GALLIUM-ARSENIDE CHARGE-SENSITIVE PREAMPLIFIER FOR OPERATION IN A WIDE LOW-TEMPERATURE RANGE

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A charge-sensitive preamplifier for temperature operation between 1 and 120 K has been developed and evaluated. It uses double-gate GaAs MESFETs selected for their low 1/f noise. These devices are operated with both gates interconnected emulating single-gate MESFETs of double gate-length, obtaining in this way a value for $A_f$, the coefficient of the 1/f noise spectral power density, of $1.7 \times 10^{-13} \text{V}^2$ at 77 K and $3.8 \times 10^{-14} \text{V}^2$ at 4 K. The latter is one fourth the value exhibited by the original device before modification and two orders of magnitude less than the value measured at 300 K. At the optimum bias operating point device transconductance is 6 mA/V, input capacitance is less than 5 pF and the power dissipation 360 μW. The circuit configuration consists of a double-cascode loaded with a bootstrapped current source. In this way, a high gain-bandwidth product is obtained despite of the low dynamic output resistance, 3000 Ω, exhibited by the MESFETs at the operating point. Equivalent noise charge was measured for detector capacitances up to 35 pF. Using a semi-Gaussian weighting function minimum values of 58 and 20 rmse- at 77 and 4 K, respectively, were determined. Noise slopes are 4.68 and 3.64 e-/pF at 1 μs shaping time for 77 and 4 K, respectively. A rise time of 20 ns was measured at the receiving-end of a 1 m length 50 Ω coaxial cable terminated at sending-end, when detector capacitance was $C_D = 35 \text{pF}$. The total power dissipation of the preamplifier is less than 10 mW.

1. Introduction

Low-noise preamplifiers capable of operation at very low temperatures are of extreme importance for particle detectors which must operate in a cryogenic environment. A broad range of detectors require, for fundamental reasons, a low-temperature environment to work. Cryogenic particle detectors like bolometers, Josephson junctions and superconducting granules operate inside refrigerators, at temperatures ranging from tenths of mK to about 1 K. A low-noise preamplifier installed in a 1 K heat-sink of a dilution refrigerator, allows a short connection of a high-impedance bolometric detector to its front-end electronics. In this way signal integration in the parasitic capacitance is strongly reduced as well as any sort of RFI pick-up which may arise when using a front-end electronics located outside, at room temperature. Also, by establishing the temperature of the hot point at only 1 K, the thermal power injected into the detector through the thermal conductance of the low capacitance interconnecting link is kept to a minimum [1–7].

On the other hand, ionization chambers calorimeters, filled with a cryogenic liquid like LAr, LKr, or LXe working at 87, 120 and 165 K, respectively, require also front-end electronics installed close to the collecting electrode, therefore at low temperature, to avoid the charge transfer delay from electrodes to preamplifier input and to reduce the contribution to the equivalent noise charge (ENC) of the interconnection's capacitance [8].

Low-noise preamplifiers for applications at temperatures above 77 K have been built using Si JFETs, which reach the best noise performance from 120 to 150 K [9]. This temperature is obtained by incorporating a thermal resistance between the device and the thermal bath to rise the actual operating temperature. This solution has as a main limitation the fact that power dissipation cannot arbitrarily be reduced if the thermal resistance is not kept at a sufficiently high value to maintain the optimum operating temperature. As a compensation, Si JFETs exhibit very low 1/f noise.

Below the mentioned optimum temperature, Si JFETs deteriorate their performance as the freezeout of carriers starts to take place. Si MOSFETs can also be employed even at temperatures well below 77 K as carriers are created in the inversion layer by the strong electric field in the Si oxide [9]. Anyhow Si MOSFETs show a highly dominant 1/f noise due to surface states in the Si–SiO₂ which does not reduce, and in many cases even increases, with decreasing temperature [10,11]. Also Ge JFETs have been operated at a temperature of 4 K as the low value of the ionization energy of dopant impurities in Ge, 20 meV compared to 45–50...
meV in Si, prevents the freezing out of carriers even when the temperature reduces to 4 K [12].

Attracted by the absence of carrier freezeout, the high transconductance at low input capacitance and the strong reduction of the 1/f noise with the reduction of temperature observed in GaAs MESFETs, we have proposed [1,7] the use of these devices in the realization of low-noise front-end electronics for cryogenic particle detectors. Following a research on the static and noise performances of selected devices [1], we have designed the first voltage-sensitive preamplifier built totally in GaAs [4]. Further versions have shown improved figures both in noise performance and in power dissipation [6].

