Analysis and improvement of ramp gain error in single-ramp single-slope ADCs for CMOS image sensors

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Abstract

The single-ramp single-slope (SRSS) analog-to-digital converter (ADC) is a promising candidate for column-parallel architectures. This paper, for the first time, quantitatively analyzes the ramp gain error in the ramp-input stage (RIS) that is the crucial component in such ADC, and verifies it using behavior model simulation. In contrast to the conventional passive RIS, an active RIS is proposed to alleviate the ramp gain error. With an active feedback loop, the ramp gain becomes a product of two reciprocal ratios of capacitance, whose departure from unity is attributed to the relative deviations rather than the absolute values in the passive RIS. The proposed concept is designed and verified by simulations in a 0.18- μm CMOS process.

Keywords: CMOS image sensor (CIS), analog-to-digital converter (ADC), column-parallel ADC, single-slope ADC, two-step ADC

1. Introduction

A CMOS image sensor (CIS) provides the significant advantages over charge coupled devices (CCDs) due to lower power consumption and easier System on Chip (SoC) integration. A column-parallel architecture has become popular for analog-to-digital converter (ADC) integration in CIS, in that it features a good trade-off among frame rate, column number, noise, and power consumption.

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Figure 1: Conventional passive ramp input stage: (a) schematic with V_{RMP} from a ramp generator and V_A to a comparator, and (b) timing sequence.

There are many types of ADCs employed for column-parallel architectures, such as single-slope (SS) ADCs[1–4], successive approximation register (SAR) ADCs [5–9], Δ - Σ ADCs[10], and cyclic ADCs [11, 12].

Among them, the SS ADC is most widely used for its good uniformity and circuit simplicity, although it requires 2^B clock cycles for B bit conversion. To speed up the conversion, the multiple-ramp SS (MRSS) ADCs [13] divides B -bit conversion into M -bit coarse and N -bit fine conversions, so as to shorten the conversion cycles to $(2^M + 2^N)$. And the final output code is obtained as

$$D = 2^N \cdot m + n, \quad (1)$$

where m and n are the latched codes in the coarse and fine ramp phases, respectively. Nevertheless, the required 2^M fine ramp generators not only occupy large area, but more seriously suffer from the mismatch of the ramp slopes and offsets, both of which limit the coarse ramp resolution M to only 2 or 3 bits with little speed improvement. For faster conversion, the single-ramp SS (SRSS) ADCs [14] memorize the coarse ramp voltage individually by using a holding capacitor (C_H) and two switches as shown in Fig. 1(a), termed passive ramp input stage (RIS) in this paper. Hence, only two ramp generators are required, and the conversion cycles become $(2^M + 2^{N+1})$ due to offset voltage correction.

However, the passive RIS suffers from parasitic capacitance during the fine ramp phase. As shown in Fig. 1, the fine ramp is transferred from R to A via B and C_H , whose top plate A is floating and thus sensitive to charge injection,

clock feedthrough, and parasitic capacitance. The former two induce an offset voltage (Δ_{OS}), which has been made signal-independent and covered by the fine ramp extension [14]. The parasitic capacitance (C_{PA} in gray in Fig. 1(a)) causes the fine ramp gain (G) deviated from unity given by

$$G = \frac{\Delta V_A}{\Delta V_R} = \frac{\Delta V_A}{\Delta V_B} = \frac{C_H}{C_H + C_{PA}} < 1, \quad (2)$$

leading to overall nonlinearity unless C_H is large enough, which makes it difficult to optimize the conversion speed for high resolutions. This problem has been
 20 neither analyzed, nor solved by far. An alternative approach proposed in [15] avoids this problem by replacing C_H with 2^M DAC (digital-to-analog converter) switches controlled by the latched m . However, two separate DAC-based ramp generators require the same resolution, and the local hardware expense in each
 25 column may also limit M .

