Characterization of Low Light Performance of a CMOS Sensor for Ultraviolet Astronomical Applications

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Abstract. CMOS detectors offer many advantages over CCDs for optical and UV astronomical applications, especially in space where high radiation tolerance is required. However, astronomical instruments are most often designed for low light-level observations demanding low dark current and read noise, good linearity and high dynamic range, characteristics that have not been widely demonstrated for CMOS imagers. We report the performance, over temperatures from 140 - 240 K, of a radiation hardened SRI 4K\texttimes{}2K back-side illuminated CMOS image sensor with surface treatments that make it highly sensitive in blue and UV bands. After suppressing emission from glow sites resulting from defects in the engineering grade device examined in this work, a 0.077 me\textsuperscript{−}/s dark current floor is reached at 160 K, rising to 1 me\textsuperscript{−}/s at 184 K, rivaling that of the best CCDs. We examine the trade-off between readout speed and read noise, finding that 1.43 e\textsuperscript{−} median read noise is achieved using line-wise digital correlated double sampling at 700 kpix/s/ch corresponding to a 1.5 s readout time. The 15 ke\textsuperscript{−} well capacity in high gain mode extends to 120 ke\textsuperscript{−} in dual gain mode. Continued collection of photo-generated charge during readout enables a further dynamic range extension beyond 10\textsuperscript{10} e\textsuperscript{−} effective well capacity with only 1% loss of exposure efficiency by combining short and long exposures. A quadratic fit to correct for non-linearity reduces gain correction residuals from 1.5% to 0.2% in low gain mode and to 0.4% in high gain mode. Cross-talk to adjacent pixels is only 0.4% vertically, 0.6% horizontally and 0.1% diagonally. These characteristics plus the relatively large (10\textmu{}m) pixel size, quasi 4-side buttablility, electronic shutter and sub-array readout make this sensor an excellent choice for wide field astronomical imaging in space, even at FUV wavelengths where sky background is very low.

Keywords: CMOS, SRI mK\texttimes{}nK, ultraviolet, dark current, glow suppression, dual-gain.

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1 Introduction

For decades the detector of choice for ultraviolet (UV) space applications has been the microchannel plate (MCP) coupled to various kinds of 2D charge sensors. Space missions such as GALEX, Neil Gehrels Swift, and Hubble Space Telescope Cosmic Origins Spectrograph adopted MCP imagers because their sensitivity extends into the far-UV (FUV), while the wide bandgap of their photocathode depresses sensitivity to optical photons and allows them to be operated at room temperature. However, MCPs have relatively low quantum efficiency (QE)[1], require high voltages in space where electrical breakdown is a concern, and are damaged by over-illumination. The latter
issue has restricted many astronomical UV telescopes from viewing fields containing bright stars, particularly in the FUV where over-illumination causes severe depletion of the photocathode.

The drawbacks of MCPs have motivated the development of Silicon-based UV imagers (CCD and CMOS) for astronomy, where higher QE provides more sensitivity for a given telescope size, enabling powerful observatories on moderate-sized platforms. The ability to view fields containing bright stars and regions where high dynamic range is required, such as the Galactic plane and Magellanic Clouds, is another significant scientific advantage, especially for wide-field of view (FoV) systems. On the technical side, avoiding high voltages and being able to match pixel sizes to the specific application are additional positive factors. For wide FoV applications, CCDs and CMOS detectors can be assembled in close butted mosaics with ∼85% fill factor, supporting imaging applications on a much larger scale than MCPs[1].

The Ultraviolet Explorer (UVEX) mission, which motivates the investigation documented here, requires two 12K × 12K focal planes, imaging in FUV(1390 − 1900 Å) and NUV (2030 − 2700 Å) passbands at a plate scale of 1.02″ per 10 μm pixel, and a high efficiency Long Slit Spectrometer covering the 1150 − 2650 Å wavelength range, at 0.4 Å/pixel sampling, with a single detector. The sky in the FUV is so dark that even a small amount of readout noise dominates photon shot noise from sky in a typical 900s exposure, so the ideal sensor will have very low read noise (∼2 e⁻) and low dark current (<10⁻³ e⁻/s) at temperatures readily accessible with radiative cooling (170K).

While CCDs can meet some of these requirements at beginning-of-life, CMOS image sensors have very significant advantages. Since charge is converted to voltage within the pixels, there are negligible charge transfer losses, a problem which is particularly severe in CCDs when background is low, and gets worse with radiation damage, requiring complex post processing based on frequent recalibration as demonstrated on HST[2][3]. Also, CMOS imagers do not require a shutter. While
CCDs can also be operated without a shutter, the required subtraction of averaged parallel overscan lines adds shot noise and delivers poor cancellation of image smear in the presence of pointing jitter.

In the CMOS architecture studied here, the 5-transistor pixel design (Figure 2) eliminates dead time between exposures. Charge transfer to the sense node defines the end of one exposure and the start of the next, allowing charge collection to continue during readout without charge smearing. We use Rolling Shutter mode which requires one frame time to clear charge at the start of a sequence of exposures. At only $3\text{s}$ for a $4K \times 4K$ sensor this overhead is much shorter than a typical CCD readout. Subsequent readouts between exposures of the same scene do not represent any overhead. An important application of this is to extend the dynamic range by combining consecutive short and long exposures, e.g. pairing $3\text{s}$ and $300\text{s}$ exposures extends dynamic range by 100 without dead time between exposures due to readout.

Another advantage of CMOS imagers is the flexibility they afford for subarray readout. Since pixels are addressed directly, subarrays can be read at high cadence without disturbing signal integration on the remainder of the sensor. This feature further extends the dynamic range to bright targets which would saturate in the normal frame time, and enables high time resolution studies of rapidly varying sources. For space applications where guide stars are required, subarray readout can provide rapid centroiding of bright stars without interfering with the use of the rest of the device for deep science exposures.
Fig 1: The mK×nK Imager reticle consists of four modular building blocks which are photo-composed (“stitched”) to create custom dimensions in 1k increments. The sizes shown have been successfully produced, for ground and space missions.

