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Analysis of the Sampling Noise Cancellation Technique in a Track-and-Hold Amplifier

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Abstract—The sampling noise cancellation (SNC) technique has been applied to different discrete-time circuits in recent literature, but the cost of this technique has not been well studied. This paper gives a detailed analysis of the SNC approach in a track-and-hold amplifier (THA). Closed-form expressions are derived for each noise source in the THA during the sampling phase and are verified by simulations. The analysis shows that while the sampling noise can be cancelled, the noise from the extra amplifier and switches can become dominant and even outweigh the benefits from SNC. Therefore, the SNC technique should be used with attention to the trade-offs, as described in this paper.

Index Terms—track-and-hold amplifier, kT/C noise, sampling noise cancellation

I. INTRODUCTION

The sampling process is ubiquitous in almost every discrete-time circuit, such as switched-capacitor amplifiers and filters, analog-to-digital converters (ADC), etc. The well-known kT/C noise associated with every sampling action has been regarded as a fundamental limit in such circuits, as a result of which every 6 dB improvement in signal-to-noise ratio (SNR) requires a quadrupled sampling capacitor, which leads to higher power consumption and larger chip area. Recently, a sampling noise cancellation (SNC) technique has been proposed and targeted to break this kT/C noise limit [1]. It has been demonstrated in closed-loop sampling circuits [1]-[3] as well as open-loop track-and-hold circuits [4]-[6]. Despite the popularity of the SNC technique, its potential cost has not been examined in detail. This paper takes the track-and-hold amplifier (THA) in which the SNC was initially proposed [1] as an example to investigate the trade-offs behind this technique.

The underlying principle of SNC is to treat the sampling noise as a time-varying offset for each sampling period. From this point of view, the methodology of auto-zeroing can be applied to cancel the kT/C noise. Namely, the sampled noise voltage is first stored on a capacitor and then subtracted from the output. However, as auto-zeroing results in a higher noise floor, one can also expect that the SNC achieves the noise cancellation function at the expense of something else. This paper aims to answer this question by a thorough analysis of the SNC operation and it is organized as follows: Section II reviews the conventional THA and its sampling noise.

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Fig. 1. Conventional THA (a) schematic (b) timing diagram.

Section III presents the THA with SNC and provides a detailed analysis for each noise source. Section IV shows the simulation results and explains the trade-offs in the SNC technique. Finally, Section V draws conclusions.

II. CONVENTIONAL THA

Fig. 1 shows the conventional THA and its timing diagram. The THA is controlled by three clocks and operates in two phases: sampling phase (φ_1 and φ_2) and amplification phase (φ_3). The falling edge of φ_2 is delayed by T_{DEL} compared to that of φ_1 in order to avoid signal-dependent charge injection from the sampling switches. The total noise of the THA is the summation of sampling noise from φ_1 and φ_2 phases and amplification noise from the φ_3 phase. This paper only focuses on the sampling noise, while the analysis of the amplification noise can be found in [7], and methods to cancel the thermal noise during the amplification phase are discussed in [8] and [9].

The sampling noise calculation of a conventional THA starts with the φ_1 switch noise. At the end of φ_1 , a noise voltage V_{n1} is sampled on the right plate of C_S , as shown in Fig. 2(a). The total charge $Q_{\varphi 1}$ stored on C_S and C_F is:

$$Q_{\varphi 1} = (V_{n1} - V_{in})C_S + V_{n1}C_F. \quad (1)$$

During the φ_2 phase, the switches controlled by the φ_2 clock also generate noise. Assume that they generate instantaneous noise voltages V_{n2L} and V_{n2R} at the end of φ_2 , as shown in Fig. 2(b). This gives rise to a voltage change ΔV_{nx} at the amplifier input node:

$$\Delta V_{nx} = (1 - \beta)V_{n2L} + \beta V_{n2R} \quad (2)$$

where β is $C_F/(C_S + C_F)$. However, since $Q_{\varphi 1}$ cannot change in this phase, no extra noise is sampled during φ_2 . In phase

Fig. 2. Conventional THA noise model in (a) φ_1 phase and (b) φ_2 phase.