Therefore, this paper firstly quantifies the ramp gain error effect on linearity of the SRSS ADCs, termed linearity constraint, and verifies it by behavior model simulation. In order to improve linearity, an active RIS is proposed and verified by simulation. The conclusions are drawn at the end of this paper.

30 **2. Linearity constraint**

The following analysis is based on the SRSS ADC with the fine ramp expanding by $\pm 0.5\Delta_C$ or $\pm 2^{N-1}\Delta_F$ [14], where $\Delta_C = \frac{V_{FS}}{2^M}$, $\Delta_F = LSB = \frac{V_{FS}}{2^{M+N}}$, and V_{FS} is the full scale voltage. As shown in Fig. 2, the m^{th} fine ramp under investigation consists of three parts termed negative extension, main fine ramp,
 35 and positive extension, where a horizontal line and the following vertical line are termed the code and the corresponding code width, respectively. Although only $G < 1$ arises in the passive RIS in Fig. 1(a), both $G < 1$ and $G > 1$ will be examined in that both cases do arise in the proposed RIS.

2.1. Basic properties

40 Two properties can be easily obtained in the RIS operation. Firstly, a code in the m^{th} fine ramp does not appear in the final output codes unless its step partly

Figure 2: The m^{th} 3-bit fine ramp with different fine ramp gain (G) and without offset voltage ($\Delta_{OS} = 0$).

or totally falls into $[m\Delta_C, (m+1)\Delta_C)$ which is set by the coarse ramp. Secondly, if there is no offset voltage, i.e., $\Delta_{OS} = 0$, the m^{th} positive extension always starts with $(m+1)\Delta_C$ regardless of G . This can be explained as follows. At the end of the coarse ramp phase, $(m+1)\Delta_C$ is stored on Node A as V_A . Then, Node B is switched from V_{REF} to V_{RMP} , making V_A decreased by $(2^N + 2^{N-1})G$ in LSB. In the fine ramp phase, V_A is increased by G every step and reaches $(m+1)\Delta_C$ after completing the negative extension and the main fine ramp that contain $(2^{N-1} + 2^N)$ number of steps in total.

For the convenience of discussion, the final output codes in the same (m^{th}) fine ramp are categorized into four types according to the step locations of their fine ramp codes with respect to $[m\Delta_C, (m+1)\Delta_C)$.

- 1) The totally appearing code (TAC) is defined as a final output code with its fine step completely inside $[m\Delta_C, (m+1)\Delta_C)$, and its code width always equals G .
- 2) The first appearing code (FAC) is defined as a final output code with its fine step across $m\Delta_C$, above which the step portion is its code width denoted as W_{F0} and W_F for $\Delta_{OS} = 0$, and $\Delta_{OS} \neq 0$, respectively.
- 3) The last appearing code (LAC) is defined as a final output code with its fine step across $(m+1)\Delta_C$, below which the step portion is its code width denoted as W_{L0} and W_L for $\Delta_{OS} = 0$, and $\Delta_{OS} \neq 0$, respectively.

Obviously, $W_{L0} = G$ based on the second property.

- 4) The non-linear code (NLC) is either missing code or a pair of duplicate codes, the number of which in each fine ramp is denoted as p_0 and p for $\Delta_{OS} = 0$, and $\Delta_{OS} \neq 0$, respectively. Since (1) can be rewritten as $D = 2^N \cdot m + n = 2^N \cdot (m - 1) + (n + 2^N)$, the same final output code is possibly made up of different pairs of coarse and fine codes, termed a pair of duplicate codes. When NLC is a pair of duplicate codes, its code width is the sum of both code widths, and the number of NLCs is defined positive. When NLC is missing codes, its code width is 0, and the number of NLCs is defined negative.

Based on the two properties aforementioned, the widths of FAC and LAC as well as the number of NLC will be examined in details in Section 2.2 and 2.3.

