# A RESET-NOISE FREE HIGH SPEED READ-OUT MODE
# FOR A FLOATING-DIFFUSION DETECTOR WITH RESISTIVE FEEDBACK

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*Abstract.*

**A different mode of operating the floating diffusion-detection node is investigated. This new mode has no reset-noise and therefore no need for reset-noise suppression. The demand on the bandwidth of the on-chip amplifier is therefore relaxed from about 3 times the pixel rate (a minimum requirement to suppress reset noise sufficiently) to about half the pixel rate. Also the time needed to dump the charge on the floating diffusion can be relaxed from some nanoseconds to half the pixel clock time. The optimal signal processing is an ordinary low-pass filter.**

**It will be shown that in the new mode one can maintain equal or better noise performance than with time-discrete processors like Correlated Double Sampling (CDS) or any other signal processing.**

*Introduction.*

The flow of charge through the horizontal register can be regarded either as a series of charge packets or as a current. In the first case one uses a capacitor to convert charge into a voltage and in the second case a resistor. In many CCD imagers the choice is made to use a floating diffusion for the capacitor which must be reset after every pixel.

Also in high-speed devices where charge is clocked at 36 MHz or 72 MHz it is very difficult to design a good signal processor. To give an example one clock interval of 72 MHz has a period time of 13.9ns in which 6.9ns for video (50% duty cycle), 4.6ns for the reset-noise hold level and 2.3ns for switching the resetfet on. Generating the necessary pulses which have rise and fall times less then ns and have the proper timing is a problem. When using a current which is converted into voltage no special timing is required and signal processing is time-continues filtering.

The intention of this feasibility study is to investigate whether using the current output is a good alternative for a floating diffusion schema in high-speed applications.

The analysis is focussed on high frequencies. The 1/f-noise contribution is therefore neglected.

*The destructive charge readout with floating diffusion and resetfet.*

The traditional floating diffusion configuration [1] for reading-out charge packets is shown in Fig.1. The capacitance between input and ground (C) is the charge sensing capacitance. The voltage source $e_n$ represents the equivalent noise of the on-chip amplifiern at the input. The current source i represents the charge dumped (qN) at the floating diffusion, for one charge packet the following relation holds

$$\int_{-\infty}^{\infty} i(t)\ dt = q\ N \qquad (1)$$

When the detection node capacitance (C) has received a charge packet the resetfet will clamp (reset) the capacitance to a reference potential $V_{ref}$ and the capacitance (C) is prepared to receive the next charge packet. Due to this reset noise is generated.

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This resetnoise can reach high values like an equivalent of 40 electrons. With signal processors like Correlated Double Sampling, Integrate and dump, Delay Line processing, Clamp sample etc. one can suppress the resetnoise and obtain reasonable noise levels. The signal is always resampled at a hold capacitance resulting in the sin(x)/x spectrum.

The noise voltage at the output of the on-chip amplifier ($G.e_n$) divided by the output voltage due to one-electron ($G.q/C$) gives an easier quantity to calculate noise performance. Its dimension is $e/\sqrt{Hz}$ and called the noise electron density NED [5].

It is shown [2,3] that a the noise limit which can be obtained with optimal[3] filtering is:

$$S_{proc}(f) = M\left(e_n \frac{C}{q}\right)^2 sinc^2\left(\frac{f}{f_h}\right) \tag{2}$$

Without going into details the signal processor's figure of merit (M) for the optimal filter, a dimension less quantity, for a one-output CCD is:

$$M = \frac{T_h}{T_v} + \frac{T_h}{T_{rs}} = 5 \tag{3}$$

in this equation the different times are resp.:
$T_h$ :the inverse of the horizontal clock frequency $T_h=1/f_h$.
$T_v$ :the time video is available (charge is hold at capacitor C) and usually is $T_h/2$,
$T_{rs}$: the reset hold level which is usually about $T_h/3$.
The integrate and dump processing is the optimal processing. The integrate action gives the optimal filtering and the dump the sin(x)/x spectrum.

The number of equivalent noise electrons after signal processing $N_{proc}$ within a bandwidth B

$$N^2_{proc} = \int_0^B M \cdot \left(e_n \frac{C}{q}\right)^2 \cdot sinc^2\left(\frac{f}{f_h}\right) df \tag{4}$$

In practice it is very difficult to build such an optimal processor for high speeds and the figure of M=5 will hardly be met.

