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MODELING DETECTOR FOR RADIATION MONITORING SYSTEMS

The detector characteristics are determined mainly by the physical properties of the semiconductor crystal as a sensitive element of the primary converter, as well as by the features of the electrical signal registration process.

The process of registering ionizing radiation (IR) consists of converting a non-electrical quantity that characterizes it into an electrical signal. In other words, one type of energy – IR energy – is converted into another, more convenient for processing and storing information. A current or voltage pulse occurs in the radiation sensor directly as a result of ionization of its active medium – a semiconductor; this pulse carries extensive information. First of all, it is correlated with the moment in time of the nuclear process. In addition, the pulse marks the fact of radiation emission within the solid angle at which sensor is visible from source. The pulse amplitude often serves as a measure of the energy loss of radiation in sensor. The pulse shape differs for different types of radiation, as well as for different areas and angles of radiation penetration into the sensor.

Most of the information transmitted by pulses is characterized by a continuous spectrum: a pulse can appear at any time with different amplitudes. In addition, electrical signals from the detector generally arrive against a background of interference, which significantly reduces the reliability of the transmitted information. Interference can be caused by parasitic electrical signals (circuit element noise, external interference) or extraneous sources of IR. The reliability of measurements increases if the signals have qualities different from interference.

The paper proposes a structural diagram and creates a multichannel digital amplitude analyzer that provides selection of the input signal by the pulse shape at high loads, which is not available in existing devices today. The use of such an analyzer made it possible to increase the energy resolution, stability of the detector with an even greater increase in the input load of the measuring path.

Keywords: registration of ionizing radiation, nuclear process time, continuous spectrum, energy resolution

Introduction. Currently, almost all industries and many branches of science use IR sources. Nuclear power plants, gamma installations of various capacities, flaw detectors, counters and many other equipment are widely used in the defense industry, medicine and agriculture. However, the most important branch of IR use in Ukraine after the elimination of the combat nuclear potential is nuclear energy [1]. The country has five nuclear power plants (NPP) with two types of reactors, which generate about 40% of the country's total electricity [2].

In this regard, dosimetry problems are becoming increasingly important, which today has become an independent scientific and technical direction of nuclear physics. Dosimetry essentially solves the problems of connecting physical quantities with the expected radiation effects of IR use. The main task of dosimetry - identifying IR sources that pose a danger to the environment and humans - is today solved using a variety of technical recording means with varying degrees of efficiency. A comparative analysis of such means and methods of their application for registration and dosimetry is presented in this section [3]. In addition, the existing variety of terms and quantities in this field requires some clarification to ensure the reliability of the presented research results.

Main part

Analysis of initial parameters crystal. Semiconductor material for use in ionizing radiation sensors must meet the previously specified conditions and be characterized by a set of the following functional qualities [3].

To obtain a signal in the external electric circuit, in which the sensor is included, the charge carriers must move quickly enough in the electric field of crystal, and to obtain a signal linear in energy, its losses on the way to the electrodes must be minimal (absent). In the general case, this means that it is necessary to ensure the highest value of mobility both types of charge carriers (with their unipolar collection - mobility of electrons) and the minimum concentration of traps that capture carriers in the process of their drift to the electrodes.

To register small signals, it is necessary to have minimal loss currents at sufficiently high voltages applied to the sensor. This means that semiconductor material must be highly resistive. It must effectively absorb X-ray or gamma radiation quanta in a sufficiently small volume. Therefore, the greater the density of material and its atomic number Z, the greater absorption cross-section on an individual atom and the ability of the material to slow down ionizing radiation as a whole. This quality is fundamental for ensuring significant sensitivity of the sensor with its sufficiently small dimensions. This quality plays a particularly important role in the case of high-energy gamma-ray spectrometry, since the efficiency of recording high-energy gamma-quanta strongly depends on the thickness of sensor and the properties of active medium (cross-section of the photoelectric effect is proportional to Z^5).

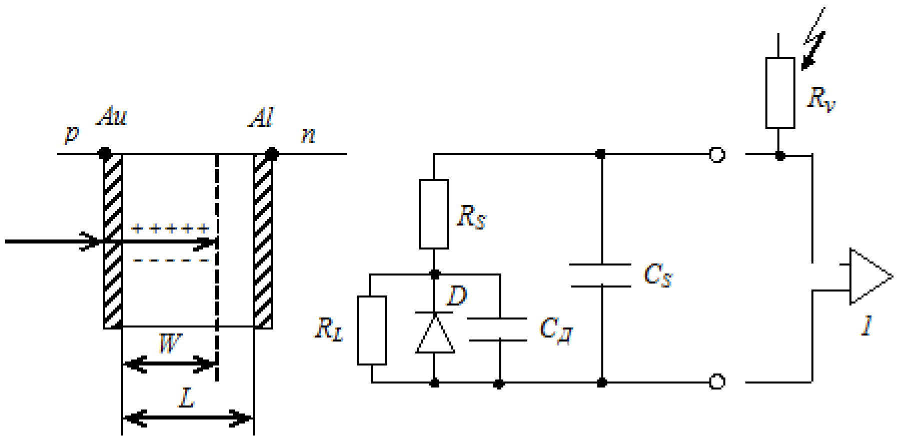
In some cases, it is also important to be able to create blocking contacts that would prevent the transfer of free carriers into the working area of sensor, i.e. such contacts when the positive electrode does not inject holes into the semiconductor, and negative electrode does not inject electrons. The problem of creating blocking contacts is solved, as a rule, either by selecting appropriate metals as contact materials, or by creating n^+- and p^+-regions in the semiconductor by diffusion or ion doping with appropriate impurities. In this case, n^+-contact is used as a positive electrode, and p^+-contact as a negative one. Such structures are called pin-structures [4, 5].