We present in this work a charge-sensitive preamplifier realized, for the first time, using exclusively GaAs MESFETs. It operates in a wide range of low temperatures. It is intended to be used either with high-impedance bolometric detectors or with ionization detectors working at cryogenic temperatures. We describe in the following section the characteristics of the devices selected; further on we discuss the particular circuit configuration adopted and finally present the results obtained.

2. Characteristics of GaAs MESFETs at low temperatures

2.1. Static

The low ionization energy of dopant impurities used in GaAs and the high mobilities of carriers make it possible to use these devices at temperatures down to 1 K. In effect, the binding energy of Si and S impurities is 6 meV. As the carrier concentration is proportional to \( \exp\left(-\frac{E_d}{2kT}\right) \), where \( E_d \) is the dopant ionization energy, \( k \) the Boltzmann constant and \( T \) the absolute temperature, it can be observed that the value of \( \sim 10^{-1} \) calculated for \( E_d = 6 \) meV at an estimated die temperature of \( \sim 15 \) K gives a tolerable reduction in the carrier concentration in GaAs compared with a factor \( \sim 10^{-8} \) calculated for the Si doped with P (\( E_d = 45 \) meV) at the same temperature. High electron mobility contributes in obtaining a large transconductance even at low temperature in GaAs MESFETs. In the devices selected for the realization of the present charge-sensitive preamplifier, at \( T = 4 \) K a value of 6 mA/V at an \( I_{DS} \) current of 0.6 mA was measured.

GaAs MESFETs are structures in which the Schottky diode constituting the gate is built on one side only. Near pinch-off, the drain-to-source dynamic resistance has a limited value. This fact imposes a limit on the value of the voltage amplification factor \( \mu = \frac{R_{ds}}{R_m} \) where \( R_{ds} \) is the dynamic output resistance and \( R_m \) the transconductance of the device. In order to obtain high open-loop gain this limitation has to be circumvented by a proper circuit configuration.

Fig. 1. Cross section of a double-gate GaAs MESFET (a). When both gates are connected together (b), the device becomes equivalent to a single-gate MESFET with double gate-length.

Regarding gate leakage current, the high value observed in GaAs MESFETs at room temperature reduces by more than five orders of magnitude when the operating temperature reduces from 300 to 100 K. In the selected devices the value of 10 nA measured for \( I_{G} \) at 300 K decreased to less than 10 fA at 120 K. At 4 K the leakage current is totally negligible.

2.2. Noise performances

Noise sources in GaAs MESFETs are the following ones: thermal noise in the channel, shot noise due to gate leakage current, thermal noise induced into the gate, 1/f due to the drain current, and generation-recombination (g-r) in the depletion region. At low frequencies 1/f and g-r noise sources are dominant. At low temperatures, shot noise due to gate leakage current becomes negligible.

Representing 1/f noise as a noise current generator between drain and source terminals, the monolateral spectral power density \( S_f \) can be expressed according to the experimental Hooge's law as

\[
S_f = \frac{\alpha_{HH} I_{DS}^2}{fN},
\]

where \( \alpha_{HH} \) is the Hooge's parameter, \( I_{DS} \) the drain current, \( f \) the frequency and \( N \) the number of carriers in the channel [13]. If the device is operated below saturation, and referring to the input terminals, the spectral power density of the voltage noise generator \( S_{1/f} \) in series with the gate can be written as

\[
S_{1/f} = \frac{\mu \alpha_{HH} I_{DS} V_{DS}^2}{g_m I_{DS}^2} f^{-1}.
\]
where \( q \) is the electron charge, \( \mu_n \) the electron mobility, \( V_{DS} \) is the drain–source bias voltage, \( S_m \) the device transconductance, \( L \) the gate length.

Generation–recombination noise in junction field-effect transistors is originated by the random trapping and emission of carriers at deep levels in the depletion region [14]. In GaAs the density of impurities and defects is higher than in Si, consequently total g–r noise is higher. In addition, a large number of trap levels gives origin to a distribution of the g–r noise spectral power density which results from the addition of the Lorentzians due to the single trap [15]. The total spectrum has a \( 1/f \) distribution. The g–r noise generator can be represented as a noise voltage source in series with the gate terminal having a spectral power density given by the following expression:

\[
S_{v_{gr}} = 4kT \sum_{i} \frac{\rho_{gr}(\tau_i/\tau_0)}{1 + (2\pi f)^{2}/\tau_i^2},
\]

where \( \rho_{gr} \) is an effective resistance depending inversely on the gate area, and directly on the deep-level trap concentration in the depletion region, \( \tau_i \) the characteristic time constant of the \( i \)th trap and \( \tau_0 \) a normalizing value [16]. In GaAs MESFETs another low-frequency noise source is originated in the interface between the conducting channel and the semi-insulating substrate.

