Delbruck T, Berner R: Temporal Contrast AER Pixel with 0.3%-Contrast Event Threshold. In ISCAS 2010: 2010:(in press).

Temporal Contrast AER Pixel with 0.3%-Contrast Event Threshold

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ABSTRACT

Bioluminescence analysis methods require measurement of small temporal variations of scene brightness over a 2d spatial array. Conventional solutions use low-noise frame-based image sensors and high resolution ADCs to achieve the required sensitivity to small fluctuations of less than 1%. Here we report a pixel design for address event representation (AER) temporal contrast detection that is optimized for detecting small relative changes of intensity. Detected changes exceeding a threshold asynchronously generate events. The pixel uses two successive gain stages to memorize and amplify changes of intensity, allowing detection of changes as small as 0.3% of intensity; a factor of about 50 better than prior capabilities for event-based temporal contrast sensors.

Keywords: AER, DVS, silicon retina, neuromorphic, vision sensor, bioluminescence, spike-based, event-based

1. INTRODUCTION

Some applications in neuroscience and bio-sensing are based on imaging florescence or other luminance changes in neurons and cells. Typically these fluctuations are quite small: In voltage sensitive dye recording, for example, they can be a less than 1% [1]. Calcium-sensitive florescence fluctuations can be substantially larger, on the order of 10-20% [2]. The time scale of fluctuations ranges from ms to seconds. Conventional approaches for measuring the fluctuations rely on the use of monochromatic low noise image sensors whose output is quantized at high voltage resolution and frame rate, and subsequently digitally processed. The high data rate and required post-processing limits potential application in portable or freely moving setups. The aim of the work presented here is to enable focal plane detection and event generation for very small relative intensity changes.

In previous work, we built a high performance dynamic vision sensor (DVS) which operates on the same basic principles as reported here [3, 4]. DVS pixels respond asynchronously to relative intensity change (temporal contrast) by emitting streams of binary ON and OFF pixel address-events which signal quantized increases and decreases of log intensity. The original DVS is limited to detect minimum contrast changes of 10-15% by pixel threshold matching variations. The sensitive DVS (sDVS) pixel proposed here adds an additional gain stage and a new mechanism to store past illumination values to increase the pixel sensitivity. In contrast with [5], it uses only a single capacitive feedback amplifier rather than two such stages, which should decrease pixel area.

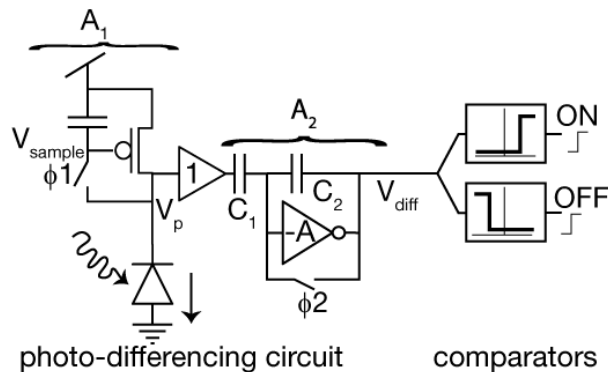

Fig. 1 Principle of circuit operation.

The rest of this paper presents design, implementation and characterization results on a single test pixel.

2. PIXEL DESIGN

Fig. 1 shows the principle of the sDVS pixel operation. It consists of a 2-stage differencing amplifier (A_1 and A_2) followed by ON and OFF event comparators. Each amplifier memorizes its input on the emission of an event and amplifies the change in the input to the output. When the overall change signal exceeds either the ON or OFF threshold, an event is generated that restarts the process.