The listed requirements apply both to the quality of sensor material and to its design.

The structural diagram of the semiconductor sensor and circuit diagram of secondary converter preamplifier, i.e. the entire detector, are shown in fig. 1.

To ensure the directed movement (drift) of the charge carriers created by IR, a forward bias voltage is applied to the contacts of sensor *D* through the load resistance R_V (fig. 1.a). The resulting pulse voltage drop $U = \frac{Q}{C}$ is in most cases not proportional to the energy lost by γ-quantum. The difference in the charge collection time leads to a spread in the duration, and therefore in the amplitude of pulses. In addition, capacitance of the sensor itself does not remain constant. Therefore, it is necessary to use a charge-sensitive preamplifier 1 in the detector circuit (fig. 1.b).

The equivalent circuit of the semiconductor sensor contains, in addition to the diode *D* itself, depletion zone capacitance $C_Д$, parasitic capacitance C_S, leakage resistance R_L and "trajectory" resistance R_S. The latter is a combination of the resistances output electrodes. The diode capacitance also depends on the voltage and quality of crystal.



a b

Figure 1 – Structure of the sensor (a) and equivalent circuit (b) for the semiconductor detector: designations – in the text

This dependence can be approximately represented as [6]:

$$C_d = 21 \cdot 10^3 A(\rho U_b)^{-\frac{1}{2}}, \text{ pF}, \tag{1}$$

where A – is the area of sensor, sm^2;
ρ – is the specific resistance of the semiconductor material;
U_b –is the blocking voltage.

The given dependence can be used for comparative evaluation of the sensor switching modes.

One of the important characteristics of the sensor is the level of parasitic components of the signal - noises not related to the physical processes of interaction of the crystal with IR. The noise level determines the minimum threshold for recording IR energy.

The conversion of the energy lost by particle in sensor into an electrical signal of the corresponding amplitude occurs with an accuracy characterized by the resolution of system[6]. The latter depends on many factors, in particular, on the properties of the amplifier. Indeed, since the amplitude of signal generated by semiconductor sensor is small, the distortion of amplitude spectrum is caused, first of all, by modulation by noise pulses arising in it and in the resistances. Chaotically adding up with useful signals, the noise "washes out" original amplitude spectrum. The distribution of noise by amplitude is Gaussian:

$$p(U) = \frac{1}{\sigma\sqrt{2\pi}} \cdot e^{-\frac{(U_i - \overline{U})^2}{2\sigma^2}}, \tag{2}$$

where σ^2 – is the dispersion or mean square deviation of the amplitude U_i from the mean value $\overline{U}$.

Let us assume that all other factors distorting the signal amplitude spectrum are negligibly small compared to influence of noise and register monochromatic charged particles leaving all the energy in the sensor. In this case, the measured signal amplitude spectrum (fig. 2) is also determined by expression (2). However, now $\overline{U}$ – is the average signal amplitude, and σ is determined by noise, where $\sqrt{\overline{U_ш^2}} = U_ш$ is equal to the root-mean-square noise voltage . The width of the curve at half-height is called the resolution $\frac{1}{2}\Delta$ (FWHM). Substituting the value into equation (2), it is easy to obtain $p(U) = \frac{1}{2}p(\overline{U})$. Having measured the resolution in energy units (in electron volts), it is possible to determine what part of the energy corresponds to the noise level recalculated to the input of a given amplifier [7].

The absolute value of the capacitance $C_Д$, as well as the parasitic capacitance C_S, largely determines the noise level, and with it the energy resolution of charge-sensitive preamplifier. The current flowing through leakage resistance R_L is another source of noise, which also leads to deterioration in the energy resolution.

For the subsequent devices of the detector to operate - the amplitude analyzer, discriminator, coincidence circuit - an amplifier with a high gain is required. Usually, the amplifier consists of two separate units: the preamplifier and the main amplifier. This division is due to the desire to minimize the input capacitance *C*, which affects the resolution, while the preamplifier is located next to the sensor. The signal, amplified by the first unit to a level at which the noise of the subsequent amplifier has practically no effect, is transmitted to the second unit via a matched cable. Particular attention should be paid to obtaining a minimum of noise in the preamplifier [8, 9].

