performance is obtained, both in terms of risetime and of noise. Note that with a detector capacitance of 33 pF, which is a realistic value for HPGe segmented detectors, a risetime of 14 ns and an electronic noise of 1.03 keV FWHM are obtained, which looks adequate for the considered applications.

TABLE I
PERFORMANCE OF THE PREAMPLIFIER

Cin [pF]	Shaping time [μs]	keV FWHM - Ge	Rise time [ns]
33	1	1.373	14
	1.5	1.203	
	2	1.142	
	3	1.032	
	4	1.038	
	5	1.041	
	6	1.047	
	8	1.126	
66	1.5	1.893	21
	3	1.642	
	4	1.57	
	5	1.562	

V. CONCLUSION

A cold discharge mechanism for fast low–noise charge preamplifiers has been proposed. The circuital topology has been analyzed theoretically and tested experimentally, finding a good agreement with expectations. The obtained results are satisfying and tests in a real cryogenic apparatus are under way. The price to be paid is that three electrical feed throughs between the FET$R_f - C_f$ cryogenic ensemble and outer environment are required rather than two, due to the necessity to close the second feedback loop. Alternatively the whole FET$R_f - C_f$ ensemble could be used warm outside the cryostat, in which case a single feedthrough would be required. Such latter arrangement is under study.

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