1/f Noise Measurement in CMOS Image Sensors

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Abstract— **This paper describes an in-situ pixel source follower power spectral density (PSD) measurement method that does not require any specialized test equipment. This method requires a dual port CMOS image sensor with analog outputs that allows for differential time series noise measurements. We describe the sensor circuits and measurement techniques used for collecting data. We derive an estimator for the PSD based on the measured data. We also present a technique for estimating the confidence interval of the the PSD, based on Bootstrap re-sampling. Using our estimate of the PSD, we derive estimators for the BSIM3 1/f noise model parameters AF and KF. We also determine confidence intervals for these estimators. Using these techniques we present source follower PSD measurements for a CMOS image sensor fabricated in a 0.18μm CMOS process with 3.3μm×3.3μm pixels. We also present the calculated values of AF and KF based on the measured PSD.**

I. INTRODUCTION

Pinned photodiodes and feedback–based reset noise reduction techniques have significantly reduced read noise in CMOS image sensors [1], [2]. In order to further reduce read noise in these sensors, detailed characterization of the residual read noise is essential.

Read noise in pinned photodiode and other low reset noise pixels is typically dominated by 1/f noise caused by the pixel source follower [2]. This is especially true as pixel size scales down allowing very little area for the source follower transistor. In order to reduce this noise component, precise measurement methods are required to estimate the power spectral density (PSD) of the pixel source follower transistor.

Measurement of 1/f noise in MOS transistors is well understood, but it typically requires special test structures and equipment [4]. In addition, the electrical and geometrical conditions used for measuring 1/f noise in test structures are often significantly different from the actual pixel source follower transistor under normal operation. 1/f noise measurements are often used to estimate parameters for a specific noise model. For example, the SPICE NLEV3 gate referred 1/f noise model for a MOS transistor is $S_{V_g}(f) = \frac{KF}{(C_{ox}LWf^{AF})}$, where AF and KF are process dependent noise parameters [5]. In this model AF and KF are very sensitive to electrical bias and layout dimensions. Therefore, AF and KF should be estimated using data from transistors with the identical size and bias conditions as the pixel source follower transistor during sensor read–out.

This paper introduces an in–situ Differential pixel Source Follower spectral Noise (DSFN) measurement method that does not require any specialized test equipment. This method requires a dual port CMOS image sensor with analog outputs to allow differential time series noise measurement. We present the sensor circuits and measurement techniques used for collecting data. We derive an estimator for the PSD based on the measured data. We also present a technique for estimating confidence intervals of the PSD estimator based on the Bootstrap [6]. Pixel source follower PSD measurement results are presented for a CMOS image sensor fabricated in a 0.18um CMOS process with 3.3μm×3.3μm pixels.

II. THEORY

The goal of this paper is to estimate the gate referred PSD of the pixel source follower transistor in a standard 3T [7] or 4T [2] CMOS image sensor, named **M2** in Figure 1. Unfortunately, there are other circuits in the read–out path that add noise, both internal and external to the sensor. In order to mitigate these effects and isolate the source follower transistor noise, we have developed the DSFN measurement method. Figure 1 depicts the architecture of a typical dual port 3T CMOS image sensor. Although the DSFN method can be used with any multi-port CMOS image sensor with differential analog outputs, in this paper we will only describe the measurement method for a dual port sensor. Two output ports are required to produce a differential noise measurement of the pixel source follower transistor. Since the signal path for each pixel is single–ended, we simultaneously measure two pixels to form a true differential signal path, allowing a significant reduction in both internal and external common mode noise. The DSFN measurement method is based on taking four differential measurements from the same pixel pair. These measurement are appropriately scaled and combined to estimate the PSD of the pixel source follower transistor.

