ELECTRONICS AND ELECTRICAL ENGINEERING
ISSN 1392 – 1215 2007. No. 5(77)
ELEKTRONIKA IR ELEKTROTECHNIKA

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MIKROELEKTRONIKA

Leakage Current Compensation for the 0.13 µm CMOS Charge Sensitive Preamplifier

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Introduction

The active hybrid pixel detector (APD) can be a valuable solution in front-end electronics for radiation sensors. Detection systems with high sensor pixilation can benefit from low power, low parasitic, high front-end channel density, and low cost per channel. In addition, the APDs are characterized by good radiation tolerance [1-3] and can integrate large amount of additional signal processing and functions in analog, mixed-signal and digital domains, offering further advantage in terms of power.

Submicron CMOS technologies ensure radiation hardness required by detector readout systems in present high-energy physics experiments as well as in other application fields [2-5]. CMOS is the most widely used technology for readout and control of imagers.

CMOS low-noise preamplifiers for the active hybrid pixel detector were presented in [6]. APD are based on Si, GaAs, CdTe sensors connected to dedicated front-end electronics chips using the bump and flip-chip technology. A typical electronics read-out system consists of a charge sensitive preamplifier (CSP), realized with an amplifier with a capacitive feedback, a shaping filter, and an ADC. The incoming particles generate charge in the detector which is being integrated by the preamplifier and filtered with the shaper, in order to obtain the proper time evolution. Then, ADCs convert the signal amplitude for the successive storing in a computer. A general description of the electronic readout system which is based on a CSP was previously reported in [6-7].

In this paper the authors attempt to derive some design aspects for the CMOS preamplifier of hybrid pixel detectors used in particle tracking, operating in the nanoampere region.

The experimental simulation results obtained with CSP for APD imaging implemented in CMOS 0.13 µm are reported. In that work, we are simulated a scheme for producing a high-value feedback device and leakage current compensation circuits.

Analysis of Mathematical Formulae of Equivalent Noise Charge

Table 1 displays the mathematical formulae used to calculate the contributions from the different sources of noise [2, 8]. The abbreviation in this table: ENC - equivalent noise charge, q - electron charge, τ_p - peaking time, C_{fb}, C_{det} and C_{in} - the feedback, detector and input transistor capacitance, respectively, R_S - the serial resistance, R_P - the parallel resistance of the sensor bias resistor and the preamplifier feedback resistor, $I_{leakage}$ - detector leakage current, g_{mTA1}, W_{T1A} and L_{T1A} - transconductance, channel width and length of the input transistor, C_{ox} - oxide capacitance per unit area, K_f - flicker noise coefficient, T - temperature, e, k - Neper and Boltzmann constant, respectively.

The total noise of the front-end readout system can be always represented by combination of two noise sources. The noise can either be a current or a voltage and they are often referred to as parallel and serial noise. Because these two sources are not correlated, the total input referred ENC calculation can be performed separately. The total noise value is the square root of the sum of the noise contributions squared.

It is useful to express the noise sources as an equivalent noise charge (ENC) at the preamplifier input. It is clear, that the ENC depends on the characteristics of the charge sensitive preamplifier. The ENC corresponds to the charge that must be delivered to the front-end in order to achieve a signal to noise ratio equal to the unity.

The ENC depends on the detector (plus parasitic) capacitance C_{det}, on the detector leakage current $I_{leakage}$, on the input FET capacitance C_{in} and series noise (including both thermal and flicker noise).

For this reason, attention must be paid to the minimization of CSP capacitance and the detector leakage current. Problems with the leakage current become crucial especially after irradiation, which yields as high as a few nanoamperes of current flowing into the input of the readout system. So, it is necessary to minimize this influence.