3. TEST CHIP

The sDVS pixel is built as a test structure in a 0.5um 2P 3M process, together with other structures (Fig. 2). Local analog buffers (Buffers) and an analog mux (aMUX) bring out internal pixel signals and a shared digital output (AER) brings out the pixel events. A global metal 3 shield

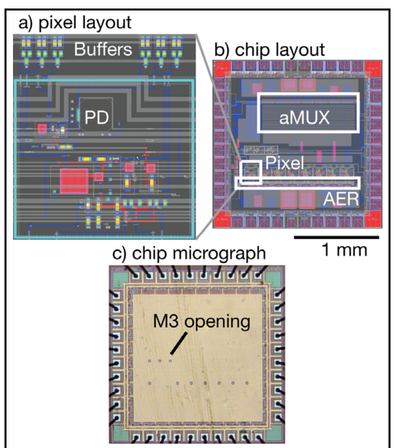

Fig. 2 Layout. a) pixel layout; b) chip layout; c) chip micrograph.

covers everything but the photodiodes. No effort was made to minimize layout area on this test chip since the primary interest was to study effects not modeled well by SPICE simulation, such as charge injection and junction leakage.

4. DETAILED PIXEL OPERATION

Fig. 3 shows the detailed pixel schematic. The first stage transimpedance amplifier is formed by the photodiode PD and pfet load M_{pu}. Closing switch M_{sw1} forms a diode-connected load for PD photocurrent, and the gate of M_{pu} comes to equilibrium where it sources the photocurrent. Opening M_{sw1} memorizes the photocurrent I_{bg} on M_{pu}. Subsequent changes in photocurrent are converted to voltage V_p by the drain conductance of M_{pu} in parallel with PD's Early effect conductance g_{pd}, which is due to its voltage-dependent junction width modulation. Just as in MOSFET channel length modulation, g_{pd} is approximately proportional to current. The M_{pu} drain conductance is also increased by the Miller effect of parasitic capacitive feedback to the floating gate of M_{pu} through its gate-drain overlap capacitance C_m. The net result is that after M_{sw1} is opened, a subsequent change ΔI in photocurrent away from its memorized value I_{bg} is transduced to a voltage ΔV_p:

$$\Delta V_p = \Delta I / g_p \tag{1}$$

where g_p is the conductance looking into V_p:

$$g_p = g_{pu} + g_{pd} + g_{miller} \tag{2}$$

and where g_{pu} is the M_{pu} drain conductance, g_{pd} is the photodiode conductance, and g_{miller} is the Miller effect capacitive feedback from V_p to V_{pu}. All three conductances are proportional to I_{bg}:

$$\begin{gathered} g_{pu} = I_{bg}/V_{Epu}, \quad g_{pd} = I_{bg}/V_{Epd} \\ g_{miller} = \frac{\kappa C_m}{C_m + C_{pu}} \frac{I_{bg}}{U_T} \end{gathered} \tag{3}$$

where V_{Epu} and V_{Epd} are the M_{pu} and PD Early voltages and κ is the back-gate coefficient. Since the total conductance is proportional to I_{bg}, the response to a fixed contrast ($\Delta I/I_{bg}$ or $\Delta \ln(I)$) is invariant to absolute illumination:

$$\frac{dV_p}{d\ln(I)} = \frac{dV_p}{\Delta I / I_{bg}} = \frac{\Delta I}{g_p} \equiv \frac{\Delta I}{I_{bg}} V_E \equiv \frac{\Delta I}{I_{bg}} A_p U_T \tag{4}$$

where we define V_E as the equivalent "Early voltage" looking into V_p and A_p as V_E's multiple of the thermal voltage U_T. As in any logarithmic detector, the response signifies input temporal contrast, rather than absolute illumination change. This invariance is desirable because it removes much of the effect of spatially varying scene illumination and exposes the reflectance changes instead. A_p is the dimensionless gain of this logarithmic detector.

V_p is buffered by a source follower to V_{sf}, which drives the second stage differencing amplifier A_2, which amplifies changes in V_{sf} by a controlled gain to produce V_{diff}. The second stage differencing amplifier is identical to the one in [3]. The capacitor ratio C_1/C_2=10 sets the gain. V_{diff} is compared to its reset level by the ON and OFF comparators (M_{ONp}, M_{ONn}, M_{OFFp}, M_{offn}). The OFF comparator is inverted once more to form nOFF. In reset, nON and nOFF are both high with don<diff and doff>diff. One of them goes low for sufficiently large changes of V_{diff}. The logic low nON and nOFF signal levels start the communication cycle from the pixel to the peripheral pixel address communication circuits, which are not shown here.