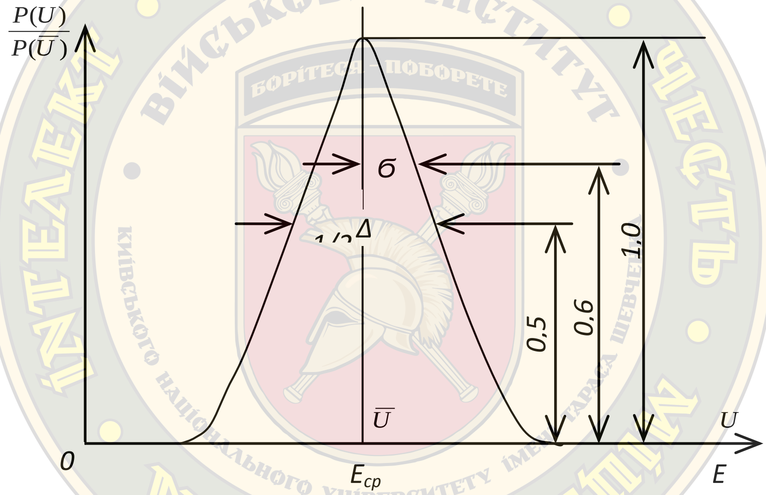


Figure 2 – Energy line expansion due to noise

To analyze the noise, let us consider in more detail the equivalent circuit of preamplifier. Noise, like the signal, can be expressed numerically in units of voltage, charge, or energy. With energy losses *E*, electron-hole pairs are formed $N = \frac{E}{W_p}$, giving a charge *Q* on the total input capacitance *C*. If $\tau_{вх} = RC$ is large compared to the charge collection time, then the signal amplitude $U' = \frac{Q}{C}$. For further consideration, we will take into account the action of the forming circuits. As a result of passing through the differentiating and integrating circuits with $\tau_и = \tau_д = \tau$ (this case is often used in practice), the signal will decrease by a factor of *e* = 2,72, i.e. will be $U = \frac{Q}{C \cdot e}$. Similar reasoning can be carried out for noise, only starting with voltage $U_ш$, then moving on to charge $Q_{ш} = U_{ш} C \cdot e$, the number of charge $N_{ш} = \frac{Q_{ш}}{q}$ carriers, and finally to the equivalent noise energy [10]:

$$E_{ш} = \frac{U_{ш} C \cdot e \overline{W_p}}{q}.$$

Often, when evaluating the noise properties of amplifiers, the signal-to-noise $\eta = \frac{U}{U_{ш}}$ ratio is used. Knowing and the signal, it is easy to determine $U_{ш}$ and $\frac{1}{2}\Delta$.

The spectral density of the parallel noise current is:

$$\frac{\bar{i}_p^2}{\Delta f} = 2qI + \left(\frac{4kT}{R_p}\right), \tag{3}$$

where I – is the sum (in absolute value) of all currents acting parallel to the sensor; R_p – is the resistance of all resistors connected parallel to the sensor; Δf – is a fragment of the spectral characteristic; T – is the absolute temperature.

This spectral density can be expressed by one equivalent noise resistance R_p, the value of which is determined by the relation:

$$\frac{1}{R_p} = \frac{qI}{2kT} + \frac{1}{R_S}. \tag{4}$$

Parallel noise is frequency independent, but the voltage it creates on input capacitance C, just like the input signal, depends on frequency inversely proportional to:

$$\frac{\bar{u}_p^2}{\Delta f} = 4kT \frac{1}{R_p} \frac{1}{(\omega C)^2}. \tag{5}$$

Another source of input stage noise is determined by the input amplifier device and its amplification principle. This noise does not depend on the input elements, so it is convenient to take it into account by the equivalent noise resistance R_S, connected in series with the amplifier input. For a field-effect transistor, the series equivalent noise resistance is $R_S \approx \frac{1}{S}$, where S – is the slope of the transistor's input characteristic. The intensity of the series noise is also frequency-independent and is:

$$\frac{\overline{u}_s^2}{\Delta f}=4kTR_S \cdot \quad (6)$$

In some cases, especially when registering X-rays, the noise component of transistors of type $\frac{1}{f}$ plays a significant role. This noise can be determined by the formula:

$$\frac{\overline{u}_s^2}{\Delta f}=\frac{A_f}{f^{\alpha}} , \quad (7)$$

where A_f – is a constant coefficient depending on the transistor manufacturing technology $\alpha \approx 1$.

The total noise voltage of the noise sources at the amplifier input is equal to:

$$\overline{U_{ш}^2} = (4kT\frac{1}{R_p}\frac{1}{\omega^2 C^2} + 4kTR_s + \frac{A_f}{f})\Delta f = N(\omega)\Delta f , \quad (8)$$

where $N(\omega)$ – is spectral density of input noise; Δf – is the narrow differential frequency bandwidth $f = \frac{\omega}{2\pi}$; .