For the general case the capacitance C must be replaced by the total capacitance of the detection node [5].

*Destructive charge readout with resistive feedback.*
This new mode needs an on-chip high ohmic resistor in the same position as the well known reset transistor Fig.1 and an off-chip inverter stage (-A) for the feedback.
The use of a resistor to convert charge into voltage is a very old one, just as the well known floating-diffusion detector with on-chip amplifier [1,4].
New is the feed back of the output voltage of the on-chip amplifier by an on-chip resistor. The resistor is placed on-chip because an off-chip resistor increases the capacitance C with some pF's due to bondpads, wiring and stray capacitance and can never obtain a sufficient low noise level.

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[3] *Optimal is defined as: minimize the noise after filtering with constraints that the reset noise is fully suppressed and the resolution stays the same.*
*The solution is the two-slope integrator:one slope integrates the reset hold level and the second slope the video+resetnoise. It is also called integrate and dump.*

The tranfer function H(f) between (charge) current I(f)[4] and the output voltage $V_{out}$

$$\frac{V_{out}(f)}{I(f)} = H(f) = -\frac{A\,G}{1+A\,G}\;\frac{R}{1+\dfrac{j\,2\,\pi\,f\,R\,C}{1+A\,G}} \tag{5}$$

in which R is the on-chip feedback resistor, C the capacitance at the floating diffusion, A the gain of the external feedback amplifier and G is the gain of the on-chip amplifier
With a high open-loop gain (A.G->∞) the transfer function reduces to

$$V_{out}(f) = R\,I(f) \tag{6}$$

The average current per pixel with N electrons in one charge packet

$$<i> = \frac{1}{T_h}\int_{-\infty}^{\infty} i(t)\,dt = \frac{q\,N}{T_h} \tag{7}$$

The average output voltage $V_{out}=R.q.N/T_h$
The noise spectral density ($A^2/Hz$) of the feed-back resistor is 4kT/R and the noise spectral density ($V^2/Hz$) of the on-chip amplifier $e_n^2$. They give rise to a noise spectral density at the output of the feed back amplifier of

$$H(f)\,H(f)^{*}\left[\frac{4\,kT}{R} + \frac{e_n^2}{R^2}\,[1+(2\,\pi\,f\,R\,C)^2]\right] \tag{8}$$

which reduces with the aid of (5), (6) and (8) to

$$4\,kT\,R + e_n^2\,[1+(2\,\pi\,f\,R\,C)^2] \tag{9}$$

The noise power in a bandwidth B

$$4\,kT\,R\,B + e_n^2\left[B + \frac{[2\,\pi\,R\,C\,B]^2}{3}\,B\right]$$

divided by the average output voltage ($q.1.R.f_h$) due to one electron squared gives also an expression for the noise in electrons

$$\frac{4\,kT}{q^2\,f_h^2\,R}\,B + \left[\frac{e_n}{q\,f_h\,R}\right]^2 B + \left[\frac{e_n\,C}{q}\right]^2\left[2\,\pi\,\frac{B}{f_h}\right]^2\frac{B}{3} \tag{11}$$

*Discussion.*
The integrate and dump has a noise spectrum which is sin(x)/x (2) shaped whereas the other mode has a triangular noise spectrum (9), Fig 2.
In the case of an image sensor one has the benefit of the fact that at equal noise power the subjective evaluation of the noise in the new mode is even better due to the triangular shape of the noise spectrum

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[4] *The Fourier transform of the current i(t) is I(f). All capitals with argument f form a pair with lower case with argument t by means of the Fourier transform.*

and the fact that the human eye is less sensitive for noise at high spatial frequencies.

The ratio between the noise electrons in both cases: the noise electrons in the resistive feedback case (11) divided with the noise electrons in the traditional case (4) is

$$\frac{4\,kT\,R}{f_h^2\,R^2\,C^2\,\langle M\rangle\,e_n^2}+\frac{1}{f_h^2\,R^2\,C^2\,\langle M\rangle}+\frac{4\,\pi^2}{3\,\langle M\rangle}\left[\frac{B}{f_h}\right]^2 \tag{12}$$

the figure of merit for the integrate and dump (M) must be corrected for the sin(x)/x

$$\langle M\rangle = \frac{M}{B}\int_0^B sinc^2\left(\frac{f}{f_h}\right)\,df \tag{13}$$

Clearly the most rigtly part is only determined by the ratio between bandwidth and clocking frequency ($B/f_h$). This ratio will allways be smaller then 1/2 since no signal reconstruction is possible and therefore not usefull for outputsignals above half the sample rate.