The ϕ_1 and ϕ_2 switch timing must be such that $\tau_1 < \tau_2$, so that M_{sw1} opens and V_p settles before M_{sw2} opens to store V_p on the input of the second stage amplifier. If τ_2 is too short, then residual changes in V_p will be amplified by the second stage, resulting in spurious events. Signals ϕ_1 and ϕ_2 are generated by starved NOR gates that OR row and column acknowledge signals. After both acknowledge signals are removed, ϕ_1 and ϕ_2 slew towards Vdd at a rate set by the refr1 and refr2 biases (Fig. 3 and Fig. 4).

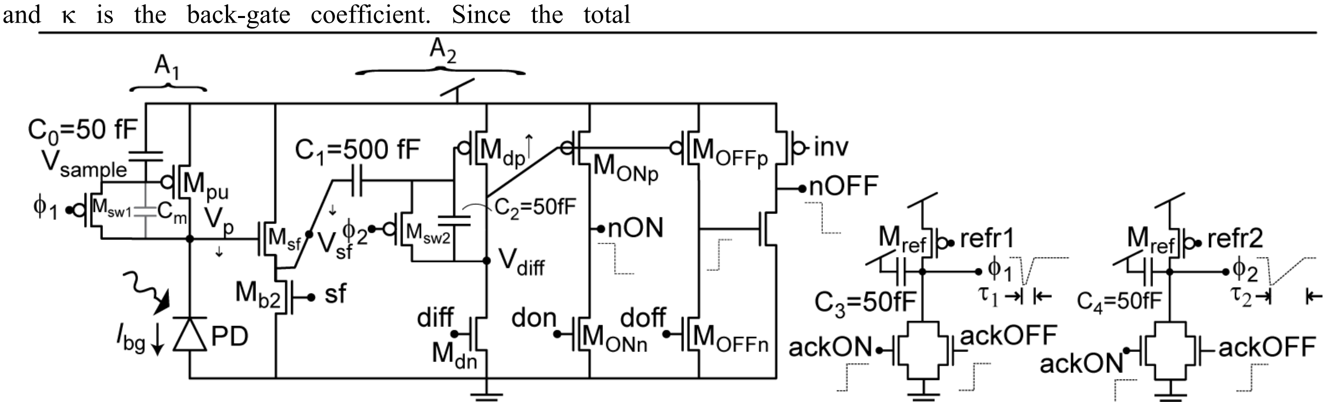

Fig. 3 Detailed pixel schematic.

5. CHARACTERIZATION

The pixel was tested by mounting a red LED over the chip so that external light was excluded. The LED was powered so that the voltage across it could be slightly modulated by an HP33120A function generator. Neutral density filters (Kodak Wratten #96) were slipped between the LED and chip to vary the overall illumination. Absolute and relative illumination levels were measured with a Tektronix J17 photometer using the J1812 irradiance head. The nominal LED irradiance onto the chip was about 3 W/m^2, corresponding to bright office lighting levels shining directly onto the die. Digital AER output from the pixel was captured by a USB board together with the function generator sync output using a SiLabs C8051F320-based AER capture board..

Fig. 4 shows pixel internal signal responses during generation of an event. The reset signals ϕ_1 and ϕ_2 are pulled low by each event and slew towards Vdd with rates set so that the second stage amplifier is held in reset until the first stage has come to equilibrium.

Fig. 5 shows how with sinusoidally varying input the pixel outputs ON events while the illumination increases and OFF events while it decreases.

With static input, junction leakage from the source diffusion of M_{sw1} to Vdd slowly turns off M_{pu}. This causes V_p to drop, simulating an increase of photocurrent. Eventually a resulting ON event is generated, starting the process over again.

Fig. 6 shows the digital event outputs of the pixel in response to a very small contrast square wave illumination with 0.3% contrast (max/min=1.003) at irradiance of 3W/m^2.