The frequency response $K(\omega)$ of spectrometric amplifiers extends from low to high frequencies and the noise $U_{ш}$ level at the amplifier output is determined by the integral expression:

$$\overline{U}_{ш}^2 = \frac{1}{2\pi}\int_0^{+\infty} \left|N(\omega)\right|\left|K(\omega)\right|^2 d\omega . \quad (9)$$

The limiting effect of the amplifier bandwidth $K(\omega)$ also affects the signal shape. The dependence of amplifier output signal on time can be determined using the inverse Fourier transform formula:

$$S_2(t)=\frac{1}{2\pi}\int_{-\infty}^{+\infty} S(\omega)K(\omega)e^{j\omega t}d\omega .$$

Selecting the best frequency response of spectrometric path in order to obtain the maximum signal-to-noise ratio is the essence of optimal filtering [10].

According to this theory, the square of the maximum possible signal-to-noise ratio is equal to [11]:

$$(\eta_{макс}^{\infty})^2 = \frac{2}{\pi}\int_0^{\infty} \frac{U^2(\omega)}{U_{ш}^2(\omega)} ,$$

where $U(\omega)$ and $U_{ш}(\omega)$ – are the spectrum of the signal and noise at the amplifier input, respectively.

It has been theoretically shown that the maximum signal-to-noise ratio in this case is achieved with equal integration and differentiation $\tau_{CR}=\tau_{RC}=\tau$ time constants. In this case, the noise level is minimal at some optimal time constant τ_0:

$$\tau_0=C\sqrt{R_S R_P}\ . \tag{10}$$

Then noise level at the amplifier output is determined by integral expression:

$$\overline{U^2_{ш}}=\frac{1}{2\pi}\int_0^{\infty} N(\omega)\frac{\omega^2\tau^2}{(1+\omega^2\tau^2)^2}d\omega=4kT\frac{R_S}{8\tau}+\frac{4kT\tau}{8C^2R_P}+\frac{A_f}{2}. \tag{11}$$

As can be seen from (11), serial noise depends inversely, parallel noise – proportionally, and noise of type $\frac{1}{f}$ does not depend at all on τ. The minimum noise $\tau=\tau_0$ value at is equal to $\overline{U^2_{ш.мин}}=\frac{kT}{C}\sqrt{\frac{R_S}{R_P}}$ (excluding noise of type $\frac{1}{f}$). Considering that with CR-RC formation the amplitude of the output signal does not depend on τ, the minimum noise corresponds to the maximum signal-to-noise ratio:

$$\eta^{RC}_{макс}=\frac{S_{2макс}}{U_{ш.мин}}=(\frac{2}{e})\frac{Q}{\sqrt{4kTC}}\sqrt[4]{\frac{R_P}{R_S}}=(\frac{2}{e})\eta_{\infty}. \tag{12}$$

This formula shows the ratio of output voltage amplitude to root-mean-square voltage of the output noise. The input signal of the spectrometric amplifier is a charge Q or energy E released by ionizing radiation in the sensor, so in practice it is common to express noise level also in units of charge or energy. Taking $\eta^{RC}_{макс}=1$, we find the equivalent root-mean-square noise charge for CR-RC formation:

$$\sigma^{RC}_q=(\frac{e}{2})\sqrt{4kTC}\sqrt[4]{\frac{R_S}{R_P}}=1{,}36\sigma_q, \tag{13}$$

where σ_q is the minimum possible noise charge.

The presented model of the primary converter allows, taking into account the real properties of the crystal, to calculate the dependences of the energy equivalent noise on the time constant of the input stage of the preamplifier (fig. 3).

When designing the analyzer, a comparison was made of two methods for determining the amplitude, differing in resolution and computational costs: first is determining difference between local minima and subsequent maximum; the second is determining difference between successive maxima.

Figure 3 shows an out-of-scale image of a typical segment of the preamplifier output signal, containing two pulses [12, 13]. The duration of the leading edge of individual pulse is – $\Delta t=20$ ns, constant of the trailing edge is – $\alpha=2020$ µs. The amplitude is expressed in ADC discretes. The simplest first method for determining the pulse amplitude in such an input sequence is finding the difference between the local minimum and the subsequent local maximum. Let us introduce the following notations for the input signal: $f_1(t)$ – function describing the decay of the first pulse; $f_2(t)$ – function describing the decay of the second pulse.