2.2. $G \neq 1$ and $\Delta_{OS} = 0$

In this case as shown in Fig. 2, $W_{L0} = G$ according to the second property, but W_{F0} varies depending on G explained as below. When $G < 1$, the m^{th} fine ramp is squeezed towards $(m + 1)\Delta_C$. Once their steps have been partly or totally squeezed above $m\Delta_C$, p_0 codes in the negative extension, that used to be invisible, appear and become duplicate with the codes in the $(m - 1)^{th}$ main fine ramp. Therefore, W_{F0} becomes

$$W_{F0} = 2^N(1 - G) - (p_0 - 1)G \quad (3)$$

in unit of Δ_F , where p_0 is an integer making $0 < W_{F0} \leq G$, and consequently satisfies

$$p_0 = \left\lceil 2^N \left(\frac{1}{G} - 1 \right) \right\rceil. \quad (4)$$

- When $G > 1$ in Fig. 2, the m^{th} fine ramp is stretched away from $(m + 1)\Delta_C$. Once their steps have been totally stretched below $m\Delta_C$, p_0 codes in the main fine ramp, that used to be visible, disappear and become missing codes. Because the number of missing codes is negative, (3) also applies to $G > 1$.

Using (3), a ramp gain range can be derived as follows to guarantee linearity, termed linearity constraint. Monotonicity can be interpreted as a lower bound

Table 1: Code widths and non-linear code numbers with offset voltage.

Case	$r = \Delta_{OS} \% G$	W_F	W_L	p
1	$G - W_{F0} < r < G$	$W_{F0} + r - G$	$W_{L0} - r$	$p_0 + 1$
2	$0 \leq r \leq G - W_{F0}$	$W_{F0} + r$	$W_{L0} - r$	p_0
3	$-W_{F0} < r < 0$	$W_{F0} + r$	$-r$	$p_0 + 1$
4	$-G < r \leq -W_{F0}$	$W_{F0} + G + r$	$-r$	p_0

of G so that only the greatest code in the negative extension can have its step partly squeezed above $m\Delta_C$, which is equivalent to $p_0 = 1$ and $W_{F0} < G$. Meanwhile, no missing codes can be interpreted as an upper bound of G so that only the smallest code in the main fine ramp can have its step partly stretched below $m\Delta_C$, which is equivalent to $p_0 = 0$ and $W_{F0} > 0$. Substituting these two conditions into (3), the linearity constraint is obtained as

$$1 - \frac{1}{2^N + 1} < G < 1 + \frac{1}{2^N - 1}. \quad (5)$$

2.3. $G \neq 1$ and $\Delta_{OS} \neq 0$

A non-zero offset voltage Δ_{OS} makes the whole fine ramp move vertically, but can be covered if

$$-2^{N-1}G < \Delta_{OS} < (2^N + 2^{N-1})G - 2^N. \quad (6)$$

80 Within this range, Δ_{OS} can be decomposed as $qG + r$ in LSB, where q and r are the quotient and remainder, respectively. r has the same sign as Δ_{OS} , and reflects the Δ_{OS} effect on non-linearity as explained as follows.

- When $r \geq 0$, the whole fine ramp is moved upwards by r . As r increases from 0 to $G - W_{F0}$, both FAC and LAC remain unchanged, and therefore, 85 $p = p_0$ with $W_F = W_{F0} + r$ and $W_L = W_{L0} - r$.
- Once $r > G - W_{F0}$, the FAC becomes a TAC in that its step is totally moved into $[m\Delta_C, (m+1)\Delta_C)$. Meanwhile, the code just below the old FAC, that used to be invisible, becomes the new FAC with $W_F = r -$

Figure 3: Block diagram of an SRSS ADC.

90 $(G - W_{F0})$. As a result, the numbers of the duplicate or missing codes are increased and decreased by 1, respectively, leading to $p = p_0 + 1$. Note that due to $p_0 < 0$ for missing codes, $p_0 + 1$ means that the number of missing codes is decreased by 1.