In this paper we report on the characterization of an SRI Nk×Mk CMOS sensor[4] made by Sarnoff International for scientific imaging from space. This radiation tolerant design (Table 1) has significant flight heritage (Parker Solar Probe, Solar Orbiter, SOLO HI EIS) [5] [6]. It can be fabricated in 1k increments up to 4k×4k from the same reticles (Figure 1) and the compact address logic allows 4k×4k tiles to be mosaiced with high fill factor, enabling large focal plane areas. The device used in this work has a 4K×2K format, as used on the Europa Clipper mission. Ultimately for large FOV astronomical applications such as UVEX a 4K×4K size is attractive. Such devices have been made successfully with the same reticles for Lawrence Livermore National Laboratory’s National Ignition Facility [4]. All sizes have very similar electrical characteristics since they are fabricated using the same reticles, at the same foundry, using the same processes.

| Radiation Parameter               | Value                                    |
|----------------------------------|------------------------------------------|
| Total Ionizing Dose              | > 100 Krad                               |
| Non-Ionizing Energy Loss         | > 3 × 10^{11} protons/cm^2 @ 63 MeV protons |
| Single Event Latchup             | > 100 MeV cm^2/mg                        |
| Single Event Upset               | > 25 MeV cm^2/mg                         |

Table 1: Radiation tolerance of SRI Nk×Mk, reproduced from data sheet.
2 Provenance of Sensor Under Test

The 4K×2K device used for this study was produced together with identical wafers for the Europa Clipper mission [4]. The Jet Propulsion Laboratory’s Advanced Detectors, Systems, and Nanoscience Group at JPL’s Microdevices Laboratory (MDL) applied their end-to-end post fabrication processing including bonding, thinning (to $7 \pm 0.3 \mu m$), and delta-doping process using Molecular Beam Epitaxy [7] [8] [9] [10] at wafer level. This process has been demonstrated over the years on a number of CCD and CMOS formats to provide high and stable quantum efficiency across the UV and visible spectrum [9]. Every photon not reflected or absorbed in the anti-reflection (AR) coating is both detected and collected into pixels [4].

JPL applied various AR coatings and metal dielectric filters (MDFs), returning several die to SRI for packaging and screening tests. The in-band UV QE and out-of-band rejection obtained with the delta doping and direct deposition of MDFs will be described in a companion paper. An earlier broadband AR-coated version was measured for Europa-Clipper band [11].

Unfortunately, pinhole defects in the photo-resist that protects the device during the etching of back surface to reveal the bond pads caused damage at random locations on the delta doped wafer. An engineering grade device without any coatings, but with tens of these “etch-through” defects was provided to Caltech for evaluation at lower temperatures than previously explored. At these temperatures several strong glow sources became apparent.

We will show that the relatively simple on-board circuitry for row/column decoding (no timing generator or ADCs) allows for sufficient glow suppression to meet the demanding dark current requirement with ample margin (§5.5), while read noise rivals the best CCDs (§5.7). Linearity is good (§5.2), inter-pixel correlation is low(§5.8), and dynamic range is exceptional when CMOS
capabilities for dual gain (§4.3.1) and dual exposure duration are harnessed. Readout overheads are insignificant.

3 SRI mK×nK CMOS detector

3.1 Pixel Topology

Each pixel is made up of a light-sensitive pinned photodiode, a transfer gate, and 4 transistors to reset the sense node and select it during readout (Figure 2). A pixel read cycle begins by resetting the sense node to the reference potential, PIX VREF by closing transistors RESET and MIMSELECT. If MIMSELECT is high during the time charge is sensed, the MIM capacitance is in parallel with the sense node increasing the conversion gain ($e^-/\mu V$) and well capacity at the expense of increased read noise (in $e^-$). By reading out the voltage before and after connecting the MIM capacitor, one can read the same signal in high then low gain, to provide low noise then high signal capacity. Transfer Gate and reset operations are performed for an entire row and cannot be executed on single pixels. This prevents CCD-style Correlated Double Sample (CDS) timing if using off-chip digital subtraction, though in-pixel circuitry is provided for analog subtraction of an entire row concurrently with a conventional short delay between baseline and signal samples.

The detector has the ability to reset and transfer charges from the photodiodes to their sense nodes over the whole array in a single pulse using a signal labelled SNAP. This is used to run the detector in “Global Shutter” mode, resulting in concurrent exposure start and end for all of the pixels of the array. However we exclusively use Rolling Reset mode in which lines are reset and read sequentially as this has yielded better noise performance.
3.2 Output Buffer

The RAW video outputs are buffered by p-type transistors in a Source-Follower topology. We use 3.3 kΩ pull-up resistors to 3.3 V as the load. No other processing occurs: we do not invoke the optional RC filtering.

Alternatively, the CDS video output pins of the detector can be buffered by n-type transistors, also in a Source-Follower topology. In this case an analog CDS circuit embedded in each pixel, stores the reset levels of a row before subtracting the signal obtained after charge transfer. This subtraction is done using switched capacitors and does not deliver the optimum noise performance, so we do not use this analog CDS mode.