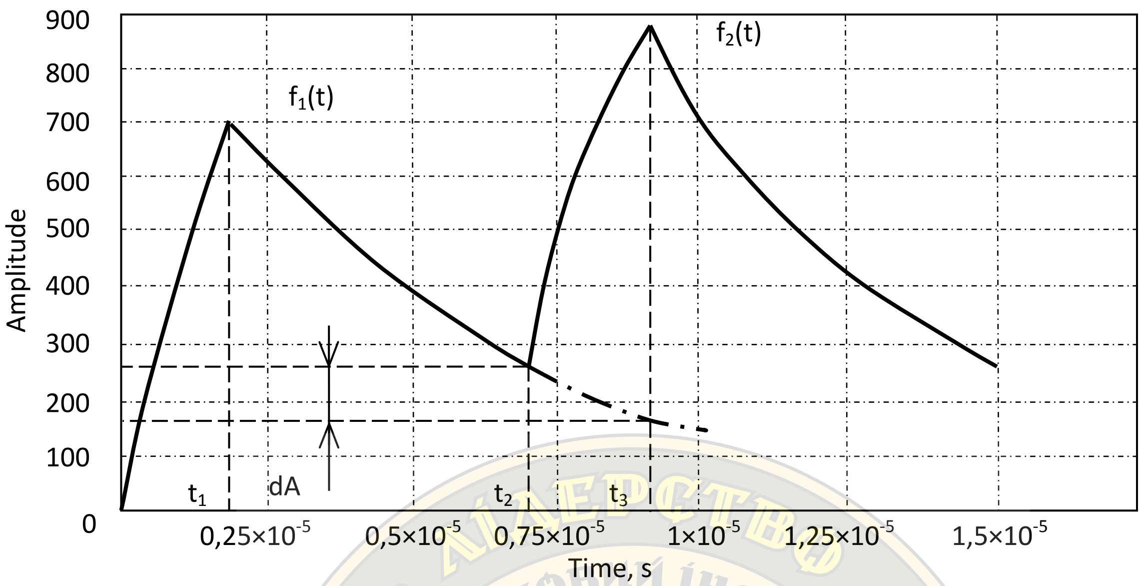


Figure 3 – Analysis of structure typical fragment input pulses of preamplifier

Then, according to the first method, the amplitude of the second pulse is found as:

$$A^{*} = f_2(t_3) - f_1(t_2), \tag{14}$$

where t_3 and t_2 – are the moments of time indicated in figure 3.

True pulse amplitude:

$$A = f_2(t_3) - f_1(t_3). \tag{15}$$

Hence, the error in determining amplitude is:

$$dA = A - A^{*} = f_1(t_2) - f_1(t_3). \tag{16}$$

Since the pulse decay constant is determined by the physical characteristics of the sensor and preamplifier, it has a constant value for all input pulses [13]. Then absolute value of the input signal at a given time t_1 can be expressed as follows:

$$f_1(t_1) = \sum_{n=0}^{\infty} A_n e^{-\alpha \tau_n}, \tag{17}$$

Similarly for the moments of time:

$$f_1(t_2) = \sum_{n=0}^{\infty} A_n e^{-\alpha(\tau_n + \delta t)} = e^{-\alpha \delta t} \cdot f_1(t_1), \tag{18}$$

$$f_1(t_3)=\sum_{n=0}^{\infty} A_n e^{-\alpha(\tau_n+T)}=e^{-\alpha T}\cdot f_1(t_1), \qquad (19)$$

where $\delta t=t_2-t_1$, $T=t_3-t_1$.

Substituting (18) and (19) into (16), we obtain:

$$dA=f_1(t_1)\cdot\left(e^{-\alpha\delta t}-e^{-\alpha T}\right). \qquad (20)$$

Because

$$\delta t=(t_3-t_1)-(t_3-t_2)=T-\Delta t, \qquad (21)$$

where $\Delta t=t_3-t_2-$ is the duration of the leading edge, then:

$$dA=f_1(t_1)\cdot e^{-\alpha T}\cdot\left(e^{-\alpha\Delta t}-1\right). \qquad (22)$$

From expression (22) it is evident that the error in determining the amplitude using the first method depends on last maximum of the signal $f_1(t_1)$ (sum of amplitude last pulse and the dips from previous pulses), the time between the last and determined pulses T and the duration of the leading edge of determined pulse Δt [14, 15]:

$$dA=\Phi\left(f_1(t_1),T,\Delta t\right). \qquad (23)$$

For a qualitative analysis of the obtained dependence on the spectrometer load value (the number of pulses recorded in 1 s), we will assume that pulses of the same amplitude A are received at the input. Then the mathematical expectation at ADC input:

$$M_{ампл}=\sum_{n=0}^{\infty} A\cdot e^{-\alpha T_M n}=\frac{A}{1-e^{-\alpha T_M}}, \qquad (24)$$

where T_M – mathematical expectation of the pulse repetition period; $T_M=1/F_3$, where F_3- loading frequency.

Substituting (23) into (21), we obtain:

$$dA_M=A\frac{e^{\alpha\Delta t}-1}{e^{\alpha T_M}-1}.$$

Figure 4 shows a graph of the dependence mathematical expectation error on the loading frequency for A=700, $\Delta t=20\cdot10^{-9}$ s, $\alpha=20\cdot10^{-6}$ s. It is evident from the graph that the magnitude of the error at a loading frequency of about 10^6 c^{-1} can reach 1-2%.