Total low-frequency noise is determined by the mentioned sources and shows an approximately \(-1\) slope in the power spectrum distribution. For that reason we refer to it hereafter as “\( 1/f \)” noise although it is understood that it is originated by different mechanisms. The spectral power density \( S_f \) of the total \( 1/f \) noise source in series with the gate terminal can be expressed as

\[
S_f = A_f f^{-1}.
\]

We have observed that the coefficient \( A_f \) is highly
dependent on temperature and biasing conditions. Particularly, $A_f$ decreases by more than two orders of magnitude when the temperature decreases from 300 to 4 K. From 300 to 77 K this reduction is more than two orders of magnitude. This behaviour is consistent with the known increase of the trap time constant $\tau$ (eq. (3)) with decreasing temperature and consequent shift towards lower frequencies of the $g-r$ noise spectrum “tail” [17]. On the other hand we have verified, as it will be shown below, that at very low temperature and above 100 Hz low-frequency noise appears as Hooge’s type. Its reduction with decreasing temperature is attributed to the dominant temperature dependence of the electron mobility term in eq. (2).

Another important characteristic observed during evaluation of our devices was the reduction of $A_f$ with decreasing biasing voltage $V_{DS}$ consistent with that expressed by Lauritzen and Sah while stating the theory on g–r noise [14]. This reduction is also consistent with eq. (2). For that reason devices employed in the realization of the preamplifier have been biased at low $V_{DS}$ near the linear region.

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Fig. 3. Superimposed output characteristics of a 3SK164 before and after short-circuiting the gates. Note the increase of the output dynamic resistance in the second case. (a) $T = 4$ K; (b) $T = 300$ K.

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Fig. 4. Noise voltage density of a 3SK164-M at 4 and 77 K. The values at 1 kHz, 6.2 nV/Hz$^{1/2}$ at 4 K and 13 nV/Hz$^{1/2}$ at 77 K are given as a reference. As the input capacitance is 5 pF, $H_f$ becomes $2 \times 10^{-25}$ J and $9 \times 10^{-25}$ at 4 and 77 K, respectively.

III. FRONT-END ELECTRONICS
The $g-r$ noise expressed by eq. (3) has an inverse dependence with the gate area [14]. Eq. (2) predicts an inverse dependence of the $1/f$ noise with the square of the gate length. This makes clear the need of using large-area devices, with long gate-length, to obtain reduced low-frequency noise. Unfortunately, commercial GaAs MESFETs are manufactured for high frequencies, with short gate-length, and with low input capacitance, therefore the gate area is kept small. The factor of merit used to evaluate the quality of a field effect transistor regarding the $1/f$ noise is the parameter $H_f$ introduced by V. Radeka. $H_f = A_f C_i$, where $C_i$ is the device's input capacitance. We have measured the lowest $H_f$ for a GaAs MESFET at 4 K [18]. The value of $5.7 \times 10^{-26}$ J measured in a NE41137 is only one order of magnitude larger than the best quality Si JFET at 300 K [19]. This confirms that GaAs devices give now excellent noise performance at low temperatures.

The $1/f$ noise current is $I_{G} = 10 \, \text{fA}$ at 100 K.

Fig. 5. Gate leakage current of a 3SK164-M as a function of temperature. At 120 K, $I_G = 30 \, \text{fA}$. The minimum measured value was $I_G = 10 \, \text{fA}$ at 100 K.

Fig. 6. The simplest version of a dominant-pole voltage amplifier. The limited value of the impedance contributed by $Q_1$ and the current source at the node A limits both the dc voltage gain as well as the bandwidth.

Fig. 7. The double cascode loaded with a bootstrapped current source circuit configuration adopted to increase the impedance at node A. Note that $Q_1$, $Q_2$, $Q_4$ and $Q_6$ are all biased at the same $V_{DS} = V_{GS}$ and same $I_D$. 

$V_{CC} = 6 V_{GS}$
should be remarked, however, that noise performances are very much batch-dependent. We have tested several devices of the same batch and got similar results, but different from other devices of the same type but from different batches. This result is not surprising as the 1/f noise is, as mentioned before, process-dependent.