4 Experimental Setup

4.1 Cryostat

A cryostat with Liquid Nitrogen (LN2) cooling is configured for operation of the detector at temperatures down to 130 K. A large printed circuit board, dubbed the Vacuum Interface Board (VIB), is located between the front and back sections, sealed with O-rings on each side. The VIB trans-
ports all required signals into the vacuum on internal layers as well as supporting electrical components in vacuum (such as the zero-insertion-force socket housing the detector) and in atmosphere (FPGA, connectors to detector controller, test points, and switches that support trouble-shooting). The detector socket is thermally strapped to the nitrogen tank at 77 K, while the VIB has cut-outs to minimizing conduction to the outside world at room temperature. We tuned the thermal link to the liquid nitrogen tank for 120 K equilibrium temperature without active heating. Detector socket temperature is servo controlled to 140 K with millikelvin stability. Detector electronics are mounted directly on the cryostat in close proximity to the VIB to limit the length of the video path.

![Fig 3: CMOST Camera. The open cryostat is shown with detector inserted into its zero-insertion-force socket. The controller sits on top of the cryostat, and the Vacuum Interface Board provides row and column address generation and routing of video signals.](image)

4.2 Camera Controller

The detector controller is an “Archon” made by Semiconductor Technology Associates Inc\(^1\). It offers versatile configuration including clock and bias drivers, LVDS outputs, and 4 slots, each

\(^1\)http://www.sta-inc.net/archon/
with 4 video processing channels. An intuitive waveform definition language has been added by Caltech Optical Observatories (COO).

Direct addressing of rows and columns is supported by the detector to allow efficient subarray readout or user selectable raster scanning direction. This requires a mix of direct selection of the 1K pixel blocks and binary coding for row selection within those blocks. As this complexity is not supported by the Archon controller, we have added an external FPGA to generate the addresses, a Xilinx Zynq SoC on Picozed board which plugs into the atmospheric side of the Vacuum Interface Board.

4.3 Readout Mode

The detector offers a large range of possible readout schemes thanks to its versatile addressing topology [12]. In Global Shutter mode, one must invoke the in-pixel (analog) CDS processor to avoid having a full frame time between measuring baseline (post reset) level and signal. This uses switched capacitors that introduce additional noise (when reset). To avoid this additional noise source, all performance measurements reported here we have used Rolling Shutter (line by line reset) and digital CDS.

Rolling Reset is widely used in CMOS detectors. Light is accumulating at all times, so the transfer of charge from photodiodes to their sense nodes, defines both the end of one exposure and start of the next. Each line is addressed sequentially. The sense nodes for all pixels in one line are reset concurrently. In our ”line-wise digital CDS scheme”, baseline samples are digitized sequentially for the entire line, then charge is transferred from photodiodes to all sense nodes concurrently for the same line. The resulting signal levels for each pixel are digitized sequentially. The process then repeats for the next line. The difference between the baseline and signal samples
is calculated in the host computer post-facto. 1.5 s is required to scan a frame with 2048 lines at 700 kpix/s, with 16 channels being read in parallel each servicing a block of 256 columns.

The 1.5 s frame time results in a small skew in exposure start times across the image area but this has negligible effect in the long exposures anticipated in our application. When acquiring a sequence of frames there is a 1.5 s overhead to reset the pixels at start of sequence (e.g. after telescope slew) but no dead time between frames within a sequence. Table 2 and Figure 4 describe a multi-frame sequence for a hypothetical $4 \times 4$ pixel array.

With exposure duration being defined by the time between charge transfers, it follows that minimum exposure time is one frame time and longer exposures are executed by inserting a delay between frame scans. Faster cadence exposures can be executed when reading a subset of rows usually scanning between two preset pointers.

The Archon version used for these experiments was originally developed for CCD readout and thus has an AC-coupled differential video input. The negative input is grounded while positive input is connected to either the RAW or the CDS output of the detector selected by a mechanical switch. (We prefer RAW, as noted.) The DC level of the JFET input stage (in the Archon, post AC coupler) is set through an analog switch typically once per row. This level is chosen so that the differential input voltage is near the beginning of the ADC input range.

Each channel is digitized by a separate 16 bit ADC running at 100 MHz. An analog filter limits bandwidth into the ADC only to the extent required to prevent aliasing. The high sample rate allows each channel to be operated in digital oscilloscope mode for diagnostic purposes. When executing digital CDS, noise bandwidth is set by the number of consecutive 100 MHz samples which are averaged per pixel. This averaging occurs concurrently in FPGAs on each video card.

The line-wise digital CDS arrangement avoids the $kTC$ noise incurred when resetting the ana-
| Step | Action                                           | Output                              |
|------|-------------------------------------------------|-------------------------------------|
| 1    | Reset Sense Node                                |                                     |
| 2    | Read Sense Node                                 | Baseline Image Low Gain             |
| 3    | Transfer charge from Photodiode to Sense Node   |                                     |
| 4    | Read Sense Node                                 | Signal Image High Gain              |

Table 2: Row readout sequence in Rolling Reset mode.

log CDS capacitors. The fact that baseline and signal have identical signal path assures good gain matching for optimal rejection of low frequencies. Surprisingly, the 366 µs interval between baseline and signal sample proves to be short enough that low frequency noise sources (bias voltage drift; 1/f noise) do not negate these benefits.

![Fig 4: Readout Diagram of the Rolling Reset Mode.](image)

4.3.1 Dual Gain mode

For best noise performance the smallest possible sense-node capacitance is desired as this increases the voltage change in response to a given charge. However the signal range is then limited by available voltage swing. To allow greater signal range, a MIM capacitor in each pixel can be added in parallel to the sense node capacitance by closing the appropriate analog switch (Figure 2). Since the readout of the voltage does not affect charge on the sense node, it is possible to read the same charge packet first with MIM capacitor isolated then again after it is connected. The sequence of steps for this Dual Gain readout mode is detailed in Table 3.