Figure 2 shows the simplified read–out circuit used to present the DSFN measurement method. Assuming that SS and SR are always closed, that $RESET > VPIXEL - V_{th}$[1], and that the source follower and the output amplifiers are linear, the circuit read–out path can be further simplified as shown in Figure 3. Note that the voltage gain of the pixel source follower is A_{v1} and the gain of the output amplifiers is A_{v2}. In this model we assume that all of the noise sources $V_{n1}(t), ..., V_{n6}(t)$ are uncorrelated wide sense stationary (WSS) random processes. [2] We also assume that $V_{n1}(t)$ and $V_{n4}(t)$ are independent identically distributed (IID) random processes. Four signal paths are measured to estimate the PSD of each pixel pair. These signal paths are

$$\begin{aligned} M_1(t) = SOUT1(t) - SOUT2(t) = \\ A_{v1}A_{v2}(V_{n1}(t) - V_{n4}(t)) + A_{v2}(V_{n2}(t) - V_{n5}(t)), \quad (1) \end{aligned}$$

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[1] V_{th} is the threshold voltage of transistor **M1**.

[2] The SPICE NLEV3 1/f noise model is not stationary, but the Van der Ziel 1/f noise model is stationary due to its assumption that 1/f noise is the linear sum of Lorentzian processes, where each process has a different parameter and weighting factor [8].

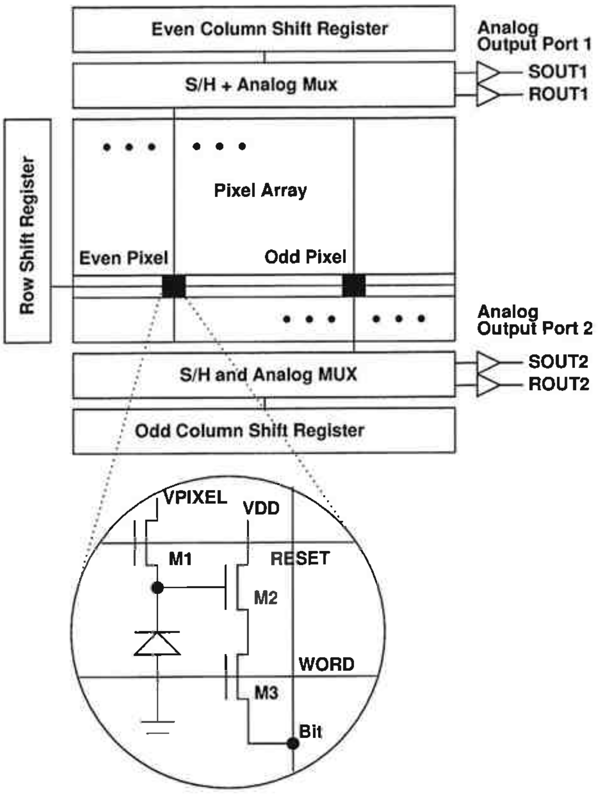

Fig. 1. CMOS Sensor Block Diagram

$$M_2(t) = ROUT1(t) - ROUT2(t) = \\ A_{v1}A_{v2}(V_{n1}(t) - V_{n4}(t)) + A_{v2}(V_{n3}(t) - V_{n6}(t)), \quad (2)$$

$$M_3(t) = SOUT1(t) - ROUT1(t) = \\ A_{v2}(V_{n2}(t) - V_{n3}(t)), \quad (3)$$

and

$$M_4(t) = SOUT2(t) - ROUT2(t) = \\ A_{v2}(V_{n5}(t) - V_{n6}(t)). \quad (4)$$

The PSD of each signal path is by definition the Fourier transform of the autocorrelation function of each signal path. It can be shown that the PSDs of $M_1(t)$, $M_2(t)$, $M_3(t)$ and $M_4(t)$ are

$$S_{M_1}(f) = A_{v1}^2 A_{v2}^2 (S_{V_{n1}}(f) + S_{V_{n4}}(f)) + \\ A_{v2}^2 (S_{V_{n2}}(f) + S_{V_{n5}}(f)), \quad (5)$$

$$S_{M_2}(f) = A_{v1}^2 A_{v2}^2 (S_{V_{n1}}(f) + S_{V_{n4}}(f)) + \\ A_{v2}^2 (S_{V_{n3}}(f) + S_{V_{n6}}(f)), \quad (6)$$

$$S_{M_3}(f) = A_{v2}^2 (S_{V_{n2}}(f) + S_{V_{n3}}(f)), \quad (7)$$

$$S_{M_4}(f) = A_{v2}^2 (S_{V_{n5}}(f) + S_{V_{n6}}(f)), \quad (8)$$

and it can also be shown that the PSD of $V_{n1}(t)$ is

$$S_{V_{n1}}(f) = \frac{(S_{M_1}(f) + S_{M_2}(f)) - (S_{M_3}(f) + S_{M_4}(f))}{4A_{v1}^2 A_{v2}^2}. \quad (9)$$

where $S_{V_{nx}}$ is the PSD of the random process V_{nx}, and assuming $S_{V_{n1}}(f) = S_{V_{n4}}(f)$.