Given a CCD imager then $C, f_h, e_n$ and B are fixed and the only degree of freedom left is the choice of the feedback resistor R.

Asume an imager with C=16 fF, $f_h$=72MHz, $e_n$=25nV/√Hz and B=30 MHz and <M>=4.17 then the value of the feed-back resistor must be at least R=2.5 Mohm. For a larger value the resistive feedback mode is even better in noise performance.
For small signals one can even bias the reset transistor in such a way that it emulates the resistor value one needs.

A practical system has not an infinite gain and the pole-zero combination must be considered. With a right choice of these time constants and feedback gain A it is possible to shape the output waveform of one pixel.
A simulation with an actual on-chip amplifier which has a bandwidth of 170MHz in one case and in the other case 36MHz is shown in Fig. 3 and Fig. 4. Clearly to see is the in-sensitivity for duration of the 'charge' pulse and bandwidth, as was expected.

## *Conclusion.*

In the new mode the old tradition is restored together with the benefits of the present: the low noise on-chip amplifier with floating diffusion and an on-chip resistor.

It is a concept for reading-out charges at high frequencies and maintaining the same or better signal to noise ratios as before without the need for complex signal processing schemes. This can come in handy with a progressive scan HDTV sensor which needs pixel rates of more then 100 MHz.

The benefits of using the resistive-feedback technique lays in the fact that signal processing is continue and one needs no time-discrete filtering like Integrate and dump, correlated double sampling etc.
The rise- and fall times play no important role anymore. One may even use up to one full clock time to dump the charge on the floating diffusion.

# LITERATURE

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## FIGURES

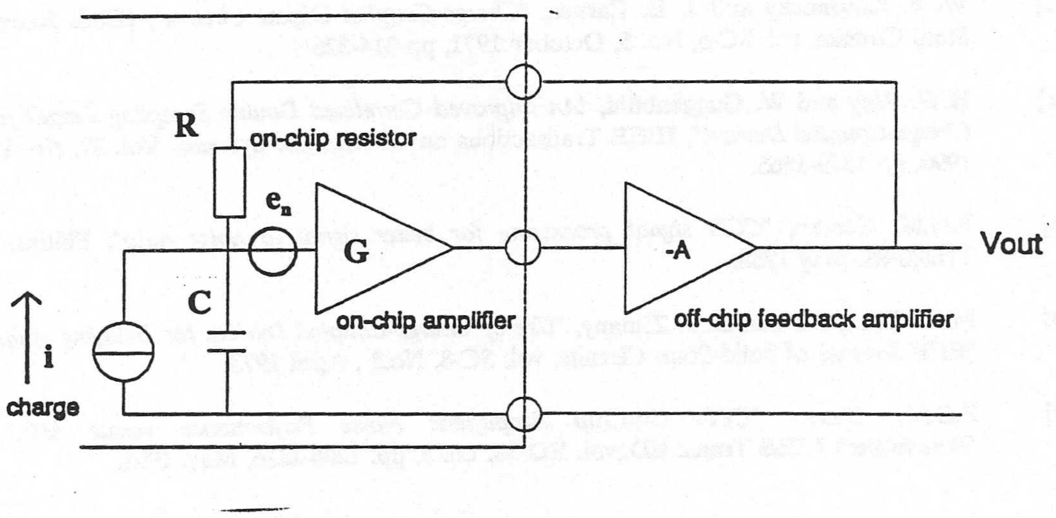

*Fig.1* *The resistive feedback read-out mode.*
*On-chip are the amplifier, the resistor and the floating diffusion. Off-chip is the negative feedback amplifier together with part of the feedback loop. The resetfet is replaced with a resistor as compared with the traditonal concept of a floating diffusion detector.*