The key here is to establish separate CDS baselines for both gain states. There is no $kT/C$ noise
Table 3: Row readout sequence to generate a dual-gain image.

| Step | Action                                           | Output                     |
|------|--------------------------------------------------|----------------------------|
| 1    | Reset Sense Node in Low Gain                     |                            |
| 2    | Read Sense Node in Low Gain                       | Baseline Image Low Gain    |
| 3    | switch to High Gain                               |                            |
| 4    | Read Sense Node in High Gain                      | Baseline Image High Gain   |
| 5    | Transfer charge from Photodiode to Sense Node     |                            |
| 6    | Read Sense Node in High Gain                      | Signal Image High Gain     |
| 7    | Switch to Low Gain                                |                            |
| 8    | Transfer charge left behind. **KEY STEP**         |                            |
| 9    | Read Sense Node in Low Gain                       | Signal Image Low Gain      |

problem since any charge trapped when opening the gain-select switch is measured at Step 4 and vanishes when the switch is closed again at Step 7.

Frame time is almost doubled since each line is digitized twice as many times. Due to the added time delay, the difference between the frames produced at Steps 9 and 2 provides a CDS sample which is inferior to the difference between frames produced at Steps 6 and 4. However, this slightly elevated read noise is irrelevant since the low-gain sample is only used for large signals that are shot-noise limited.

5 Performance

5.1 Photon Transfer Curves and Well capacity

To infer well capacity and conversion gain (e⁻/ADU), pairs of identical exposures with increasing integration time are acquired with temporally stable flux and moderate spatial uniformity. We then subtract the two frames of each pair to remove common mode patterns due to illumination or fixed electrical patterns, leaving only the random noise component of the signal. Spatial variance is then calculated for a region chosen to be free of bad pixels or other artifacts such as ROIC glow. This procedure is repeated for both low and high gains.
Conversion Gain ($e^-/ADU$), the inverse of the slope of the linear fit, is calculated separately for each gain mode. These values are then used to convert signal and variance to values displayed in electrons, for both gain states on the same Photon Transfer Curve (Figure 5). This plot format is convenient to immediately highlight the signal level at which shot noise dominates. This measurement yields readout noise of $2.1 \, e^-$ and $12.3 \, e^-$ and a Full Well Capacity of $15 \, ke^-$ and $120 \, ke^-$ in the high gain and low gain modes, respectively.

However, when only spatial variance is used in photon transfer curves, total noise is underestimated due to interpixel correlations, so system conversion gain ($e^-/ADU$) is overestimated. The correct conversion gain is obtained when the covariance with adjacent pixels is included in the photon transfer curve (§5.8). Conversion gain has been corrected for Inter-Pixel Correlation (IPC) throughout this paper resulting in a 6% reduction in dark current and read noise compared to that inferred from simple photon transfer method using variance alone. Corrected conversion gains are $1.13 \, e^-/ADU$ in the high gain and $8.0 \, e^-/ADU$ in low gain mode.
5.2 Linearity

Linearity has been measured in both high and low gain modes by varying the exposure time while the detector is illuminated with a constant flux. To do this, a red LED is attached to the LN2 tank on a separate thermally regulated platform, operating slightly above 77 K. A bench current source injects a current ranging from a few tens to hundreds of nanoamperes leading to a total power dissipation well below a microwatt, resulting in very little efficiency variation due to junction self-heating.

We infer that detector self-heating is responsible for some gain drift as readout cadence influences the level of the video signal. Sensor current consumption peaks at $\sim 70 \text{ mA}$ during readout, dissipating hundreds of milliwatts. This highlights the need for a more thermally conductive package to minimize the change in detector temperature due to self heating. The package in use is far from optimal in this regard so significant performance improvement is probably possible. It is also clearly beneficial to adopt clocking schemes which maintain the readout cadence even when idle to minimize temperature drift.

An initial transient in the signal is seen in Figure 6 after turning on the detector and continuously reading it at a constant cadence. This transient appears only where there is illumination: not in dark areas under a mask. This demonstrates that the CDS processing indeed corrects for offset drift of the RAW video levels but cannot prevent the gain change. We explore this temperature dependence in more detail in §5.3, finding that the video level settles to $1e^-$ after approximately 30 min.

In order to mitigate gain changes due to self-heating during the linearity experiment, the detector was read out at a very low duty cycle (at least 5 min per frame), so that the effect of self-heating
Fig 6: Initial transient after starting the readout of the detector at a constant frame rate. A transient is only visible on the illuminated area, revealing a gain change.

was negligible.

We estimated the non-linearity by fitting an affine function or a quadratic with non-zero value at origin. We acknowledge that this is not ideal, but fixing a zero offset leads to a large mismatch with our fit. The actual detector’s transfer function most likely requires a more complex function than a low order polynomial to produce a better fit. In order to optimize both the residuals during our fitting process and the non-linearity itself, we included a weighting factor, $w_i$, in the estimator $E$, which is defined as

$$E = \sum w_i^2 (p(x_i) - y_i)^2 = \sum \left( \frac{p(x_i) - y_i}{y_i} \right)^2$$  \hspace{1cm} (1)$$

where $y_i$ is the measured signal at exposure time $x_i$, and $p(x_i)$ is the value of the linear fit at exposure time $x_i$. This prevents the gain correction from increasing steeply at the lower end of the non-linearity curve when the denominator becomes smaller.

Figure 7 summarizes the linearity behavior in high gain (1.13 $e^-$/ADU) mode. A linear fit results in an RMS deviation of 1.4%, with larger disagreement for higher signal. Better agreement is achieved with a quadratic fit, yielding an RMS 0.2%. This is consistent with typical CMOS sensors.
where non-linearity originates in both the voltage-dependency of the PN junction capacitance that contributes to Sense Node capacitance, and in the non-linearities of the source-follower itself [13]. The inset plot shows linearity behavior for faint signals (obtained at reduced flux), demonstrating that the fit improves to an RMS deviation of 0.1% and projects to 0 signal for null exposure time. (We must extrapolate to zero-length exposure since minimum exposure time is the frame scan time which is 1.5 s when reading at 700 kPix/s/ch.)