φ_3 , the loop is closed, and the amplifier input is forced to zero. Thus, the total charge Q_{φ_3} on C_S and C_F is forced to:

$$Q_{\varphi_3} = -V_{out}C_F. \quad (3)$$

Due to charge conservation ($Q_{\varphi_3} = Q_{\varphi_1}$), the amplifier output can be expressed as:

$$V_{out} = V_{in} \frac{C_S}{C_F} - V_{n1} \frac{C_S + C_F}{C_F}. \quad (4)$$

It contains the amplified input signal as well as sampled noise. Since V_{n1} has a power of $kT/(C_S + C_F)$, where k is the Boltzmann constant and T is the absolute temperature, the total output-referred sampling noise power $\overline{V_{n,out}^2}$ is:

$$\overline{V_{n,out}^2} = \frac{kT(C_S + C_F)}{C_F^2}. \quad (5)$$

As expected, the sampling noise is expressed in a kT/C format. This result is the same as the equation in [7] where the equipartition theorem is used for the noise calculation. In summary, the sampling noise in a conventional THA comes solely from the φ_1 switch, and it is not affected by the φ_2 switches and thus not dependent on the T_{DEL} duration.

III. THA WITH SAMPLING NOISE CANCELLATION

A. SNC Principle

Fig. 3 shows the THA structure with the SNC technique [1]. Compared to the conventional THA, a two-stage amplifier is used and an extra capacitor C_2 is inserted between the two amplifier stages. The first-stage amplifier A_1 has high bandwidth and low gain, while the second-stage amplifier A_2 has low bandwidth and high gain. The timing control for this THA is the same as that for the conventional case in Fig. 1(b).

To analyse the sampling noise, we can follow the same steps as in Section II. At the end of φ_1 , the noise voltage V_{n1} is sampled on C_S and C_F , and the total charge Q_{φ_1} can still be expressed as (1). During φ_2 , the amplified V_{n1} is stored on C_2 , as shown in Fig. 4(a). When the amplifier is configured closed loop in φ_3 , the right plate of C_2 will be forced to approximately zero thanks to the high gain from A_2 . Since C_2 is floating in φ_3 , the A_1 input will remain at V_{n1} instead of settling to zero, as illustrated in Fig. 4(b). Therefore, the total charge in φ_3 is:

$$Q_{\varphi_3} = V_{n1}C_S + (V_{n1} - V_{out})C_F, \quad (6)$$

Fig. 3. THA with SNC technique.

and by combining this with (1) and noting that $Q_{\varphi_3} = Q_{\varphi_1}$, the final output becomes:

$$V_{out} = V_{in} \frac{C_S}{C_F}. \quad (7)$$

The sampling noise term V_{n1} is not present in (7), and therefore, the noise cancellation function is realized if all the other elements are ideal. However, if amplifier A_1 has a finite bandwidth, its output may not settle to $A_1 V_{n1}$ in the given time T_{DEL} [1], resulting in incomplete noise cancellation. The remaining sampling noise can be written as [1]:

$$\overline{V_{n,out}^2} = \frac{kT(C_S + C_F)}{C_F^2} \exp\left(\frac{-2T_{DEL}}{R_{out1}C_2}\right) \quad (8)$$

where R_{out1} denotes the output impedance of the amplifier A_1 . However, the above calculation only holds true if the noise from the φ_2 phase can be ignored. To guarantee an effective SNC, the other potential noise contributors (switches S_2 , S_{2L} , S_{2R} and amplifier A_1) need to be investigated carefully. These contributors in the THA have not been studied before, and thus this paper analyses these additional noise sources and verifies them by means of simulations.

B. Noise from Switch S_2

The noise generated from switch S_2 is shown in Fig. 4(c). At the end of φ_2 , a noise voltage V_{n2} is sampled on the right plate of C_2 . This makes the voltage kept on C_2 no longer $A_1 V_{n1}$ but $A_1 V_{n1} - V_{n2}$. By repeating the noise calculation steps from the above sections, the S_2 -induced output-referred noise can be obtained:

$$\overline{V_{n,out}^2} = \frac{kT}{C_2} \cdot \frac{(C_S + C_F)^2}{A_1^2 C_F^2}. \quad (9)$$

C. Noise from Switches S_{2L} and S_{2R}

To simplify the calculation, it is assumed that the ratio of the on-resistance of S_{2L} and S_{2R} satisfies $R_{on2L}/R_{on2R} = C_F/C_S$, as shown in Fig. 4(d). Under this condition, the transfer function from each noise source to the amplifier input is a frequency-independent constant. This also indicates that the bandwidth of the noise seen at V_{nx} is infinity. This noise will be filtered by the amplifier A_1 and leads to an amplifier output of $A_1 V'_{nx}$, where V'_{nx} is a band-limited signal. Like V_{n1} noise, the amplified V_{nx} noise can also settle incompletely in the given T_{DEL} duration. Following the above steps, the