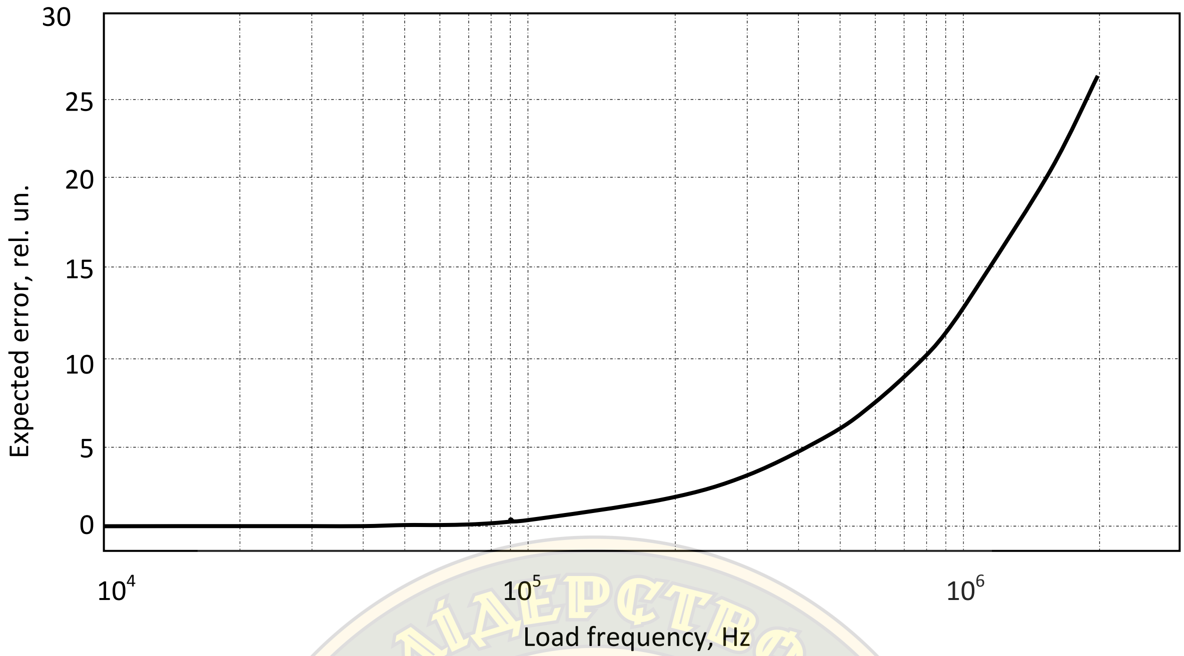


Figure 4 – Dependence of the expected error on the loading frequency when determining the pulse amplitude by directly measuring it between minimums and maximums

This error can be avoided if we assume that the coefficient depends on the sensor and preamplifier, i.e. is a constant value for a specific spectrometer and can be determined empirically. In particular, using the notations in figure 4, we obtain:

$$\alpha = \frac{\ln(f_1(t_1)) - \ln(f_1(t_2))}{\delta t}. \qquad (25)$$

Then the formula for the second method of determining the amplitude takes the following form:

$$A = f_2(t_3) - e^{-\alpha T} f_1(t_1). \qquad (26)$$

The price paid for increasing the accuracy in this method is an increase in the volume of calculations. This should be taken into account when constructing spectrometers operating in real time.

Mathematical expectation of the amplitude maximum sample:

$$M_s = A - \frac{\Delta}{2} = A\frac{1+e^{-\alpha t}}{2}. \qquad (28)$$

The variance of the error in amplitude will be:

$$D_S = \frac{\Delta^2}{12} = \frac{A^2}{12}\left(1 - e^{-\alpha e}\right)^2. \qquad (29)$$

The dependence of the maximum sample amplitude on the sampling frequency is shown in figure 5, where for definiteness the true amplitude input pulses is taken to be equal to 700 ADC quantization levels. The graph of the mathematical expectation of the maximum sample amplitude shows the location of the peak on energy spectrum. With an increase in the sampling frequency from 10 to 100 MHz, the magnitude of peak decrease from the true value (700) decreases from 1% to 0.1%. The deviation limits characterize the peak width in spectrum, which affects the resolving properties of spectrometer.