- When $r < 0$, the whole fine ramp is moved downwards by $|r|$. The LAC becomes a TAC in that its step is totally moved into $[m\Delta_C, (m+1)\Delta_C)$.
95 Meanwhile, the code just above the old LAC, that used to be invisible, becomes the new LAC with $W_L = |r|$. As r decreases from 0 to $(-W_{F0})$, the FAC remains unchanged with $W_F = W_{F0} - |r|$. As a result, $p = p_0 + 1$ due to appearance of one new code.
- Once $r \leq -W_{F0}$, the old FAC disappears due to its step totally moved
100 out of $[m\Delta_C, (m+1)\Delta_C)$. Meanwhile, the code just above the old FAC, that used to be a TAC, becomes the new FAC with $W_F = G - (|r| - W_{F0})$. As a result, p returns to p_0 due to disappearance of one code.

These four cases are summarized in Table 1.

2.4. Verification by behavior model

105 In order to support the theory, the SRSS ADC is modeled in the behavior level with adjustable G and Δ_{OS} . This behavior model describes the static

(a)

(b)

(c)

(d)

(e)

(f)

Figure 4: Typical static characteristics simulated by behavior model for different G and Δ_{os} .

Table 2: Theoretical differential non-linearity.

p	Differential Nonlinearity (DNL)			
	TAC	NLC	FAC	LAC
$p < 0$	$G - 1$	-1	$W_F - 1$	$W_L - 1$
$p = 0$	$G - 1$	—	$W_F - 1$	$W_L - 1$
$p = 1$	$G - 1$	$W_F + W_L - 1$		
p	TAC	NLC		
		TAC+TAC	FAC+TAC	LAC+TAC
$p = 2$	$G - 1$	—	$W_F + G - 1$	$W_L + G - 1$
$p > 2$	$G - 1$	$2G - 1$	$W_F + G - 1$	$W_L + G - 1$

characteristic of the SRSS ADC, and is implemented in MATLAB based on the circuit blocks as shown in Fig. 3. The counter generates a digital code increasing from 0 to 2^M and then from 0 to 2^{N+1} for the coarse and fine ramp phases, respectively, which is multiplied with Δ_C and Δ_F , respectively, and converted to the ramp voltage (V_{RMP}) by a DAC. The RIS transfers V_{RMP} to one of the comparator input by taking into account the non-idealities of gain error and offset voltage. By comparing the input voltage (V_{IN}) and the non-ideal ramp voltage, the comparator decides whether the digital code should be stored into the memory. In the end of one conversion cycle, the final code is obtained by two codes of m and n stored in the memory according to (1).

With $B = 12$ bits and $N = 6$ bits, the static characteristic of the behavior SRSS ADC is simulated as the function of Δ_{OS} sweeping from -1LSB to 1LSB, while G is chosen as 0.98, 0.99, 1.01, and 1.02, so as to represent four ranges separated by $\left(1 - \frac{1}{2^{N+1}}\right)$, 1, and $\left(1 + \frac{1}{2^{N-1}}\right)$ according to (5). Fig. 4 shows the typical simulation results, which intuitively exhibits various non-linearity phenomena in the SRSS ADC. From these simulation results, the specifications of non-linearity are extracted, including differential non-linearity (DNL) and the number of NLCs (p) as illustrated as symbols in Fig. 5.

With a view to comparison, theoretical values for p and DNL are needed. p

(a)

(b)

(c)

(d)

(e)

(f)

(g)

(h)

Figure 5: Comparison of analytical theory (lines) and behavior model (symbols) regarding DNL and number of abnormal codes as the function of offset voltage with ramp gains of (a)(b) 0.98, (c)(d) 0.99, (e)(f) 1.01, and (g)(h) 1.02.

has already listed in Table 1, where p_0 can be calculated by (4). Meanwhile, as listed in Table 2, the theoretical DNLs are derived from the theoretical code widths in Table 1, where W_{F0} can be calculated by (3) and (4). When $p < 0$ and $p > 2$, DNLs have four kinds of different values, corresponding to four kinds of codes. When p is in between, the kinds of DNL values are less than four stemming from changes of NLC, FAC, and LAC as follows. When $p = 0$, there is no NLC so that the kinds of DNL values are decreased to 3. When $p = 1$, FAC and LAC in the neighboring fine ramps become a pair of duplicate codes, and merge into one kind of NLC, so that the kinds of DNL values are further decreased to 2. When $p = 2$, FAC and LAC are duplicate with the TACs in the preceding and succeeding fine ramps, respectively, and become two kinds of NLCs, so that the kinds of DNL values are increased to 3.