Fig. 4. THA with SNC technique in (a) sampling phase (φ_1 and φ_2) and (b) amplification phase (φ_3). Noise analysis of (c) switch S_2 (d) switches S_{2L} and S_{2R} (e) amplifier A_1 . (f) Differential amplifier model.

output-referred noise caused by S_{2L} and S_{2R} can be shown to be:

$$\overline{V_{n,out}^2} = \frac{kT(R_{on2L}C_S^2 + R_{on2R}C_F^2)}{R_{out1}C_2C_F^2} \left[1 - \exp\left(\frac{-T_{DEL}}{R_{out1}C_2}\right) \right]^2. \quad (10)$$

D. Noise from Amplifier A_1

The noise from amplifier A_1 can be modeled as a noise voltage power spectral density $4kT\gamma/G_{m1}$, as shown in Fig. 4(e). This is filtered thanks to the finite bandwidth of A_1 , resulting in a final output-referred noise of:

$$\overline{V_{n,out}^2} = \frac{kT\gamma}{G_{m1}R_{out1}C_2} \cdot \frac{(C_S + C_F)^2}{C_F^2}. \quad (11)$$

where G_{m1} is the transconductance of the amplifier A_1 . Note that in this paper we only focus on noise contributions that are sampled in the network. Amplifiers A_1 and A_2 also add continuous-time noise during phase φ_3 , but this is not included in the analysis, as they don't affect the actual sample.

IV. SIMULATION RESULTS

A. Simulation Setup and Results

To verify the above analysis, testbenches for both the conventional THA and a THA with SNC are created. The same two-stage amplifiers are used in these two testbenches. Each stage of the amplifier is built with the ideal components shown in Fig. 4(f). A differential configuration of the entire circuit is simulated, which doubles the output noise power. Each switch is modelled as an ideal switch in series with a resistor with $4kTR$ noise. The transient noise simulation is performed using Cadence Spectre. The sampling period is 1 μ s, and the maximum noise frequency is set to be 5 GHz. The THA input is a DC signal of 10 mV. Other key parameters used in the simulation are summarized in Table I.

Fig. 5 shows the simulation results from the conventional THA without and with φ_2 switches noise. The simulated

TABLE I
PARAMETER VALUES IN THE SIMULATION

Parameter	C_S	C_F	C_L	C_2	γ	T
Value	4 pF	1 pF	1 pF	1 pF	1	300 K
Parameter	G_{m1}	R_{out1}	C_{out1}	G_{m2}	R_{out2}	C_{out2}
Value	10 mS	1 k Ω	20 fF	1 mS	100 k Ω	20 fF
Switches	$R_{on1} = R_{on2} = R_{on2L} = 10$ k Ω , $R_{on2R} = 40$ k Ω .					

output noise matches the calculated value, and the noise is independent on T_{DEL} as expected. Fig. 6 shows the simulated results from the THA with SNC. The individual noise sources are first simulated separately, as shown from Fig. 6(a) to Fig. 6(d). When one noise source is enabled, the other noise sources are disabled (e.g. by shorting the resistor which is in series with the ideal switch). It can be observed from Fig. 6(a) and Fig. 6(c) that the noise settles faster than what (8) and (10) predict. This is because we assumed a step response with zero initial condition when deriving (8) and (10). In practice, the amplifier A_1 output also tracks the switch noise during φ_1 , so its settling does not start from the zero initial condition, and this shortens the time needed for complete settling. Fig. 6(e) shows the result when all the noise sources are included. The total noise is smaller than the addition of the individual noise values. This is because the switch on-resistance used in the simulation is relatively large, which reduces the bandwidth of some of the noise sources.

Fig. 5. Simulated output noise of the conventional THA.

Fig. 6. Simulated output noise of the THA with SNC with only (a) S_1 noise (b) S_2 noise (c) S_{2L} and S_{2R} noise (d) amplifier A_1 noise and (e) all noise sources.

Fig. 7. (a) Ideal sampling and its noise model. (b) Bottom plate sampling and its noise model.