Table 1. Mathematical formulae of equivalent noise charge (ENC)

Source of noise	Type	The mathematical formulae of ENC
Leakage current	Parallel	$ENC^2_{\text{leakage}} = \dfrac{e^2 \cdot \tau_{\text{p}} \cdot I_{\text{leakage}}}{4 \cdot q};$
Thermal or Johnson noise	Parallel	$ENC^2_{\text{therm}} = \dfrac{e^2 \cdot \tau_{\text{P}} \cdot k \cdot T}{2 \cdot q^2 \cdot R_{\text{P}}};$
White noise	Series	$ENC^2_{\text{white}} = \dfrac{e^2 \cdot k \cdot T \cdot R_{\text{S}}(C_{\text{det}} + C_{\text{fb}} + C_{\text{in}})^2}{q^2 \cdot \tau_{\text{P}}};$
Transistor flicker noise	Series	$ENC^2_{\text{flicker}} = \dfrac{K_{\text{f}}}{2 \cdot C^2_{\text{ox}} \cdot W_{\text{T1A}} \cdot L_{\text{T1A}}} \cdot \dfrac{e^2 \cdot (C_{\text{det}} + C_{\text{fb}} + C_{\text{in}})^2}{q^2};$
Transistor channel thermal noise	Series	$ENC^2_{\text{ch}} = \dfrac{e^2 \cdot k \cdot T \cdot (C_{\text{det}} + C_{\text{fb}} + C_{\text{in}})^2}{3 \cdot q^2 \cdot g_{\text{mT1A}} \cdot \tau_P};$

One of the readout electronics requirements is that it be compatible with detector arrays from different materials. This requires the preamplifier to automatically accommodate a wide range of detector leakage current with little or no change in the output response. The idea for current compensation method lying at the background of this article had been scooped from [10]. The automatic leakage current compensation scheme is shown in Fig. 1.

The complete bias current $2I_p$ is steered through T_{2B} while the current through T_{2A} is turned off. The input node is therefore discharged with a net current of I_p, independent of I_{leak}. The current through T_{3A} is not changed significantly because the large capacitor C_{int} keeps the gate voltage nearly constant. The capacitor C_{int} plays a substantial role in the described leakage current compensation method. Its value has to be sufficiently high in order to prevent the circuit from oscillating. Positive leakage currents lower than $2I_p$ and negative currents lower than I_p can be compensated in each pixel.

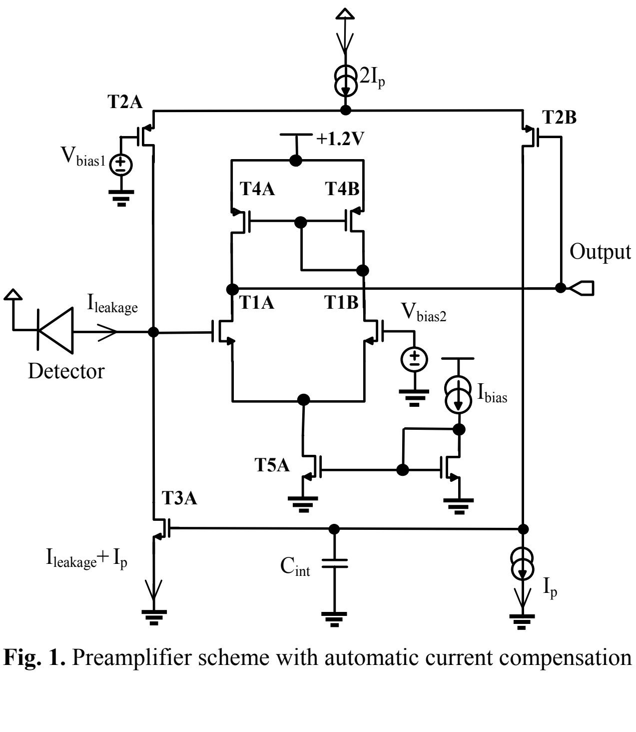

Fig. 1. Preamplifier scheme with automatic current compensation

ENC dependence on the detector leakage current is shown in Fig. 2. The leakage current simulated in the range from -10 nA to +10 nA, but the minimum of ENC is in region -3 nA…+6 nA and equal to 54…70 ē.