Fig. 5 compares the theoretical p and DNL (lines) with the behavior simulation results (symbols). For simplicity, the theoretical DNLs in Table. 2 for $p < 0$ and $p > 2$ are plotted for $G > 1$ and $G < 1$, respectively, while DNLs for $p = 0, 1, 2$ are discussed individually as below.

- The extracted DNLs and p 's are periodic functions of Δ_{OS} in that both of them are functions of r in theory while r is a periodic function of Δ_{OS} .
- When $p > 2$ or $p < 0$, the theoretical values of DNLs and p 's agree with the extracted counterparts as demonstrated in Figs. 5(a)(g) and Figs. 5(b)(h), respectively, for $p = 3$ and $p = -1$. Figs. 5(a)(g) also show that the extracted DNLs have four kinds of values as expected in theory, which is manifested in Figs. 4 (a)(f).
- When $p = 2$, the theoretical values of DNLs and p 's agree with the extracted counterparts as demonstrated in Figs. 5(a)(c) and Figs. 5(b)(d), respectively. Figs. 5(a)(c) also show that the extracted DNLs have three kinds of values as expected in theory which is manifested in Fig. 4(b).
- When $p = 1$, Figs. 5(c)(e) show that the extracted DNLs have two kinds of values as expected in theory, which is manifested in Figs. 4 (c)(d).

155 Figs. 5(c)(e) also show that the DNLs of TAC are matched between theory
and extraction, while the extracted DNLs of NLC are independent of Δ_{OS} ,
which is explained as follows. For $G = 0.99$ in Fig. 5(c) as r increases from
0, FAC and LAC of the adjacent fine ramps become a pair of duplicate
codes and combined as one NLC. Their widths are expressed in Case 2 in
160 Table 1, and therefore, this NLC width is $W_{N1} = W_F + W_L = (W_{F0} +$
 $r) + (W_{L0} - r) = W_{F0} + G$. Although the same combination also arises for
 $G = 1.01$ in Fig. 5(e) as r decreases from 0, the difference is that due to
 $r < 0$, W_F and W_L are expressed in Case 3 in Table 1, and therefore, this
NLC width is $W_{N2} = W_F + W_L = (W_{F0} + r) + (-r) = W_{F0}$. However,
165 when extracting p from the behavior model simulation, the algorithm finds
a pair of duplicate codes if its code width is greater than G . Therefore,
due to $W_{N1} > G$, the former NLC is regarded as a pair of duplicate codes
so that the extracted p 's equal 1 in Fig. 5(d), while due to $W_{N2} \leq G$,
the latter NLC is not regarded as a pair of duplicate codes so that the
170 extracted p 's equal 0 rather than 1 in theory in Fig. 5(f).

- When $p = 0$, the theoretical values of DNLs and p 's agree with the ex-
tracted counterparts as demonstrated in Figs. 5(e)(g) and Figs. 5(f)(h),
respectively. Figs. 5(e)(g) also show that the extracted DNLs have three
kinds of values as expected in theory, which is manifested in Fig. 4(e).

175 In summary, the proposed theory is verified by giving perfect explanations
for the simulation results of the SRSS ADC behavior model.