Fortunately, this reduction in the noise bandwidth can be easily incorporated into the existing equations by multiplying (9) by $R_{on2}/(R_{on2} + R_{out1})$ and replacing R_{out1} in (10) and (11) by $(R_{on2} + R_{out1})$, respectively. The remaining difference between the simulation and calculation results in Fig. 6 is mainly due to the finite maximum noise frequency specified in the transient noise simulation (set to 5 GHz). This makes the simulated noise value in general lower than the calculated value. In summary, the theoretical calculations can be used as an accurate indication for the actual circuit noise.

Fig. 8. Simulated output noise of the THA with SNC by (a) sweeping C_2 ($R_{on1} = R_{on2} = R_{on2L} = 10$ k Ω , $R_{on2R} = 40$ k Ω) (b) sweeping R_{on2} ($C_2 = 5$ pF, $R_{on1} = R_{on2L} = 10$ k Ω , $R_{on2R} = 40$ k Ω) (c) sweeping R_{on2L} ($C_2 = 1$ pF, $R_{on1} = R_{on2} = 10$ k Ω , $R_{on2R} = 4R_{on2L}$).

B. Discussion

As shown in Fig. 6(a), the SNC technique can indeed cancel the original noise from the sampling switch S_1 . However, the total sampled noise of the THA with SNC as shown in Fig. 6(e) is even higher than that of the conventional THA in this particular simulation example. Putting aside the intricate equations, we can interpret this behaviour intuitively from two aspects.

First, the actual sampling moment is different in these two scenarios. In the conventional THA case, the actual sampling happens at t_1 in Fig. 1(b). This means that all the noise injected at moment t_1 will be maintained in the system while the noise at moment t_2 is not critical. However, the SNC moves the actual sampling moment to t_2 [4]. Although it could cancel the sampling noise at t_1 , more noise sources are present at t_2 due to the extra amplification and sampling process on C_2 . This approach thus transfers the noise origin from one component to another one, instead of removing it.

Second, the cancelled noise is not actually the classic kT/C noise that is shown in Fig. 7(a). Due to the bottom plate sampling, switches are needed at both the top and bottom plates of the capacitor in Fig. 7(b). In this case, only the voltage across the capacitor (V_C) has the noise power of kT/C . The plate voltages V_{out_top} and V_{out_btm} are correlated and have a much larger noise bandwidth. The SNC technique only cancels the noise on V_{out_btm} instead of cancelling the real kT/C noise across the sampling capacitor. The analysis of SNC in prior art [1][4] assumes a zero R_{on_top} and ignores its noise contribution.

To make the SNC technique effective, we need to make sure that the sum of the noise from (8) to (11) is smaller than (5). For simplicity, considering the noise from S_{2L} and S_{2R}

is dominant in Fig. 6, we can set the noise in (10) less than that in (5) as an example. Note that R_{out1} in (10) needs to be replaced by $(R_{on2} + R_{out1})$ as explained in Section IVA. This yields:

$$(1 + \lambda)(R_{on2} + R_{out1})C_2 > (R_{on2L} + \lambda^2 R_{on2R})C_S. \quad (12)$$

where $\lambda = C_F/C_S$ and T_{DEL} is assumed long enough. Several design choices can be made to satisfy (12), as shown in Fig. 8. First, C_2 can be made larger at the expense of higher A_1 power. But if C_2 is designed to be even larger than C_S , this will be against our initial intention to reduce the total sampling capacitance. Second, the R_{on2} and R_{out1} can be increased to reduce the noise bandwidth, but this may lead to long T_{DEL} and is thus suboptimal. Third, R_{on2L} and R_{on2R} should be sufficiently small. However, small R_{on2L} also implies that a low output impedance from the driving stage in front of the THA (not shown in the figures) is needed.

For a given bandwidth requirement, the sampling capacitor C_S size can be reduced thanks to SNC, and this can relax the input driver specification from the speed perspective. But the on-resistance of S_{2L} and the driver output impedance still need to be kept low due to the noise restrictions in (12). This can instead make the driver consume more power due to the low noise spectral density requirement. Therefore, the input driver needs to be designed carefully with the strict noise budget as well as the speed specification.

V. CONCLUSION

This paper examines the principle of the SNC technique in a THA thoroughly. All the noise sources in the sampling phase are included in the analysis, and the theoretical calculations are verified by simulations. The trade-offs between switch on-resistance, amplifier power, driver strength, total capacitance, and the final noise power are discussed. Overall, the SNC technique can indeed reduce sampling noise, but it comes at the cost of increased power or area inside the THA, or increased power in the driving stage in front of the THA.

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