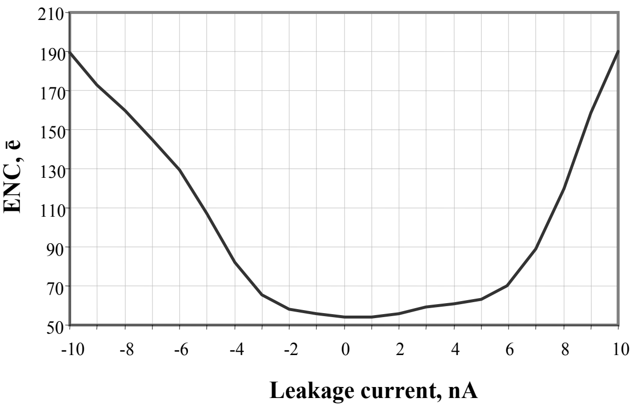

Fig. 2. ENC as a function of the detector leakage current (Q_{in}=1 kē, C_{det}=30 fF, T=27 °C)

The feedback consists of a capacitor C_{fb} in parallel with a resistance R_{fb}, which is implemented by PMOS transistors (T2A and T2B) operating in the linear region. The series of current pulses released by the detector are integrated on the feedback capacitance of the charge amplifier producing voltage steps of approximately Q_{in}/C_{fb} at its output, where Q_{in} is the total charge resulting from integrating the pulse. The value of the feedback capacitor C_{fb} is chosen as the C_{GD} capacitance of input transistor (1):

$$C_{\text{fb}} = C_{\text{GD}} = W_{\text{T1A}} \cdot C_{\text{GD0}} \approx 1fF; \qquad (1)$$

where C_{GD0} is the gate-drain overlap capacitance/channel width.

The value of the feedback resistance R_{fb} is typically in the tens Megaohm range. In our case, this resistance is:

$$R_{fb} = \frac{1}{g_{\text{mT2A}}} + \frac{1}{g_{\text{mT2B}}} \approx 100M\Omega; \qquad (2)$$

where g_{mT2A} and g_{mT2B} are common-source transconductances of transistors T2A and T2B.

The feedback capacitor C_{fb} is discharged by the feedback resistance and the decay time is externally controlled through the bias current 2Ip (Fig. 3). The preamplifier output is directly used for hit discrimination and the discharge must be completed before the next signal arrives. A fast discharge is required in this case. This can lead to a reduction in the peak amplitude if the discharge starts before the signal has reached its maximum due to limited rise time of the amplifier (Fig. 3).

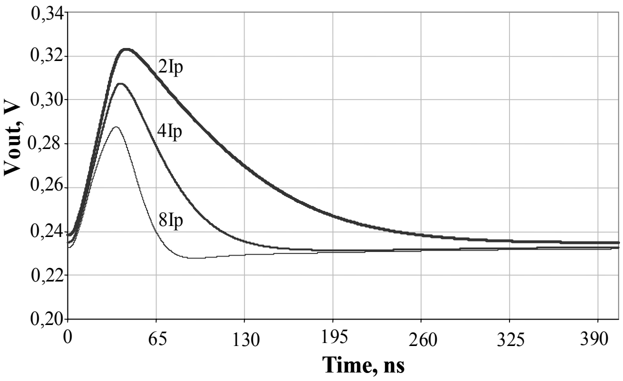

Fig. 3. Preamplifier output signals variation of the bias current 2Ip (Q_{in}=1 kē, C_{det}=30 fF, T=27 ˚C)

Fig. 4 depicts the response at the output of the preamplifier for five different values of input charges, namely 1 kē, 5 kē, 10 kē, 15 kē and 20 kē. The gain of the chain is approximately equal 41.42 mV/kē, while the peaking time is 35 ns.

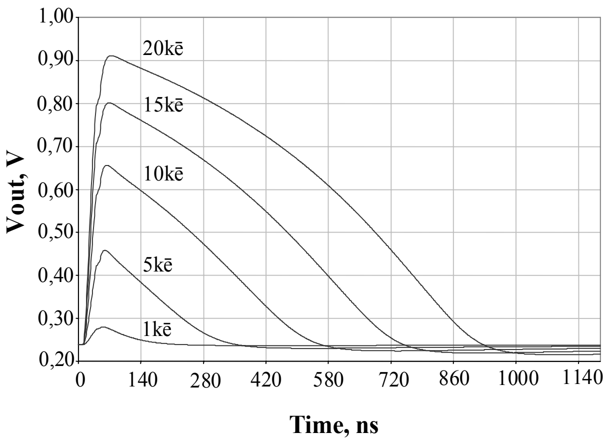

Fig. 4. Transient response at the output of the voltage amplifier (input charge 1 kē, 5 kē, 10 kē, 15 kē and 20 kē, C_{det}=30 fF, T=27˚C)

Fig. 5 shows the peaking time versus detector capacitance. The simulation shows the peaking time of 30 ns at of 0 fF input capacitance and a weak dependence on the input capacitance. From this figure, a charge-to-voltage conversion gain is an exponential function. The gain at the input capacitance C_{det}=0 fF increases up to 68 mV/kē. The expected peaking time (C_{det}=30 fF) for the preamplifier is about 35 ns. The deterioration of the preamplifier speed is due to the increase of capacitance in the coupling with the discriminator stage.