3. Active RIS

The proposed theory indicates that for the passive RIS in Fig. 1(a), the
linearity constraint imposes a lower bound on C_H as

$$C_H > 2^N C_{PA}, \quad (7)$$

by substituting (2) into (5). Therefore, for a high-resolution SRSS ADC using
the passive RIS, B can hardly be equally divided into M and N , involving a

Figure 6: Proposed active ramp input stage: (a) schematic with V_{RMP} from a ramp generator and V_A to a comparator, and (b) timing sequence.

Figure 7: Equivalent circuits for the active RIS in (a) sampling, (b) holding, and (c) follower modes.

tradeoff between speed, area and power consumption. In order to overcome this drawback, an active RIS is proposed.

3.1. Operation principle

Fig. 6(a) shows the circuit for the proposed active RIS, which is essentially a flip-around sample-and-hold circuit built by an operational transconductance amplifier (OTA). Controlled by the timing shown in Fig. 6(b), the active RIS sequentially operates in three modes as shown in Fig. 7.

3.1.1. Sampling mode

At the beginning of a coarse ramp phase, S_2 and S_5 are turned off, while the others are turned on. The active RIS goes into the sampling mode as shown in

Fig. 7(a). V_{RMP} from the coarse ramp generator is sampled by the OTA, and simultaneously shorted with A to compare with the input signal (V_{IN}) by the comparator (not shown). Therefore, the node voltages are

$$V_X = V_O = \frac{A_{OL}}{1 + A_{OL}} (V_{REF} + V_{OS}), \quad (8)$$

and

$$V_Y = V_A = V_{RMP}, \quad (9)$$

where A_{OL} and V_{OS} are the open-loop gain and the offset voltage of the OTA, respectively.

190 3.1.2. Holding mode

Once $V_{RMP} > V_{IN}$, the active RIS goes into the holding mode as shown in Fig. 7(b), in the same way as the flip-around operation [16] as follows. S_4 turns off first to avoid the signal-dependent charge injection, and then S_1 and S_5 change their states immediately. Thus, the instantaneous coarse ramp voltage (V_{HLD}) is memorized with the node voltages expressed as

$$V_A = V_Y = V_O = \frac{A_{OL}}{1 + A_{OL}\beta_X} \left(\beta_X V_{HLD} + \frac{V_{REF} + V_{OS}}{1 + A_{OL}} \right), \quad (10)$$

where

$$\beta_X = \frac{C_{HX}}{C_{HX} + C_{PX}}. \quad (11)$$

3.1.3. Follower mode

When the fine ramp phase begins, S_6 turns off first, which injects signal-independent charges to A due to $V_A = V_Y$. Then, S_2 and S_3 change their states so that Node B is switched from V_{REF} to V_{RMP} , and the active RIS goes into the follower mode as shown in Fig. 7(c). V_{RMP} , that is now from the fine ramp generator, changes the voltage on B , and then propagates through O to A , whose voltage changes are denoted as ΔV_B , ΔV_O and ΔV_A , respectively. Hence, its ramp gain becomes

$$G = \frac{\Delta V_A}{\Delta V_B} = \frac{\Delta V_A}{\Delta V_O} \frac{\Delta V_O}{\Delta V_B} = \frac{\beta_A}{\beta_X} \frac{1}{1 + \frac{1}{A_{OL}\beta_X}} \quad (12)$$

in that

$$\frac{\Delta V_O}{\Delta V_B} = \frac{A_{OL}}{1 + A_{OL}\beta_X}, \quad (13)$$

and

$$\frac{\Delta V_A}{\Delta V_O} = \beta_A = \frac{C_{HA}}{C_{HA} + C_{PA}}. \quad (14)$$