Figure 8 summarizes the linearity behavior in low gain (8.0 e−/ADU) mode. This shows similar behavior: the linear fit yields an RMS value of 1.5%, while the quadratic fit gives better agreement with RMS deviation 0.4%. Note that the faint signal is neglected since low gain mode will not be utilized for very faint signals.

Fig 7: Linearity in high gain mode. Faint signal ranges from 0 e− to 1000 e−.

Fig 8: Linearity in low gain mode. Note that the faint signal is neglected since low gain mode will not be utilized for very faint signals.
5.3 Temperature dependence of the gain

Temperature stability requirements in flight will probably be driven by the temperature dependence of the gain. To evaluate this, we recorded identical flat fields at moderate intensity for temperatures varying between 140 K to 180 K. Light intensity was held constant while regulating LED temperature to millikelvin precision. We allowed a settling time of 30 min between each 5 K step.

Figure 9 shows that the temperature dependence of gain is not linear but gradually flattens as the temperature increases. Around 165 K, the gain varies by 0.15 %/K. It is not possible to determine from this experiment whether the gain dependence on temperature is caused by variation in QE, sense node capacitance, Source-Follower transconductance, or load resistance.

It would be interesting to reproduce this experiment to characterize the temperature dependence of the offset. While the per-row CDS processing removes any slow DC-offset variation, one could track the baseline samples, however, the AC-coupling of our video acquisition board currently prevents such measurement.

![Temperature dependence of the gain](image)

Fig 9: Temperature dependence of the gain between 150 K and 180 K. Around 165 K, the gain varies by 0.15 %/K.

5.4 Multiplexer (Mux) glow

Electronics surrounding the image area (primarily logic and pixel drivers on the left edge of the chip) act as glow sources that become visible in long exposures. Figures 10a and 10b show Dark
Maps generated from the difference of 18 hour and 1 hour exposures to suppress fixed patterns and reveal $\text{me}^-/\text{s}$ effects.

It can be seen that light is emitted by the row select logic to the left of the image and the column select logic well below the bottom of the image. The device under test also had several strong glow sources in the image area to the right and below of the area displayed.

We suppress photo-emission from the address logic by reducing the levels of bias voltages during exposure, where not required in the charge collection process. The only remaining biases are $\text{VDD}$ driving the logic and the Transfer Gate that keeps the charges in the photo-diode. Experience shows that the three biases $\text{V}_{\text{Ref}}$, $\text{V}_{\text{Reset High}}$, and $\text{V}_{\text{MIM High}}$ must be kept on to ensure a valid bias of the Sense Node before readout. Some settling time was added after turning the other biases on again and starting a readout of the image area. This settling time, though dozens of seconds, is negligible compared to the long integration times required for glow to be visible. This glow-mitigation technique greatly increased the area where dark current is less than $1\text{ me}^-/\text{s}$.

5.5 Dark current

Dark current has been measured at temperature between 140 K and 240 K. One hour was allowed for settling between each temperature change. No hysteresis from heating or cooling the detector is observed. We evaluate dark current within a sub-window in the region of the detector minimally affected by glow. In spite of the very low signal, good signal to noise ratio is obtained by using a large number of pixels and very long exposure times.

McGrath et al. [14] list different mechanisms as possible sources of dark current, each having their own temperature dependence. We do not attempt to identify the dominant source, noting only that the dark current decrease follows the expected Arrhenius law, until a a floor is reached at
(a) 18 hours exposure in the dark while keeping electronics in its default state.

(b) “Low glow” 18 hours exposure in the dark while shutting down support electronics.

Fig 10: Dark Maps of the top left quadrant of the detector show fine structures and glow sources in and around the image area. Most of the glow coming from the left edge of the chip where lies the logic and pixel drivers is turned off by lowering the biases during the exposure.
0.08 me$^-$/s for temperatures below 165 K.

Given 2 e$^-$ read noise and 900 s exposure time, the dark current can rise to 1 me$^-$/s before increasing total noise by 10%. These requirements imply that the operating temperature could be as high as 184 K before shot noise on dark current becomes significant.

The striations to the right in Figure 10b are due to light scattering from a bright defect at the lower edge, which we assume will not be present in science grade devices. A dozen or so larger, approximately semicircular, white spots are are not due to cosmic rays but appear to represent weak glow sites of order 2 me$^-$/s/pixel at peak, affecting fewer than 1% of pixels.

Fig 11: Dark current for a wide range of temperatures. A simple fit of an Arrhenius law shows the expected temperature dependency.

5.6 Optimizing Speed and Noise Performance

The STA Archon controller digitizes pixels at a rate of 100 MHz and supports sample averaging in real time for each channel to provide low-pass filtering. This is the digital equivalent of the analog integrator found in traditional CDS processors.

Readout noise tends to increase with higher pixel frequency, as noise bandwidth increases
when there is less sample averaging. Conversely, when pixel rate is low the increased interval between baseline and signal samples results in poorer rejection of low frequency noise processes which begin to dominate. We explored multiple pixel frequencies from 700 kHz to 1.4 MHz for numerous averaging window widths.

For any given pixel time, when the averaging window is too wide, the sampling of the video signal begins before it settles sufficiently, with adverse effects on gain and linearity behavior. Therefore we check Gain and linearity using low level flat illumination to infer changes in system gain and thus the true noise in electrons.

![Pixel Speed-Noise Optimization](image)

*Fig 12: Pixel Speed Noise optimization. For each pixel frequency, an optimum window range is found.*

When the window becomes too wide, the gain decreases and the noise rises. For each pixel frequency, there is therefore an optimal window width at the bottom of each curve in Figure 12. In this paper, we only use the optimum sample width for any given pixel frequency.