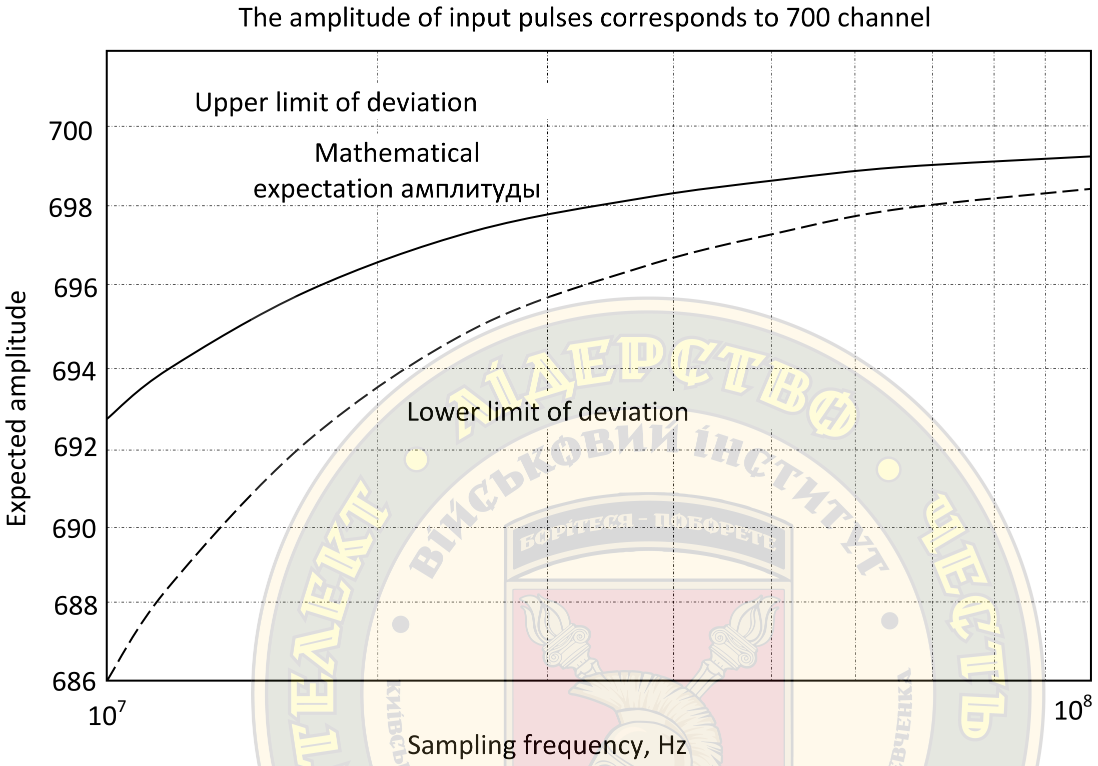


Figure 5 – Dependence of sampling error on its frequency

For a more complete comparison of the resolution methods described above, the process of measuring pulse amplitude was simulated based on the model developed above [12,13].

At same time, the use of digital signal processing technique proposed in this work allows both methods to obtain an energy resolution not exceeding 40 keV. This is sufficient to create a dosimeter with compensation of the energy dependence sensitivity (EDS). In addition, these results show that second method is optimal for creating high-resolution spectrometers - less than 10 keV.

Thus, a comparison of the two methods for determining the amplitude showed that at a low loading frequency ($<10^5$ s^{-1}), first method is preferable: it is characterized by a smaller volume of calculations with the same accuracy characteristics. At high loading frequencies (about 10^6 s^{-1} and more), preference should be given to the second method, which retains its accuracy and resolution.

Thus, the obtained model allows us to estimate the effect of sampling frequency on the magnitude of shift peaks of the energy spectrum and resolution properties of the spectrometer when choosing parameters of its hardware implementation.

Therefore, for the effective use of ADC bit depth, the constant component must be subtracted from its input signal. The dynamic range of signal change must be matched with the diagnostic range of ADC. To solve this problem, a corresponding circuit of the digital spectrometer is proposed, shown in fig. 6.