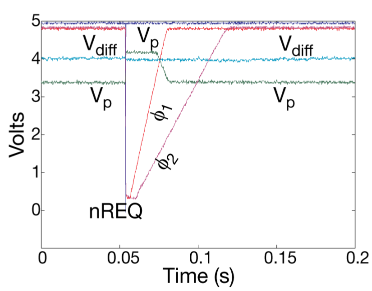

Fig. 4 Pixel internal signals during generation of a single ON event. nReq is the pixel event handshake signal.

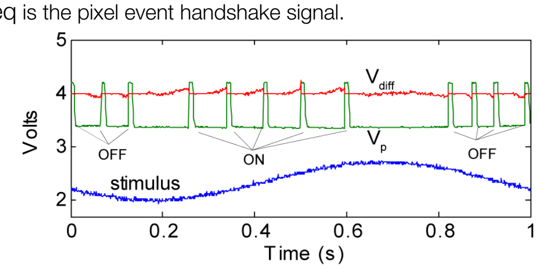

Fig. 5 Internal pixel signals in response to a single sinusoidal cycle of varying intensity. Rising *stimulus* increases light intensity, resulting in ON events; falling *stimulus* results in OFF spikes.

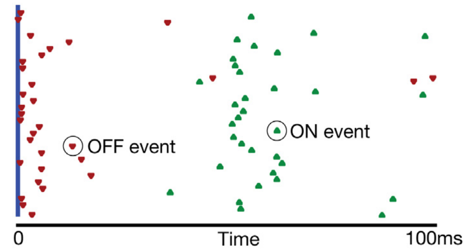

Fig. 6 Spike raster responses to repeated square wave stimulus; each row is a repeat of the square wave. Stimulus frequency is 10 Hz, stimulus contrast is 0.3%, and stimulus irradiance is 3W/m^2.

A logarithmic pixel gain is measured in terms of contrast, or equivalently, relative change. The measured contrast gain of A_1 is 3350 mV/e-fold, so a contrast of 1% (step ratio of 1.01) results in ΔV_p=33.5 mV. This gain corresponds to A_p=134. The measured gain of A_2 is about 6; it is smaller than 10 because the source follower buffer gain is reduced by back gate effect. Thus the overall contrast gain A_1A_2=20 V/e-fold. Therefore a comparator threshold of 50 mV is referred back to an equivalent contrast threshold of 0.25%, in rough accordance with our measurements of achievable minimum event threshold for reliably detecting changes.

Modulation of firing time of the background ON events still can be used to detect the modulation of illumination even when the modulation is too small to directly generate events. The modulation of the input must be at a lower frequency than the background firing rate. Fig. 7 shows the interspike intervals (ISIs) of the background ON events together with points marking the zero-crossing phase of a sinusoidal stimulus with relative intensity change of 0.3%, i.e., from normalized intensity of minimum 1 to maximum 1.003. The ON and OFF thresholds are set so that events are not normally generated. The modulation of ISI, although noisy, is clearly visible over single cycles of the stimulus. Our experimental setup could not produce a smaller variation of photocurrent to test still smaller contrasts.

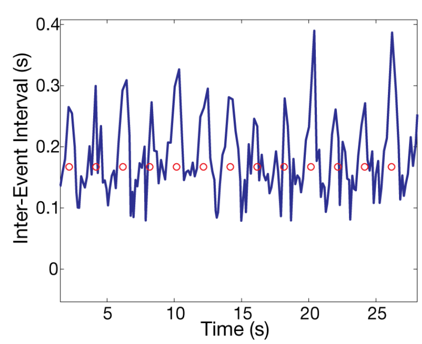

Fig. 7 Inter-event interval modulation in response to 0.3% subthreshold input variation at mean irradiance of 3W/m^2. Circles mark the start of each phase of the sinusoidal variation of illumination.