We assert that the PSD of V_{n1} is a tight upper bound on the PSD of the pixel source follower transistor **M2**. This assertion is correct for most designs, because the reset transistor **M1** and the access transistor **M3** are operated in the linear region[3] when measurements are made, and therefore contribute very little 1/f noise to the estimated PSD. This observation can be clearly seen in the Van der Ziel 1/f noise model for MOS transistors [8]. In addition, the source follower bias transistor can be made large enough to contribute little 1/f noise. Note that the noise from each transistor in each pixel is assumed to be independent.

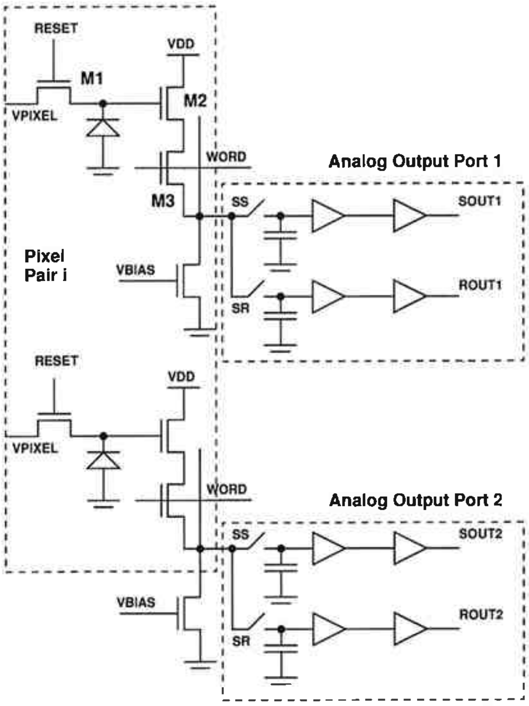

Fig. 2. Simplified Read–out Circuit

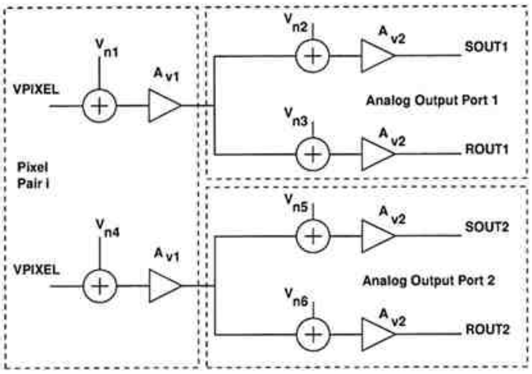

Fig. 3. System Noise Model

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[3] A MOS transistor is operated in the linear region when the gate to source voltage minus the transistor threshold voltage is much greater than the drain to source voltage.

$S_{V_{n1}}(f)$ for each pixel pair i is estimated by measuring the four signal paths $M_1(t)$, $M_2(t)$, $M_3(t)$, $M_4(t)$, and then estimating the PSD of each using

$$\overline{S_{M_{z,i}}(f)} = \frac{1}{P}\sum_{k=0}^{P-1}\frac{1}{T}\left|\frac{1}{N}\sum_{l=0}^{N-1} M_{z,i,k}(l\Delta t)e^{\frac{-j2\pi l w\Delta t}{T}}\right|^2, \tag{10}$$

where z is the signal path, N is the number of samples in each measurement k, P is the number of independent measurements, Δt is the time between each sample, T is the entire sampling period, $f = \frac{w}{T}\ \forall\ 0 \leq w \leq N-1$, and $M_{z,i,k}(l\Delta t)$ is a specific measurement sample. The estimated average PSD of the two source follower transistors in pixel pair i is

$$\overline{S_{V_{n1},i}(f)} = \frac{(\overline{S_{M_{1,i}}(f)}+\overline{S_{M_{2,i}}(f)})-(\overline{S_{M_{3,i}}(f)}+\overline{S_{M_{4,i}}(f)})}{4A_{v1}^2A_{v2}^2}. \tag{11}$$