5.7 Read noise

CMOS sensors can exhibit significant spatial variation in readout noise, predominantly due to single electron traps close to the channel of the pixel buffer MOSFET. On some pixels, this creates
a bistable video level known as Random Telegraph Noise (RTN), whose amplitude and characteristic time constant vary from pixel to pixel, with characteristic switching rate being dependent on temperature.

In Figure 15 a time series of a pixel exhibiting strong RTN can be seen jumping between three states after CDS processing, representing a negative transition between CDS samples, a positive transition, or no transition.

The noise map in Figure 13 was created by computing the (temporal) standard deviation of each pixel from 94 dark frames.

Fig 13: Noise Map of the detector in high gain. Excess noise is visible in the bottom rows, next to the amplifiers. The bad column generates noise in its surrounding as well as four distinct clusters.

The histogram in Figure 14, was created from data in subset at top left of the temporal noise map, chosen to exclude defective areas. Instead of following the ideal Gamma distribution, the measured distribution is skewed with many pixels being noisier than expected.
Given this highly skewed distribution, it may not be obvious what is the best metric for noise performance. The modal noise is $1.33 \times 10^{-2}$, the median noise is $1.43 \times 10^{-2}$, and the mean is $1.70 \times 10^{-2}$. We propose that a better metric is the square root of the average variance across all pixels not excluded as “too noisy” to be used. The square of this “RMS noise” times the number of pixels in a photometric aperture then represents expected total variance which can be used for making SNR projections. Figure 16 shows the RMS Noise as a function of the number of pixels included. As more pixels are classified as “bad” the RMS Noise improves but at the expense of lost information where interpolation across bad pixels is employed.

Fig 14: Noise distribution of pixels in a clean sub-array of the detector. The ideal gamma distribution describes distribution if noise were distributed normally in the detector.
Fig 15: Fluctuating level of a pixel affected by Random Telegraph Noise. The amplitude of the step is much higher than the read out noise. Note that the size of the step is constant while transition timing is random.

Fig 16: Square root of average noise variance for all pixels included in window, after exclusion of noisiest pixels, plotted for three different pixel rates.

5.8 Inter-Pixel Correlation (IPC) - Cross-correlation

Pixels may couple to their neighbors via two main effects [15][16]: 1) Brighter Fatter Effect (BFE) [17] [18] is the tendency for previously accumulated charge to repel incoming charge, allowing neighboring pixels to collect additional electrons; 2) Inter-Pixel Capacitance leads to voltage coupling between neighboring pixels [19]. The method described below is used to evaluate the total contribution of inter-pixel effects (BFE and Inter-Pixel Capacitance) to cross-correlations with-
out distinguishing them. This total effect will be referred to as the Inter-Pixel Correlation (IPC). A method to distinguish underlying inter-pixel phenomena using correlations in flat fields is described in papers by Hirata et. al. [15] and Choi, et. al [19].

We extracted the impulse response from a difference of two flat frames using a 2D autocorrelation method. To limit errors due to illumination non-uniformity and imperfections within the device itself, the area we used for processing was limited to well chosen $256 \times 256$ pixel sub-arrays. This is only enough to provide a $\sim 0.4\%$ error bar with our results according to [15]. We performed a Monte Carlo simulation using fake shot-noise limited frames with no IPC to confirm this error bar. Forty-eight pairs of frames were then combined to reduce the errors. The normalized kernel of the inferred impulse response is shown in Figure 17.

![Figure 17](image_url)

**Fig 17**: Left: Auto-correlation of the difference of two flat frames. The noise is reduced by averaging $48 \times 256 \times 256$ pixel patches. Right: Impulse response or IPC reconstructed from the Autocorrelation in low gain mode.

We reproduced this analysis for a range of illumination levels (Figure 18). We define the horizontal (vertical) correlation as the ratio of the first horizontal (vertical) neighbors over the central peak of the PSF. Here, the noise associated with each sample is evaluated by computing the stan-
standard deviation of the correlation coefficient of pixels far away from the center of the PSF. Both horizontal and vertical directions display a similar rate of IPC increase with signal, a characteristic associated with decreasing slope in the variance versus mean as signal increases. Both effects are typically associated with BFE but confirmation awaits spot projection tests in which the dependence of lateral charge diffusion on signal contrast can be measured directly. It is relevant to note here that the pixel clock was slowed down to 1.25 kHz to eliminate any inter-channel coupling or bandwidth limit coming from the video acquisition chain.

![Inter Pixel Coupling VS Signal in Low Gain Mode](image)

**Fig 18:** Horizontal and Vertical Correlation of first neighbors in low gain mode.

The classical Variance vs. Mean method [20] assumes pixels are uncorrelated and therefore underestimates the shot noise of the incident photons due to the smoothing effect of the spatial correlation. Figure 19 shows such a Photon Transfer Curve. A similar Variance vs. Mean curve that integrates the power distributed in the wings of the impulse response reveals the true conversion gain. Before correction for correlations, the conversion gain is overestimated by $\sim 6\%$. The corrected Conversion Gains have been used throughout this paper wherever units are in $e^-$. 

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Fig 19: Photon Transfer curve obtained in Low-Gain mode by taking into account the pixel correlation. The total conversion gain drops by 6% from $8.5 \, e^-/ADU$ to $8.0 \, e^-/ADU$.

5.9 Cosmic Ray Statistics

We have examined the distribution of cosmic rays incident on the detector from three 18 hour dark exposures of the CMOS detector in order to confirm that event rates and number of pixels affected are as expected. This information is useful for understanding how particle event sizes (track lengths) are likely to scale if thickness is changed, and to check that the detector package and AR coatings do not produce additional events. Bad pixels were masked to avoid spurious detections, but this masking did not include the noisiest pixels.