5.1 Noise analysis

The dominant sources of noise are the photon shot noise, transistor M_{pu} drain current noise, and the shot and flicker noise of the junction leakage of switch M_{sw1}. Likely also important, but not considered here, is 1/f noise in M_{pu}. Considering only the "fast" shot noise, the total input referred contrast noise in A_1 can be computed by treating both M_{pu} and PD as shot noise sources with power spectral density (PSD) of $S_f(I) = 2qI_{\mathrm{bg}}$ [6, 7]. Summing these two noise current sources at V_p, dividing by g_p^2 to obtain the V_p PSD, and integrating over V_p's first order lowpass spectrum with cutoff frequency $f_p = I_{\mathrm{bg}} / 2\pi A_{\mathrm{p}} C U_{\mathrm{T}}$, we obtain the total input-referred noise power as

$$\sigma^2_{\Delta I / I_{\mathrm{bg}}} = \frac{q}{A_{\mathrm{p}} C U_{\mathrm{T}}} \qquad (5)$$

where q is the electron charge, A_p is the gain defined in Eq. (4), C is the V_p node capacitance, and U_T is the thermal voltage. This result is expected as a direct result of the gain-bandwidth tradeoff. In [7], we showed that a "unity gain" source follower photoreceptor with $A_{\mathrm{p}} = 1$ has input-referred contrast noise of $q / C U_{\mathrm{T}}$; here the gain is A_p times higher and thus the bandwidth is A_p times smaller. The total noise is constant and is spread over a bandwidth proportional to intensity, as seen in direct measurements of the PSD of V_p (Fig. 8).

Using measured and estimated circuit parameters $A_{\mathrm{p}} \approx 134$ and $C \approx 110$ fF, we obtain from Eq. (5) $\sigma_{\Delta I / I_{\mathrm{bg}}} \approx 0.07\%$, substantially lower than our measured minimum event threshold of 0.3%. Our measurement is based on event detection with high reliability, which would imply some multiple of the 1-sigma noise estimate of Eq. (5). However, from (5), we can also obtain the V_p output noise as $\sigma^2_{V_p} = q A_{\mathrm{p}} \mathrm{U}_{\mathrm{T}} / C \cong (2.2\ \mathrm{mV})^2$ which accords with measurements of $\sigma_{V_p} = 2$ mV. The measured σ_V is also fairly independent of illumination; over 3 decades it changes from 2 mV down to 1.7 mV. Therefore we conclude that other noise sources such as 1/f noise or power supply coupling limit our contrast sensitivity.

6. DISCUSSION

Although the new pixel increases DVS pixel sensitivity by about a factor of 50 (to 0.3% from 15%), a limitation of the present design is the long time required to reset the PD node after each event (Fig. 4). This is a direct consequence of the gain-bandwidth tradeoff at V_p; the high gain achieved here comes at the expense of bandwidth. Typically the photocurrent is rather small (e.g. 1pA) and after charge injection of M_{sw1}, V_p requires time to settle. This *RC* time can be many milliseconds under low illumination, as is often the case in practical scenarios for florescence microscopy. During this period, the second

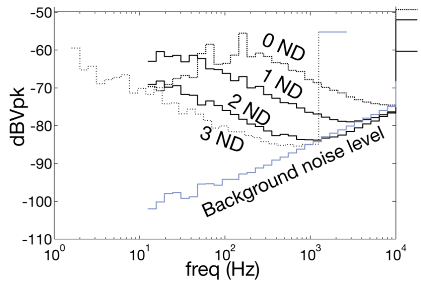

Fig. 8 Octave band noise spectra of V_p for various background illumination levels. The number *X* of decade neutral density filters is *X* ND. The vertical scale is power/octave; it rises for the background flat noise spectrum.

differencing amplifier must be held in reset. If it is not, then new events are generated in response to the settling of Vp. This same requirement also impacts the dynamic range of the circuit. As the light intensity decreases, the settling time also increases, because the conductance looking into Vp is proportional to the photocurrent. For an array of pixels, the refractory periods would need to be adjusted to handle the settling requirements of the darkest pixels. Clearly, a faster input stage which uses active transimpedance amplification (holding PD at a virtual ground) combined with reduction of the switch charge injection will improve this pixel design, at the cost of shorter integration time and hence higher shot noise limit.

7. ACKNOWLEDGEMENTS

Supported by Swiss National Science Foundation grant 200021-112354/1, the Univ of Zürich, and ETH Zürich,

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