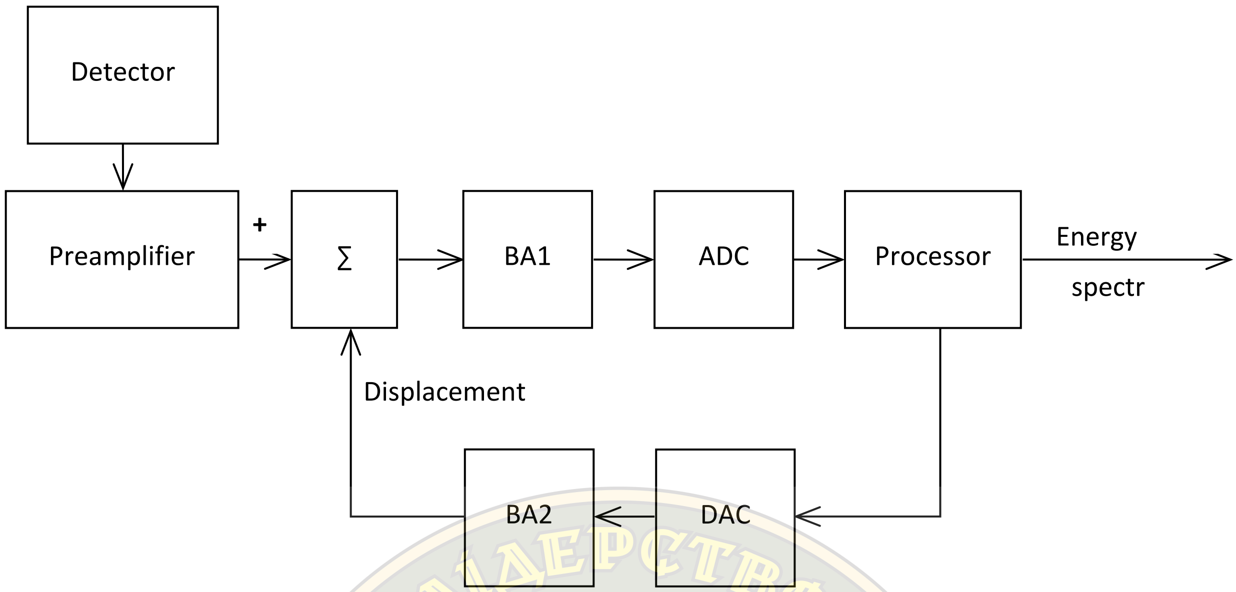


Figure 6 – Block diagram of digital spectrometer: Σ – adder; BA – buffer amplifier; DAC – digital-to-analog converter

In this diagram, the processor, based on the analysis of signals from ADC, sets the required gain factors of the first buffer amplifier and offsets for adder.

The maximum gain factor in BA1 is selected so that when measuring the minimum background, the signal at ADC input corresponds to its dynamic range. The minimum value of the gain factor should match the input signal with the dynamic range of ADC at the maximum frequency of loading pulses maximum possible amplitude and zero offset value.

For example, in order to improve the energy resolution, the sensor uses such a feature of CdZnTe as a large difference in the mobility of electrons and holes. Thus, the use of digital signal processing allows us to achieve a result similar to changing the configuration of the electrodes. This allows us to create dosimeters that solve an important practical problem – compensation for the energy dependence of sensitivity.

Conclusions. The paper presents a model of primary transducer – gamma-radiation sensor. It is based on the following properties of a semiconductor crystal: maximum quantum efficiency; maximum carrier mobility; minimum density of structural defects; maximum values of specific resistance and density. The combination of these properties provides significant sensitivity of the sensor with minimum crystal dimensions. The inconsistency of such a combination must be eliminated both in the process of crystal manufacturing (for example, a high-resistance crystal can be obtained by simultaneously using cleaning, components and compensating doping) and by subsequent processing using the methods proposed in this paper (thermal field method, ionization annealing).

The model of the primary transducer (sensor) allows calculating dependence of the energy equivalent of noise on the properties input stage of preamplifier, taking into account the real properties of the crystal. It is shown that:

– increase in the crystal volume, bias voltage and sensor capacitance increases the noise level;

– results of the analysis in relation to CdZnTe crystals used in this work indicate the possibility of sensor operating without cooling.

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МОДЕЛЮВАННЯ ДЕТЕКТОРА ДЛЯ СИСТЕМ РАДІАЦІЙНОГО КОНТРОЛЮ

Характеристики детектора визначаються головним чином фізичними властивостями кристала напівпровідника як чутливого елемента первинного перетворювача, а також особливостями процесу реєстрації електричного сигналу.

Процес реєстрації іонізуючого випромінювання (ІВ) полягає у перетворенні неелектричної величини, що характеризує його, електричний сигнал. Інакше висловлюючись, у своїй перетворюється один вид енергії – енергія ІВ – в інший, зручніший для обробки та накопичення інформації. У датчику випромінювання виникає імпульс струму чи напруги у результаті іонізації його активного середовища – напівпровідника, цей імпульс несе велику інформацію. Насамперед, він корелюється з моментом часу ядерного процесу. Крім того, імпульс наголошує на факті випромінювання радіації в межах тілесного кута, під яким датчик видно від джерела. Амплітуда імпульсу часто є мірою енергетичних втрат випромінювання в датчику. Форма імпульсу відрізняється для різних видів випромінювання, а також для різних областей та кутів попадання випромінювання в датчик.

Більшість інформації, передана імпульсами, характеризується безперервним спектром: імпульс може виникнути у час з різною амплітудою. Крім того, електричні сигнали з детектора в загальному випадку надходять на тлі перешкод, що значно знижують достовірність інформації, що передається. Перешкоди можуть бути спричинені паразитними електричними сигналами (шуми елементів схеми, зовнішні наведення) або сторонніми джерелами ІВ. Надійність вимірювань збільшується, якщо сигнали мають властивості, відмінні від перешкоди.

У роботі запропоновано структурну схему та створено багатоканальний цифровий амплітудний аналізатор, що забезпечує селекцію вхідного сигналу за формою імпульсу при великих завантаженнях, що недоступно приладам, що існують. Застосування такого аналізатора дозволило збільшити енергетичну роздільну здатність, стабільність роботи детектора при ще більшому збільшенні вхідного завантаження вимірювального тракту.

Ключові слова: реєстрація іонізуючого випромінювання, час ядерного процесу, безперервний спектр, енергетичний дозвіл