We have observed in our noise measurements, to be described in the following section, that at room temperature the low-frequency noise is predominantly of generation–recombination type. In fact, we have tested the inverse gate-area dependence in accordance with that expected for the g–r noise. At low temperatures, instead, the noise appears as Hooge’s type. This fact is proved by the $1/L^2$ dependence of the noise spectral power density found when the device was at 77 and 4 K, in accordance with eq. (2). The total low-frequency noise power spectrum shows a 1/f dependence.

2.3. Selected device

We have selected a double-gate device, the Sony 3SK164, designed for applications at relatively low frequency of operation, 900 MHz, therefore having longer gate-lengths than other similar MESFETs for higher frequencies. A modification in the normal mode of operation was introduced by connecting both gates together obtaining in this way a single-gate MESFET with double gate-length referred to hereafter as 3SK164-M (fig. 1). The result of such a simple modification in the device geometry led to a reduction in the 1/f noise by a factor four at 4 and 77 K. The noise reduction at 300 K was by a factor 1.6 (fig. 2). The cause of this behaviour was explained in the previous section.

The modified device presents improved static characteristics as can be observed in fig. 3 where the increase of the output dynamic resistance is apparent. In the same figure the optimum biasing operating point is indicated. At this point, where $I_D = 0.6$ mA and $V_{DS} = 0.6$ V, the value of transconductance $g_m$ is 6 mA/V, with a power dissipation of only 360 $\mu$W. Output dynamic resistance is 3000 $\Omega$. The coefficient of the 1/f noise spectral power density $A_f$ is $3.8 \times 10^{-14}$ V$^2$ at 4 K and $1.7 \times 10^{-13}$ V$^2$ at 77 K. The spectral power density as a function of frequency for 3SK164-M at different temperatures is shown in fig. 4.

Total input capacitance of 3SK164-M is less than 5 pF and was determined from the measurements of ENC as a function of total detector capacitance as will be explained in section 4. In consequence, the 1/f noise characteristic parameter $H_f$ is lower than $1.9 \times 10^{-25}$ J and $8.5 \times 10^{-25}$ J at 4 and 77 K, respectively.

Gate leakage current decreases from 10 nA at 300 K to less than 10 fA at 120 K (fig. 5). Below that temperature, it was not possible to accurately measure such a small current, but its exponential dependence on $1/T$ makes one presume a totally negligible leakage current at 4 K. Indirect measurements made with voltage-sensitive preamplifiers using the same devices confirmed this assumption.

3. Design of high open-loop gain amplifier using low $R_{ds}g_m$ product field-effect transistors

It was mentioned before that in order to keep low-noise performance the selected GaAs MESFETs have to be operated at low $V_{DS}$ and at low current $I_D$. In these conditions the dynamic output resistance $R_{ds}$ is low and the voltage amplification factor $\mu = R_{ds}g_m$ is also reduced. A simplified dominant pole voltage amplifier (fig. 6) will give voltage gain:

$$\frac{v_o(s)}{v_i(s)} = g_m R_f R_{ds} \frac{1}{(1 + sC_f R_f R_{ds})} \tag{5}$$

where $R_f$ is the dynamic output resistance of a current source made with the same low-$\mu$ device.

It is clear that, in order to obtain high voltage gain and high bandwidth, $Q_1$ has to be replaced by a cascode configuration to increase the reflected dynamic output impedance at the node A, and the current source must be realized using a circuit configuration which assures high dynamic impedance. A double-cascode ($Q_1, Q_2$,
Q₃) was used for the first purpose and a bootstrapped current source for the second one (Q₄, Q₅) (fig. 7).

The dynamic resistance $R_{ds}'$ contributed by the (Q₁, Q₂, Q₃) configuration at node A is

$$R_{ds}' = R_{ds3} + [R_{ds2} + R_{ds1}(1 + R_{ds2}g_{m2})]$$

$$\times (1 + R_{ds3}g_{m3}).$$

(6)

If all $R_{dsi} = R_{ds}$ and $g_{m1} = g_{m2} = g_{m}$, then $R_{ds}' = R_{ds}g_{m}^2$.

At the above-mentioned operating point, selected devices have $g_{m} = 6 \text{ mA/V}$ and $R_{ds} = 3000 \Omega$, in consequence $R_{ds}' = 1 \text{ MΩ}$.