Cosmic rays were identified as any contiguous region above a threshold set to 50 e-, well above the median read noise. This resulted in 31,877 events, or 1.45 CR/min/cm$^2$, which is in agreement with the expected 1 CR/min/cm$^2$ [21], with the excess being typical for background radiation from our concrete building based on prior in-house measurements with CCDs. Figure 20 shows four examples of cosmic ray events identified with this method.

For this experiment, the detector was facing the horizon. The incident angle of incoming cosmic rays, $\theta$, was then calculated to follow the distribution shown in Figure 21b. Equation 2
Fig 20: Example of various length cosmic ray events identified by finding contiguous regions above $50 \text{e}^-$ in dark frames.

accounts for the typical angular distribution of muons at the ground according to [21].

$$P(\theta) = \cos^2(\theta - \pi/2) \times \cos(\theta)$$

A diagram of the incident angle in reference to the cosmic ray and detector facing the horizon is shown in Figure 21a. The length $l$ of each Cosmic Ray track can then be related to the incident angle $\theta$ and the thickness $T$ of the detector by $l = T \times \tan(\theta)$.

(a) Reference diagram for incident angle $\theta$ and azimuthal angle $\phi$.

Fig 21: Muons and cosmic rays are preferentially coming from the zenith while our camera system is pointing towards the horizon. Note muons coming with a high incidence onto the detector leave a longer trace than those coming with a normal incidence.

Using this geometric model, we are able to use the cosmic ray length distribution to infer
the thickness of the detector, previously measured optically to be $7.0 \pm 0.3\mu m$. This was done by running a Monte Carlo simulation for the incident angle of cosmic rays, $\theta$, drawn from the distribution in Figure 21b. We calculated the distribution in length of the simulated cosmic ray using the angle and thickness of the detector for 4 different thicknesses: 5 microns, 10 microns, 15 microns, and 20 microns. Histograms of these simulated events were then compared to the measured distribution of track length to see which thickness agrees with the data. We estimate the measured length distribution by taking the diagonal length of the bounding box of each cosmic ray. The length distribution agrees with the expectation that shorter events are more likely, as longer cosmic ray tracks correlate with steeper incident angle. The mean track length is 3.6 pixels. The results of the Monte Carlo simulation are compared to the measured length distribution in Figure 22. The implied detector thickness is closer to 5 microns than to 10 microns, thus only slightly less than the direct optical measurement.

![Fig 22: Comparison of cosmic ray track length distribution to Monte Carlo simulation.](image)

We determine from this length distribution that the total rate of pixels affected by cosmic rays is
4.1 pixel/min/cm² and that the fractional area is $4.1 \times 10^{-4}\%$/min.

Figure 23 shows the charge distribution for all events and just multi pixel events for comparison. We observe that single pixel events produce excess counts near the minimum charge deposition, suggesting that this is due to some pixels having sufficient Random Telegraph Noise (RTN) to exceed the CR detection threshold. We assume the multi-pixel event distribution is the true distribution, and will return to masking pixels with excess RTN later. The modal value of the multi-pixel charge distribution is $175\,e^-/\text{pixel}$ with estimated width $309\,e^-/\text{pixel}$.

Fig 23: Cosmic Ray Charge Distribution for (left) all events and (right) multi-pixel events. The excess counts at lowest charge levels (at left) may be due to RTN and must be investigated further.

98% of cosmic rays are expected to be muons which create linear tracks with known deposition rate[22]. To measure this deposition rate, we plotted the total charge deposited versus track length. This was measured in pixels, and thus is projected onto two dimensions rather than being the true three dimensional track length for each event. The contribution of device thickness is significant only for track lengths less than $\sim 100\,\mu\text{m}$. Figure 24 shows an average $70\,e^-/\mu\text{m}$ charge deposition rate in the longest tracks region ($> 100\,\mu\text{m}$), which is higher than the $27\,e^-/\mu\text{m}$ reported in [22] for comparably thin silicon, but very similar to deposition rate measured with thick CCDs.
Fig 24: Charge vs. track length for cosmic ray events. The estimated slope is printed in $\text{ke}^-/\mu\text{m}$, with uncertainty estimated from the covariance matrix.

6 Summary

The SRI 4K×2K described here is an “engineering grade” device produced from a wafer that was part of a production run for the Europa Clipper mission. In spite of a post-processing error (in the etching step to expose bond pads after backside processing), which produced several severe glow sites, dark current and read noise in the unaffected areas are excellent, rivalling the best CCDs. In addition to being much more radiation tolerant than any CCD, this sensor offers higher exposure duty cycle, electronic shuttering, extended Dynamic Range (dual gain, dual exposure), and the ability to read subarrays (for guiding and/or photometry of very bright stars) while continuing long exposures on the remaining pixels. The device exhibits excellent linearity, low correlation between adjacent pixels and the weak dependence of this correlation on signal bodes well for future measurements of brighter-fatter effect. The low thickness in relation to pixel size minimizes cosmic ray track length, while the narrowness of the tracks suggests low lateral charge diffusion, though this has not yet been carefully measured.

Future tests will include projection and scanning of thousands of sub-pixel spots to characterize charge diffusion, to investigate the brighter-fatter effect, and to verify that photometry does not vary appreciably with sub pixel spot motion. More sensitive image persistence tests are also planned.
Meanwhile work is proceeding (at JPL) to validate coating performance, optimizing both in-band QE and out-of-band rejection. An essential part of this work is the validation of Quantum Yield (the production of more than one electron per photon) and the study of how QY statistics affect SNR as a function of fluence, a property shared by all silicon detectors.