The Bootstrap[4] is used to determine the confidence intervals for both $\overline{S_{V_{n1},i}(f)}$ and $\overline{S_{V_{n1}}(f)}$ by estimating their cumulative distribution functions F_i and G respectively. To explain the Bootstrap method of estimating PSD confidence intervals we need to develop additional notation. Let the set of P independent PSD estimates for pixel pair i be $\mathbf{x} = (S_{V_{n1},i,0}(f), ..., S_{V_{n1},i,P-1}(f))$ and a randomly re-sampled set of $\mathbf{x}$ with P members be $\mathbf{x}^*$, i.e. a set generated by randomly sampling $\mathbf{x}$ with replacement P times. Rewriting $\overline{S_{V_{n1},i}(f)}$ in terms of $\mathbf{x}$ we have

$$\overline{S_{V_{n1},i}(f)} = \frac{1}{P}\sum_{k=0}^{P-1} x_k, \tag{12}$$

where x_k is the kth member of the set $\mathbf{x}$, and therefore the Bootstrap replication of $\overline{S_{V_{n1},i}(f)}$ is

$$\overline{S_{V_{n1},i}(f)}^* = \frac{1}{P}\sum_{k=0}^{P-1} x_k^*. \tag{13}$$

The cumulative distribution F_i is estimated using a set of R Bootstrap replications $\overline{S_{V_{n1},i}(f)}^*$.[5] Note that each Bootstrap replication consists of randomly selecting a new $\mathbf{x}^*$ from $\mathbf{x}$ and calculating $\overline{S_{V_{n1},i}(f)}^*$. The $1-2\alpha$ confidence interval of $\overline{S_{V_{n1},i}(f)}$ is estimated by $[\overline{F}_i^{-1}(\alpha), \overline{F}_i^{-1}(1-\alpha)]$, where $\overline{F}_i$ is the Bootstrap estimate of F_i.[6] Let the set of Q pixel pair PSD estimates be $\mathbf{y} = (\overline{S_{V_{n1},0}(f)}, ..., \overline{S_{V_{n1},Q-1}(f)})$ and a randomly re-sampled set of $\mathbf{y}$ with Q members be $\mathbf{y}^*$. Then the Bootstrap replication of $\overline{S_{V_{n1}}(f)}$ is

$$\overline{S_{V_{n1}}(f)}^* = \frac{1}{Q}\sum_{i=0}^{Q-1} y_i^*. \tag{14}$$

The cumulative distribution G is estimated based on R bootstrap replications, and the $1-2\alpha$ confidence interval is estimated by $[\overline{G}^{-1}(\alpha), \overline{G}^{-1}(1-\alpha)]$, where $\overline{G}$ is the Bootstrap estimate of G.

The SPICE NLEV3 1/f noise parameters AF and KF are estimated using a least squares fit between the single sided PSD of $\overline{S_{V_{n1}}(f)}$, i.e. $2\overline{S_{V_{n1}}(f)}\ \forall\ \frac{1}{T} \leq f \leq (\frac{N}{2}-1)\frac{1}{T}$, and $\frac{KF}{C_{ox}LWf^{AF}}$. Moreover, assuming that

$$2\overline{S_{V_{n1}}(f)} \approx \frac{KF}{C_{ox}LWf^{AF}}\ \ \forall\ \frac{1}{T} \leq f \leq (\frac{N}{2}-1)\frac{1}{T}, \tag{15}$$

AF and KF are estimated based on the following minimization

$$\min_{AF,KF} \sum_{\frac{1}{T}\leq f\leq(\frac{N}{2}-1)\frac{1}{T}} (X(f))^2, \tag{16}$$

where

$$X(f) = \log(2\overline{S_{V_{n1}}(f)}) + \log(C_{ox}WL) - \log(KF) + AF\log(f) \tag{17}$$

The $1-2\alpha$ confidence intervals of AF and KF are estimated using R bootstrap replications of $\overline{S_{V_{n1}}(f)}^*$ followed by a least squares fit of the data.