Let C_{HA} and C_{HX} have a nominal value of C_H , while C_{PA} and C_{PX} have a nominal value of C_P . In order to achieve this, each of C_{PA} and C_{PX} consists of not only the parasitic capacitance, but also a manually added capacitor that plays a leading role. Assume that $C_{HA} = C_H$ and $C_{PA} = C_P$, while C_{HX} and C_{PX} deviate from their nominal values by ΔC_H and ΔC_P , respectively, i.e., $C_{HX} = C_H + \Delta C_H$ and $C_{PX} = C_P + \Delta C_P$. As $A_{OL} \gg 1$, $\Delta C_H \ll C_H$, and $\Delta C_P \ll C_P$, (12) can be rewritten as

$$G \approx \frac{\beta_A}{\beta_X} = 1 + \frac{C_P}{C_H + C_P} \left(\frac{\Delta C_P}{C_P} - \frac{\Delta C_H}{C_H} \right) \quad (15)$$

which depends on the relative deviation denoted as $\frac{\Delta C}{C} = \frac{\Delta C_P}{C_P} - \frac{\Delta C_H}{C_H}$.

3.2. Design constraints

In order to estimate C_H and C_P , design constraints are discussed. As shown in Fig. 7, the active RIS responds to V_{RMP} in two different feedback configurations with distinct feedback factors (β) and capacitive loads (C_L) as expressed by

$$\beta = 1 \text{ and } C_L = C_{PX} + C_{HX} \quad (16)$$

in the sampling mode, while

$$\beta = \beta_X \text{ and } C_L = \frac{C_{PX}C_{HX}}{C_{PX} + C_{HX}} + \frac{C_{PA}C_{HA}}{C_{PA} + C_{HA}} \quad (17)$$

in the follower mode.

3.2.1. Linearity constraint

Substituting (15) into (5), the linearity constraint for the active RIS becomes

$$C_H > \left[(2^N + 1) \left| \frac{\Delta C_P}{C_P} - \frac{\Delta C_H}{C_H} \right| - 1 \right] C_P \quad (18)$$

Note that in the active RIS, each of C_{PX} and C_{PA} consists of a manually
 205 implemented capacitor besides the parasitic one. Hence, ΔC_P accounts for
 both mismatch, and the difference in the voltage-dependent input capacitance
 between the OTA and the comparator.

3.2.2. Thermal noise constraint

Regarding the thermal noise [17], the sampled noise power goes directly to
 O by the flip-around operation, and after combined with the OTA noise power
 referred to O , is delivered to A as an output $\overline{V_{out}^2}$ by a factor of β_A^2 . For $\overline{v_{out}^2}$ to
 be less than half of the quantization noise power, the thermal noise constraint
 is given by

$$\overline{v_{out}^2} = \left(\frac{kT}{C_{HX}} + \frac{4kT}{3\beta_X C_L} \right) \beta_A^2 < \frac{\Delta_F^2}{24} \quad (19)$$

where k is the Boltzmann constant, T is the absolute temperature in degrees
 210 Kelvin, and C_L is given by (17).

3.2.3. Power constraint

For a B-bit ADC, the accuracy is determined by the static and dynamic
 performance of the OTA. The static deviation must be less than $\Delta_F/2$ for the
 OTA output at O , which imposes a lower limit on the open-loop gain (A_{OL}) of
 the OTA given by [18]

$$A_{OL} > \frac{2^{B+1}}{\beta}. \quad (20)$$

The dynamic accuracy requires that the exponential settling error at O must
 be less than $\Delta_F/2$ within a settling time given by a proportion (n) of the clock
 period ($1/f_{clk}$), which imposes a lower limit on the gain-bandwidth product
 (GBW) of the OTA given by [18]

$$GBW > \frac{(B+1)(\ln 2)}{2\pi n\beta} \cdot f_{clk}. \quad (21)$$

Therefore, the power constraint is written as

$$I_i > \left[(B+1)(\ln 2) \frac{f_{clk}}{n} \frac{C_L}{\beta} \right] \left(\frac{V_{od}}{2} \right) \quad (22)$$

Figure 8: Design space with the selected design point

in that $GBW = \frac{g_m}{2\pi C_L}$ and $g_m = \frac{2I_i}{V_{od}}$ [18], where g_m is the transconductance of the OTA, I_i is the current through the OTA input pair, and V_{od} is the overdrive voltage. Equ. (22) should be applied to both feedback configurations in (16) and (17).