We thank the SRI team for packaging multiple detectors and performing screening tests, and for their advice during our detector characterization effort. We acknowledge the support of JPL and Caltech’s President and Director’s Research and Development Fund (PDRDF Program). We also gratefully acknowledge the collaborative agreement between APL-SRI-JPL that made wafers available for this work.

References

1 O. H. W. Siegmund, J. B. McPhate, T. Curtis, *et al.*, “Development of UV imaging detectors with atomic layer deposited microchannel plates and cross strip readouts,” in *X-Ray, Optical, and Infrared Detectors for Astronomy IX*, A. D. Holland and J. Beletic, Eds., **11454**, 274 – 284, International Society for Optics and Photonics, SPIE (2020).

2 R. Massey, T. Schrabback, O. Cordes, *et al.*, “An improved model of charge transfer inefficiency and correction algorithm for the hubble space telescope,” *Monthly Notices of the Royal Astronomical Society* **439**, 887–907 (2014).

3 R. Massey, “Charge transfer inefficiency in the hubble space telescope since servicing mission 4,” *Monthly Notices of the Royal Astronomical Society: Letters* **409**, L109–L113 (2010).

4 J. Janesick, T. Elliott, J. Andrews, *et al.*, “Fundamental performance differences of CMOS and CCD imagers: part VI,” in *Target Diagnostics Physics and Engineering for Inertial Con-
finement Fusion IV, J. A. Koch and G. P. Grim, Eds., \textbf{9591}, 1 – 17, International Society for Optics and Photonics, SPIE (2015).

5 S. P. Plunkett, R. Howard, D. H. Chua, \textit{et al.}, “The Wide-Field Imager for the Parker Solar Probe Mission (WISPR),” in \textit{AGU Fall Meeting Abstracts}, \textbf{2017}, SH23D–2693 (2017).

6 R. A. Howard, A. Vourlidas, C. M. Korendyke, \textit{et al.}, “The solar and heliospheric imager (SoloHI) instrument for the solar orbiter mission,” in \textit{Solar Physics and Space Weather Instrumentation V}, S. Fineschi and J. Fennelly, Eds., \textbf{8862}, 155 – 167, International Society for Optics and Photonics, SPIE (2013).

7 D. M. Burrows, A. Wolszczan, and A. Moore, \textit{The WSPC Handbook of Astronomical Instrumentation}, vol. 2,3, World Scientific (2021).

8 M. E. Hoenk, P. J. Grunthaner, F. J. Grunthaner, \textit{et al.}, “Growth of a delta-doped silicon layer by molecular beam epitaxy on a charge-coupled device for reflection-limited ultraviolet quantum efficiency,” \textit{Applied Physics Letters} \textbf{61}(9), 1084–1086 (1992).

9 S. Nikzad, A. D. Jewell, M. E. Hoenk, \textit{et al.}, “High-efficiency uv/optical/nir detectors for large aperture telescopes and uv explorer missions: development of and field observations with delta-doped arrays,” \textit{Journal of Astronomical Telescopes, Instruments, and Systems} \textbf{3}(3), 036002 (2017).

10 J. R. Janesick, \textit{Scientific charge-coupled devices}, vol. 83, ch. Delta-doped CCDs. SPIE press (2001).

11 J. J. H. Shouleh Nikzad, Michael E. Hoenk and L. D. Bell, “High performance silicon and iii-nitride-based uv and uv/optical imaging detectors,” \textit{Journal of Instrumentation}, 3–33 (2021).
12 J. Janesick, T. Elliott, J. Andrews, et al., “Mk x Nk gated CMOS imager,” in Target Diagnostics Physics and Engineering for Inertial Confinement Fusion III, P. M. Bell and G. P. Grim, Eds., 9211, 8 – 20, International Society for Optics and Photonics, SPIE (2014).

13 F. Wang and A. Theuwissen, “Linearity analysis of a cmos image sensor,” in Electronic Imaging, Electronic Imaging, 84–90, Society for Imaging Science and Technology (2017). 2017 IS&T International Symposium on Electronic Imaging ; Conference date: 31-01-2017 Through 02-02-2017.

14 D. McGrath, S. Tobin, V. Goiffon, et al., “Dark current limiting mechanisms in cmos image sensors,” Electronic Imaging 2018(11), 354–1–354–8 (2018).

15 C. M. Hirata and A. Choi, “Brighter-fatter effect in near-infrared detectors. i. theory of flat autocorrelations,” Publications of the Astronomical Society of the Pacific 132, 014501 (2019).

16 A. C. Moore, Z. Ninkov, and W. J. Forrest, “Interpixel capacitance in nondestructive focal plane arrays,” in Focal Plane Arrays for Space Telescopes, T. J. Grycewicz and C. R. McCreight, Eds., 5167, 204 – 215, International Society for Optics and Photonics, SPIE (2004).

17 P. Antilogus, P. Astier, P. Doherty, et al., “The brighter-fatter effect and pixel correlations in ccd sensors,” Journal of Instrumentation 9, C03048–C03048 (2014).

18 A. A. Plazas, C. Shapiro, R. Smith, et al., “Laboratory Measurement of the Brighter-fatter Effect in an H2RG Infrared Detector,” pasp 130, 065004 (2018).

19 A. Choi and C. M. Hirata, “Brighter-fatter effect in near-infrared detectors. II. autocorrelation analysis of h4rg-10 flats,” 132, 014502 (2019).

20 J. R. Janesick, Photon Transfer, SPIE (2007).
21 J. Beringer, J. F. Arguin, R. M. Barnett, et al., “Review of particle physics,” Phys. Rev. D 86, 010001 (306–307) (2012).

22 D. Groom, “Cosmic rays and other nonsense in astronomical ccd imagers,” Experimental Astronomy 14(1), 45–55 (2002).

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