III. RESULTS

In this Section we present results from a standard 3T 0.18μm 4M1P CMOS image sensor with 3.3μm×3.3μm pixels. The pixel read–out circuitry is the same as shown in Figure 2, with transistor sizes M1=0.35μm/0.7μm (W/L), M2=0.35μm/0.48μm, M3= 0.35μm/0.35μm. Following the SS/SR multiplexer, the $SOUT$ and $ROUT$ signals are driven by two simple source follower circuits which are represented in the figure. Note that the read–out circuitry is identical for $SOUT1$, $SOUT2$, $ROUT1$, and $ROUT2$.

During each measurement, the appropriate outputs of the sensor were connected externally to a two stage high gain, $A_v = 1000$, low noise differential amplifier with a bandwidth of approximately 25KHz. The output of the differential amplifier was connected to a 16 bit ADC. Data captured by the ADC was transferred to a PC and processed using Matlab.

All of the measurements were performed at an ambient temperature of 25°C with $VDD = VPIXEL = 2.0$V, and $SS = SR = RESET = 3.3$V. An Agilent E3631A power supply was used to generate the bias and control voltages. No special electromagnetic shielding was used during the measurements. For each pixel, four signal paths were measured, $M_1(t)$, $M_2(t)$, $M_3(t)$, and $M_4(t)$, and each signal path was measured independently $P = 201$ times. The measurement of each signal path consisted of $N = 5000$ samples with a sampling period of $\Delta t = 20\mu$s, and a measurement period $T = 100$ms. $Q = 6$ independent pixel pairs were measured. The system gain $A_{v1}A_{v2}$ was directly measured by observing the change in $SOUT1$ as a function of $VPIXEL$ for small input signals.

Figure 4 shows the single sided gate referred PSD of a selected pixel pair with its 95% confidence interval. $R = 1024$ Bootstrap replications were used to estimate the confidence interval.

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[4] A good introduction to the Bootstrap is presented by Efron and Tibshirani in [6].

[5] The error in the estimated cumulative distribution function is an inverse function of R, and R typically must be greater than 1000 for reasonably reliable results.

[6] The Bootstrap confidence interval can be further refined using the BC_a method [6], but at a cost of additional computation.

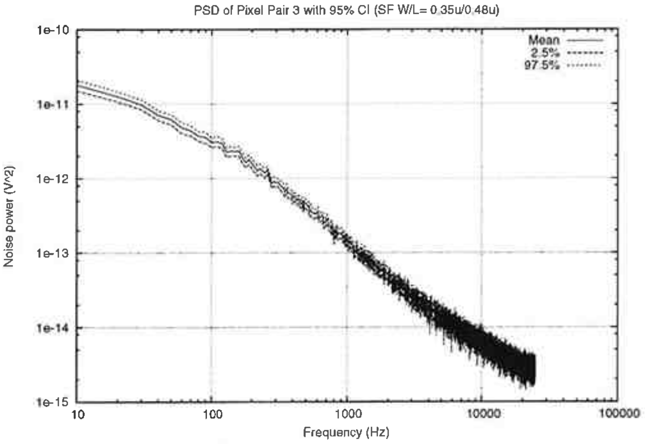

Fig. 4. PSD of Pixel Pair 3

Figure 5 shows the ensemble of single sided pixel pair PSDs. Note the wide variation in PSD between each pixel pair.