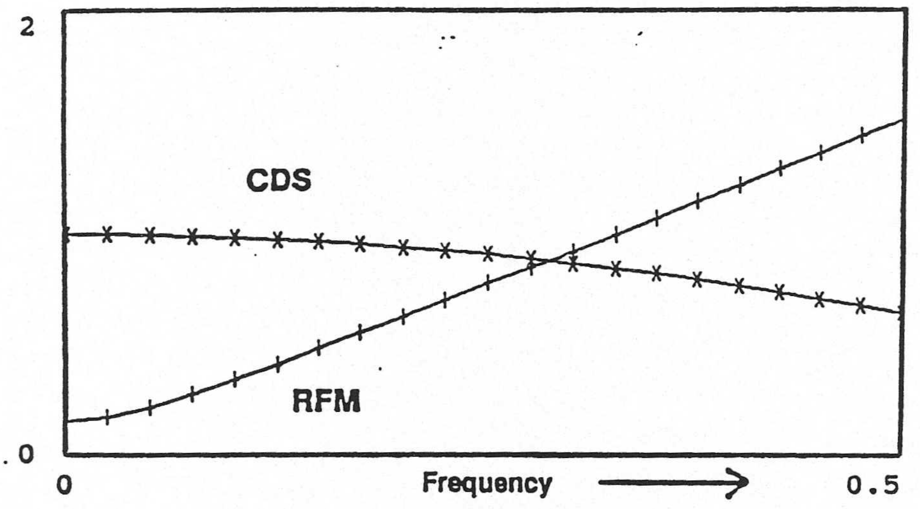

*Fig.2* *Noise spectrum after CDS* *: sin(x)/x*
*and after the resistive feedback mode (RFM)* *: triangular shaped spectrum.*
*The horizontal axis [f] is the frequency normalized by the pixel rate. Vertically the noise spectrum is shown in arbitrary units.*
*The noise power in a bandwidth upto half the pixel rate is the same in both cases. Notice the much lower noise level at lower frequency for the Resistive Feedback Read-out mode compared with CDS.*

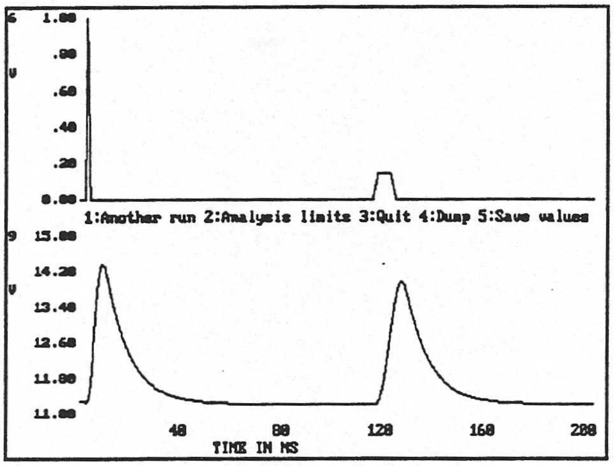

Fig. 3 *The responce (lower graph: node 9) in the time-domain of the resistive feedback floating-diffusion detector as a function of two different charge pulses (upper graph: node 6). One has a duration of 1 ns and the other 7 ns. The spacing between the two pulses is 8 pixels. The bandwidth of the on-chip amplifier is 170 MHz.*

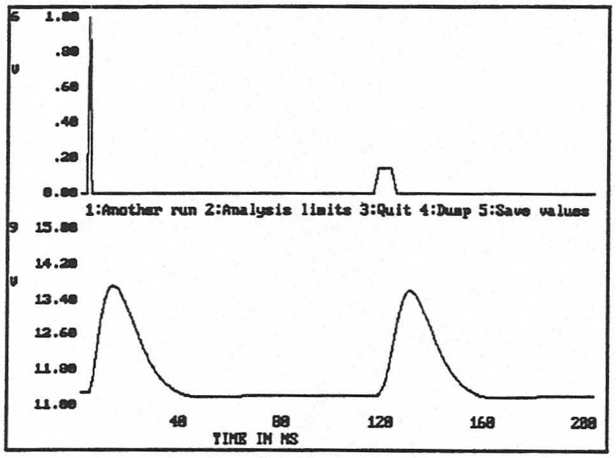

Fig. 4 *The responce (lower graph: node 9) in the time-domain of the resistive feedback floating-diffusion detector as a function of two different charge pulses (upper graph: node 6). One has a duration of 1 ns and the other 7 ns. The spacing between the two pulses is 8 pixels. The bandwidth of the on-chip amplifier is 36 MHz.*