A calculation of the dynamic resistance of the bootstrapped current source gives

$$R_{j} + R + R_{ds4}(1 + g_{m4}R)(1 + g_{m5}R_{ds5}) + R_{ds2},$$

(7)

which once more assuming identical devices give $R_{j} = R_{ds}^2$. For the selected device, $R_{j} = 500 \text{ kΩ}$.

Transistors Q₆ and Q₇ form an identical current source to the one constituted by Q₁–Q₅. The same current $I$ passes through them assuring that the drain–source bias voltage of Q₂ and Q₆ is equal to that of Q₁ and Q₄. Current $I$ is much smaller than the saturation current $I_{DSS}$, therefore $V_{GS}$ is enough to bias the transistors in the transition between linear and saturation region. In this way, low-noise operation of those transistors contributing to the input noise voltage source, is assured by biasing them at a low drain–source bias voltage.

The circuit includes a voltage-level shifter D₁–D₂.

**Fig. 9.** ENC of the GaAs charge-sensitive preamplifier as a function of detector capacitance (a) and as a function of shaping time (b). $T = 77 \text{ K}$. ENC vs $C_{D}$ is given by $\text{ENC} = 52 + 4.68 \times C_{D}(\text{pF})$ at $\tau = 1 \mu\text{s}$ ($C_{D}$ in pF). Minimum ENC is 58 r.m.s. e⁻.
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The circuit configuration described above can be employed with any other type of field-effect transistors in which low-noise operation can be assured by biasing them at low drain-source voltage. The decrease of the voltage amplification factor $\mu$ is compensated by the use of a double-cascode and bootstrapped current source. In

Fig. 10. ENC of the GaAs charge-sensitive preamplifier as a function of detector capacitance (a) and as a function of shaping time (b). $T = 4$ K. ENC vs $C_D$ is given by $\text{ENC} = 28 + 3.64 \times C_D$ at $\tau = 1$ $\mu$s ($C_D$ in pF). Minimum ENC is 20 rms.

The dc open-loop gain of the circuit shown in fig. 8 was measured and determined to be 1400. A $-3$ dB cutoff frequency was found at 180 kHz giving a gain-bandwidth product of 252 MHz. We have designed and tested voltage-sensitive preamplifiers based on GaAs MESFETs which, at the expense of a much higher power dissipation, present a gain-bandwidth product of 350 MHz measured at a closed-loop gain of 50 [6].

Although $V_{GS}$ varies somewhat with temperature the circuit configuration assures even distribution of potentials. By keeping all transistors working at low drain-source voltage, the power dissipation is kept to a minimum. For the present case $P_d < 10$ mW.

Actually gate-source Schottky junctions of similar GaAs MESFETs, which fixes the drain-source bias voltage of $Q_3$ also to $V_{GS}$. By fixing $V_{CC} = 6 |V_{GS}|$ also $Q_3$ is biased at the same $V_{DS} = |V_{GS}|$ condition.

III. FRONT-END ELECTRONICS
4. The charge-sensitive preamplifier

The circuit (fig. 8) consists of the dominant pole amplifier described in the preceding section feedbacked with a 1 pF high-Q Vitramon capacitor [21] in parallel with a 1 GΩ Eltec 102 resistor [22].

The feedback network determines a decay time constant of 1 ms at 77 K and 2 ms at 4 K. The difference in decay time constants is determined by the increase of the feedback resistor value at very low temperatures.

Charge sensitivity is fairly constant with temperature due to the high stability of the feedback capacitor.

The ENC of the charge-sensitive preamplifier was measured for several detector capacitances up to 35 pF. Measurements were done at 4 and 77 K. Results are shown in figs. 9 and 10. A semi-Gaussian weighting function with shaping time from 0.5 to 10 μs was used. Results were quite satisfactory: at 4 K and for zero detector capacitance, a minimum ENC of 20 rms e⁻ was determined. At 77 K, the ENC was 58 rms e⁻ also for \( C_D = 0 \). Fig. 11 show the response of the charge-sensitive preamplifier followed by the semi-Gaussian shaper (Ortec 450 research amplifier), to a test charge input.

The minimum ENC was found to be practically

![Image](image-url)
constant from 3 to 10 $\mu$s shaping time (we did not make measurements above 10 $\mu$s). This result is explained by the predominant contribution of the 1/f noise. In effect, when parallel noise contribution is very small, as in the present case, the ENC for the semi-Gaussian shaping is given by the following expression:

$$\text{ENC} = \left[0.75 \varepsilon_{\text{nw}}^2 \left(\frac{1}{\tau}\right) + 3.2 A_f \right] \frac{1}{C_t}, \quad (8)$$

where $\varepsilon_{\text{nw}}^2$ represents the monolateral white series noise spectral power density of the amplifier, $\tau$ the shaping time and $C_t$ the total input capacitance of the amplifier including detector, feedback, input and test capacitances.