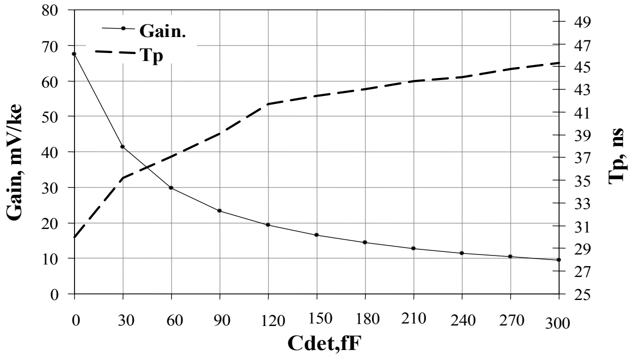

Fig. 5. Preamplifier peaking time and gain versus input capacitance (Q_{in}=1 kē, C_{det}=30 fF, T=27 ˚C)

Total Input Equivalent Noise Charge (ENC) Simulation

The simulated output noise spectral density S_{noise}(f) of the system is shown in Fig. 6. The noise response falls down from low frequencies until around 1 MHz a peak appears. This peak is caused by the detector capacitor.

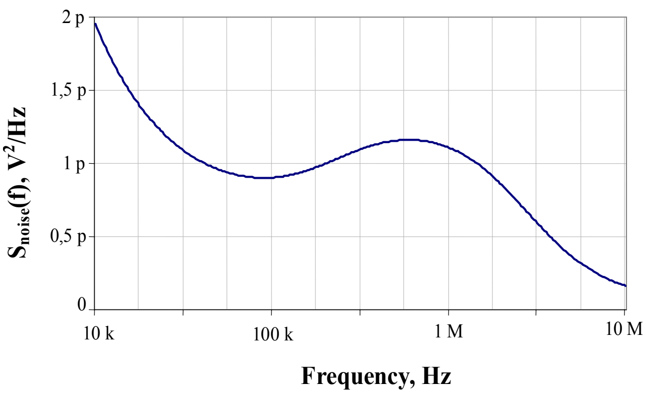

Fig. 6. Simulated output-noise S_{noise}(f) spectral density (Q_{in}=1 kē, C_{det}=30 fF, T=27 ˚C)

The total output noise voltage V_{noise} can be calculated by the following formula:

$$V_{noise} = \sqrt{\int_{\Delta f} S_{noise}(f)df}; \tag{3}$$

where Δf is the system noise bandwidth. For a given charge Q_{in} resulting in an output voltage V_{out}, the mean squared value of input-referred equivalent noise charge (ENC) can be calculated using the following formula:

$$ENC = \frac{V_{noise}}{A}; \tag{4}$$

where A = V_{out}/Q_{in} is the charge-to-voltage conversion gain.

By integrating the noise spectral density curve from 10 kHz to 10 MHz, we obtain a total output noise voltage of 2.22 mV (Fig. 7). The diagrams Fig.6 and Fig.7, however, it indicates us little about the signal noise ratio. Therefore the noise must be calculated to the input as electrons.

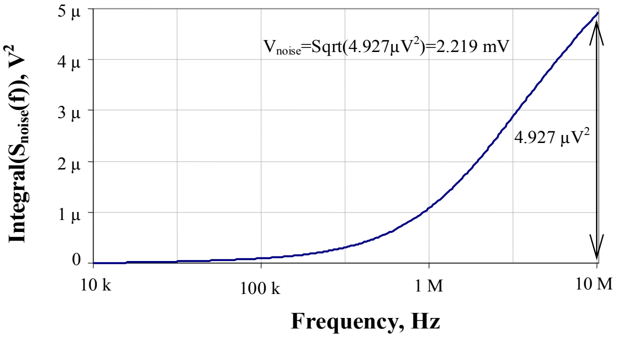

Fig. 7. Simulated integral of output-noise spectral density (Q_{in}=1 kē, C_{det}=30 fF, T=27 ˚C)

Conclusions

In this paper we present a circuit technique for biasing a low noise CMOS preamplifier that uses a MOSFET transistor as a feedback resistor and leakage current compensation for use with radiation detectors. The charge sensitive preamplifier has been designed in submicron 0.13µm CMOS technology. Noise evaluation of the charge preamplifier as input stages has been performed and reported. The minimum of ENC is in the leakage current region -3 nA…+6 nA and equal 54…70 ē.