3.3. Design space

The three constraints given by (18), (19), and (22) are embodied by a design space as shown in Fig. 8, after substituting the parameter values listed in Table 3. In Table 3, V_{od} is simplified for high gain design [18], $|\frac{\Delta C}{C}|$ is evaluated by simulation, and T is chosen in the cryogenic temperature range for long-wavelength infrared (LWIR) detectors [19]. Therefore, within the design space, $(C_P, C_H) = (450 \text{ fF}, 200 \text{ fF})$ is chosen as a design point.

Table 3: Design parameter values for the design space

B	N	V_{FS}	$ \frac{\Delta C}{C} $	T	I_i	f_{clk}	n	V_{od}
(bit)	(bit)	(V)		(K)	(μA)	(MHz)		(V)
12	6	1	2%	80	100	50	2/3	0.2

3.4. Simulation and discussion

With the design point, the active RIS shown in Fig. 6(a) is designed in a $0.18\text{-}\mu\text{m}$ CMOS process with its layout extracted for simulation. Since the other

Figure 9: Monte Carlo simulation results of $G - 1$ for active RIS (a) Histogram and fitted normal distribution of 300 samples at $V_{HLD} = 0.9V$ (b) $3\text{-}\sigma$ error bars as a function of V_{HLD}

blocks of the SRSS ADC in Fig. 3 are also necessary to perform simulation, the ramp generator, timing control, and comparator are implemented by verilog, verilog, and circuit, respectively. The comparator employs a preamplifier with a tail current to weaken the dependence of its input capacitance on V_A [20].

230 The simulation is carried out at 80K, although it is beyond the valid range of the device models provided by the foundry. Due to the absence of accurate models [21], enough margins have to be left for the increases of channel conductance and threshold voltages [22].

As the mismatch plays an important role in G , Monte Carlo mismatch
 235 simulation is performed for the active RIS. Afterwards, the ramp gain error $\Delta G = G - 1$ is extracted and statistically analyzed. Fig. 9(a) shows the histogram of 300 samples at $V_{HLD} = 0.9V$, to which a normal distribution is fitted with the mean value (μ) and standard deviation (σ) obtained. In this way, the statistical properties of ΔG for different V_{HLD} 's are obtained and depicted as
 240 $3\text{-}\sigma$ error bars as shown in Fig. 9(b), which indicates that the mean value weakly depends on V_{HLD} , and ΔG satisfies (5) even with $3\text{-}\sigma$ deviations.

The 12-bit SRSS ADC with the active RIS is simulated with its output codes

Figure 10: Simulation result of 512-point FFT spectrum for a 35.82-kHz input signal at 0.25 MS/s.

analyzed by FFT as shown in Fig. 10, where the sampling rate approximately equals $f_{clk}/(2^M + 2^{N+1})$.

245 The total capacitance is $1.3pF$ in this design. While for the passive RIS in Fig. 1, if $C_{PA} = 50fF$, $C_H > 3.2pF$ according to (7), to which over 20% margin should be added for process variations. Taking $C_H = 4pF$ in the passive RIS for example, the capacitor area can be reduced by $1350\mu m^2$ using $2fF/\mu m^2$ Metal-Insulator-Metal (MIM) capacitors, which is partly cancelled out by the OTA
250 area about $480\mu m^2$. The OTA power consumption is simulated about $180\mu W$, which seems to be a penalty from a single ADC point of view. However, in a column-parallel architecture, it is expected to be cancelled out by the power reduction of the global buffers that drive V_{RMP} and V_{REF} due to the reduced capacitive loads.

255 4. Conclusions

This paper proposed a theory quantifying the ramp gain error effect of RIS in SRSS ADCs, which is strongly supported by the behavior model simulation. The active RIS is proposed for SRSS ADCs, and designed under the guidance of the design space. Circuit simulation results verify its enhancement on linearity.
260 It is a potential solution to the parasitic capacitance problem of RIS.

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