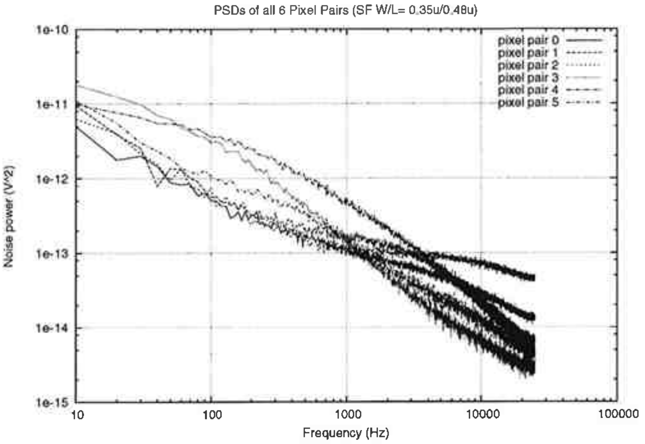

Fig. 5. PSD of all 6 Pixel Pairs

Figure 6 shows the average single sided pixel PSD with its 95% confidence interval. $R = 1024$ Bootstrap replications were used to estimate the confidence interval. Note that the 95% confidence interval in this Figure is much larger than in Figure 4.

Table I shows a summary of measured results. Note that $\overline{\rho_{AF,KF}}$ is the estimated correlation coefficient between AF and KF based on the distribution of AF and KF generated by Bootstrap re-sampling, and $\sigma^2_{1/f} = \int_{\frac{1}{T}}^{(\frac{N}{2}-1)\frac{1}{T}} \frac{\overline{KF}}{C_{ox}WLf^{\overline{AF}}}df$ is the integrated 1/f noise power for the SPICE NLEV3 model using the estimated parameters $\overline{AF}$ and $\overline{KF}$.

	2.5%	Mean	97.5%	Units
$\overline{A_{v1}A_{v2}}$		0.57		V/V
$\overline{AF}$	0.57	0.89	1.23	unit-less
$\overline{KF}$	6.4e-27	1.9e-25	1.1e-24	V^2F
$\overline{\rho_{AF,KF}}$		0.79		unit-less
$\sigma^2_{1/f}$	1.08	1.66	2.90	nV^2

TABLE I
MEASURED RESULTS

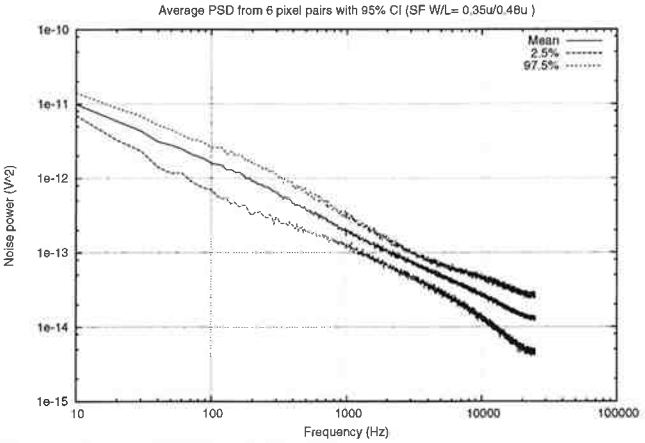

Fig. 6. Average PSD from 6 Pixel Pairs

IV. DISCUSSION

Using the DSFN measurement method almost no 60Hz, i.e. power line, interference can be observed in the final PSDs. This implies that external common mode noise was significantly attenuated. The relatively narrow confidence interval for $\overline{S_{v1,i}(f)}$, compared with $\overline{S_{v1}(f)}$, can not be explained by the difference between $P = 201$ and $Q = 6$, i.e. the number of independent samples used for averaging. The wide confidence interval for $\overline{S_{v1}(f)}$ is caused by large transistor to transistor variations in noise.

The very wide confidence intervals for $\overline{AF}$ and $\overline{KF}$ are a bit misleading due to the fact that these parameters are highly correlated. The variance in integrated 1/f noise $\sigma^2_{1/f}$ is typically a better measure of how well $\overline{AF}$ and $\overline{KF}$ estimate 1/f noise in a SPICE simulation than the variance of either $\overline{AF}$ or $\overline{KF}$. The final selection of AF and KF for SPICE simulations should be based on maximizing the integrated noise power. This will guarantee worst case noise estimates from SPICE. It should also be noted that due to the wide variation in $\overline{AF}$ and $\overline{KF}$ they should only be considered valid over the measured frequency range.

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