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Submitted for publication 2007 03 01

V. Barzdėnas, R. Navickas. Leakage Current Compensation for 0.13 µm CMOS Charge Sensitive Preamplifier // Electronics and Electrical Engineering. – Kaunas: Technologija, 2007. – No. 5(77). – P. 33–36.

We have been simulated the submicron CMOS charge sensitive preamplifier (CSP) with the leakage current compensation circuit, which allows reducing equivalent noise charge up to 54 electrons. Simulation has been performed with SPICE simulators using the BSIMV3.3 transistors parameters of the MOSIS 0.13 µm CMOS. The value of the feedback capacitor C_{fb} is chosen as the C_{GD}≈1 fF capacitance of input transistors. The feedback resistance R_{fb} is implemented by PMOS transistors operating in the linear region. CSP has the following main electrical parameters: the minimum of ENC is in the leakage current region -3 nA…+6 nA and equal 54…70 ē, total gain of the chain is K=41.42 mV/kē, while the peaking time is τ_p≈35 ns. It is a promising solution for X-ray pixel detectors. Ill. 7, bibl. 10 (in English; summaries in English, Russian and Lithuanian).

В. Барзденас, Р. Навицкас. 0.13 мкм КМОП зарядочувствительный предусилитель с компенсацией утечки тока // Электроника и электротехника. – Каунас: Технология, 2007. – № 5(77). – С. 33–36.

Зарядочувствительный предусилитель (ЗЧП) для субмикронной 0.13 мкм КМОП технологии моделирован с компенсацией утечки тока, которая позволила уменьшить эквивалентный шумовой заряд до 54 электронов. Результаты получены при помощи программных пакетов SPICE, используя КМОП 0.13 мкм транзисторные модели BSIMV3.3 компании MOSIS. Для емкости обратной связи использована емкость входного транзистора затвор – сток, которая равна C_{GD}≈1 фФ, а для резистора обратной связи R_{fb}≈100 MΩ – PМОП транзисторы, работающие в линейном режиме. Коэффициент преобразования K=41.42 мВ/кē, а длительность фронта выходного импульса τ_p≈35 ns. Ил. 7, библ. 10 (на английском языке; рефераты на английском, русском и литовском яз.).

V. Barzdėnas, R. Navickas. 0,13 µm KMOP krūviui jautrus priešstiprintuvis su nuotėkio srovės kompensavimo grandine // Elektronika ir elektrotechnika. – Kaunas: Technologija, 2007. – Nr. 5(77) – P. 33–36.

Submikroninei 0,13 µm KMOP technologijai apskaičiuotas ir išanalizuotas krūviui jautrus priešstiprintuvis (KJP) su detektoriaus nuotėkio srovės kompensavimo grandine, įgalinančia sumažinti ekvivalentinį triukšmų krūvį ENC, kurio minimumas yra 54 – 70 ē, kai nuotėkio srovė keičiasi nuo -3 nA iki +6 nA. Skaičiavimai atlikti su SPICE programų paketais, naudojant kompanijos MOSIS 0,13 µm technologijos KMOP tranzistorių BSIMV3.3 modelius su triukšminiais parametrais. KJP grįžtamojo ryšio talpa yra įėjimo tranzistoriaus užtūros ir santakos talpa C_{GD}≈1 fF, o grįžtamojo ryšio varža R_{fb}≈100 MΩ – PMOP tranzistoriai, veikiantys tiesinėje mažų srovių srityje. KJP perdavimo koeficientas lygus K= 41,42 mV/kē, o registravimo trukmė τ_p≈35 ns. Il. 7, bibl. 10 (anglų kalba; santraukos anglų, rusų ir lietuvių k.).