If we compute the noise slope squared $\text{NS}^2 = (\partial \text{ENC}/\partial C_t)^2$ we get

$$\text{NS}^2 = 0.75 \varepsilon_{\text{nw}}^2 \left(\frac{1}{\tau}\right) + 3.2 A_f. \quad (9)$$

Fitting the experimental values of NS for different $\tau$ (figs. 9 and 10) with eq. (9) it is possible to determine the value of $A_f$ and $\varepsilon_{\text{nw}}^2$ (fig. 12). The values of $A_f$, obtained with this method are totally in agreement with those obtained from direct measurements of series noise spectral power density of transistors used. In addition, it can be observed that the thermal noise does not change significantly as would be in principle expected when the temperature decreases from 77 to 4 K. This fact is attributed to the increase of the channel thermal noise equivalent resistance at low temperatures [23,24].

From the measurements of ENC as a function of $C_D$ at short shaping times, the extrapolated matching capacitance was determined. Its value at 4 K is 8 pF and slightly larger at 77 K (figs. 9 and 10). A preamplifier matching higher detector capacitances is at the moment under development.

Regarding the dynamic behaviour, the charge-sensitive preamplifier loaded with a 1 m long 50 $\Omega$ coaxial cable terminated at the sending-end shows at the receiving-end a transient response which at $C_D = 0$ has a rise time of 15 ns. The rise time increases when the detector capacitance $C_D = 35$ pF. At 77 K it is 22 ns and at 4 K

Fig. 13. Transient response of the charge-sensitive preamplifier. Pulse shape was measured at the end of a 1 m long 50 $\Omega$ coaxial cable terminated at the sending-end with a 50 $\Omega$ resistor. (a) $T = 4$ K, (b) $T = 77$ K.

Fig. 14. Physical layout of the GaAs charge-sensitive preamplifier (scale in cm).
Table 1
GaAs charge-sensitive preamplifier specifications. General: operating temperature 1-120 K, power dissipation 7 mW.

<table>
<thead>
<tr>
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<th>$T = 77$ K</th>
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<td>0.16</td>
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<td>1</td>
<td>ms</td>
</tr>
<tr>
<td>Detector matching capacitance</td>
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<td></td>
<td></td>
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<tr>
<td>Minimum ENC @ $C_D = 0$ pF</td>
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<td>9</td>
<td>pF</td>
</tr>
<tr>
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<td>58</td>
<td>rms e$^-$</td>
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<tr>
<td>Rise time @ $C_D = 35$ pF</td>
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<tr>
<td>Dynamic range</td>
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<td>84</td>
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20 ns (fig. 13). The dynamic range is 120 dB at 4 K and 84 dB at 77 K.

The highest temperature of operation is limited by the increase of the leakage current which above 120 K becomes larger than 30 fA (fig. 5) and by the tolerable ENC which also increases with temperature. In addition, above 120 K a Si JFET-based charge-sensitive preamplifier will give better noise performance due to the lower value of its 1/f noise.

Table 1 shows the complete specification of the GaAs charge-sensitive preamplifier. The circuit was mounted on a G-10 printed circuit board; a new version employing a ceramic substrate is at the moment being developed and will be ready soon.

5. Conclusions

For the first time, a charge-sensitive preamplifier which operates from 1 to 120 K has been developed and tested. The use of GaAs MESFETs and a new circuit configuration made it possible to attain a low equivalent noise charge as well as very good dynamic performance with low power dissipation.

We have confirmed our previous findings which showed that some types of GaAs MESFETs, when properly biased, give excellent noise performance at very low temperatures and make possible the realization of low-noise preamplifiers for particle detectors working at cryogenic temperatures.

The reduction by a factor four in the low-frequency noise observed when short-circuiting both gates in double-gated GaAs MESFETs makes plausible that at 4 K low-frequency noise is probably of Hooge's type. The availability of devices with longer gate-length should exhibit lower 1/f noise per unit of gate area at very low temperatures. Also, it is evident that the development of large-area devices will be useful for the realization of charge-sensitive preamplifiers with higher detector matching capacitance.

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