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COVER This issue contains a Special Section on Signals and Symbols. The cover was inspired by the blocks used to demonstrate simulated objects in paper: Recognizing Object Function through Reasoning About Partial Shape Descriptions and Dynamic Physical Properties by Stark, *et al.*

NOVEMBER

Circuit Techniques for Reducing the Effects of Op-Amp Imperfections: Autozeroing, Correlated Double Sampling, and Chopper Stabilization

CHRISTIAN C. ENZ, MEMBER, IEEE, AND GABOR C. TEMES, FELLOW, IEEE

Invited Paper

In linear IC's fabricated in a low-voltage CMOS technology, the reduction of the dynamic range due to the dc offset and low-frequency noise of the amplifiers becomes increasingly significant. Also, the achievable amplifier gain is often quite low in such a technology, since cascoding may not be a practical circuit option due to the resulting reduction of the output signal swing. In this paper, some old and some new circuit techniques will be described for the compensation of the amplifier most important nonideal effects including the noise (mainly thermal and $1/f$ noise), the input-referred dc offset voltage, as well as the finite gain resulting in a nonideal virtual ground at the input.

I. INTRODUCTION[1]

In linear active circuits, the active element most often used is the operational amplifier (op-amp), whose main function in the circuit is to create a virtual ground, i.e., a node with a zero (or constant) voltage at its input terminal without sinking any current. Using op-amps with MOS input transistors, the op-amp input current at low frequencies can indeed be made extremely small; however, the input voltage of a practical op-amp is usually significantly large (typically of the order of 1–10 mV), since it is affected by several nonideal effects. These include noise (most importantly, $1/f$ and thermal noise), the input-referred dc offset voltage, as well as the signal voltage needed to generate the desired output voltage of the op-amp. Normally, the thermal noise occupies a wide frequency band, while the $1/f$ noise, offset and input signal are narrowband low-frequency signals.

The purpose of the circuit techniques discussed in this paper is to reduce the effects of the narrow-band noise sources at the virtual ground of an op-amp stage. By reducing the low-frequency noise and offset at the op-amp input, hence the dynamic range of the circuit is improved; by reducing the signal voltage at the virtual ground terminal, the effect of the finite low-frequency gain of the op-amp on the signal-processing characteristics of the stage is decreased. Both improvements are especially significant for low-supply voltage circuits, which have limited signal swings and where the op-amp gain may be low since headroom for cascoding may not be available. The proposed techniques are applicable to such important building blocks as voltage amplifiers, ADC and DAC stages, integrators and filters, sample-and-hold (S/H) circuits, analog delay stages, and comparators.

Sections II and III present the two basic techniques that are used to reduce the offset and low-frequency noise of op-amps, namely the autozero (AZ) and chopper stabilization (CHS) techniques. A clear distinction is made between autozeroing, which is a sampling technique, and CHS, which is a modulation technique, mainly with respect to their effect on the amplifier broadband noise. The correlated double sampling (CDS) technique is described in Section II as a particular case of AZ where, as its name indicates, the amplifier noise and offset are sampled twice in each clock period. Then, Section IV treats the most important practical issues at the transistor and circuit level that are faced when implementing the offset and noise reduction techniques discussed previously. Section V presents fundamental building blocks that are used for sampled-data analog signal processing. They are all realized as switched-capacitor (SC) circuits and therefore exploit the CDS technique not only for reducing the offset and the $1/f$ noise, but also to lower the sensitivity of the circuit performance to the finite amplifier gain. Examples of SC S/H stages, voltage amplifiers, integrators, and filters are

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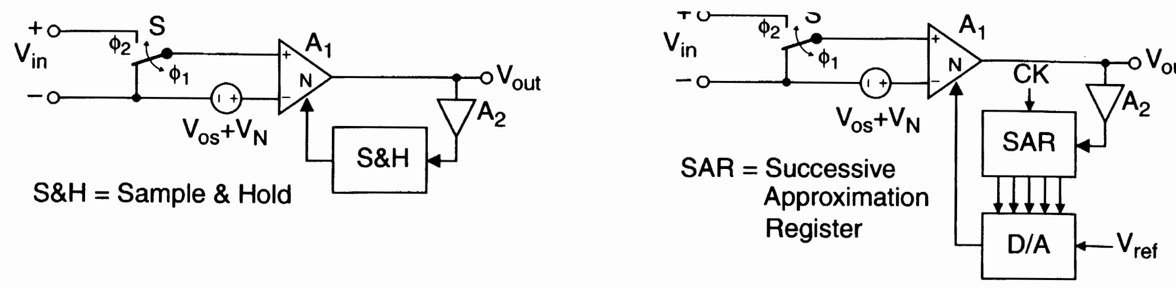
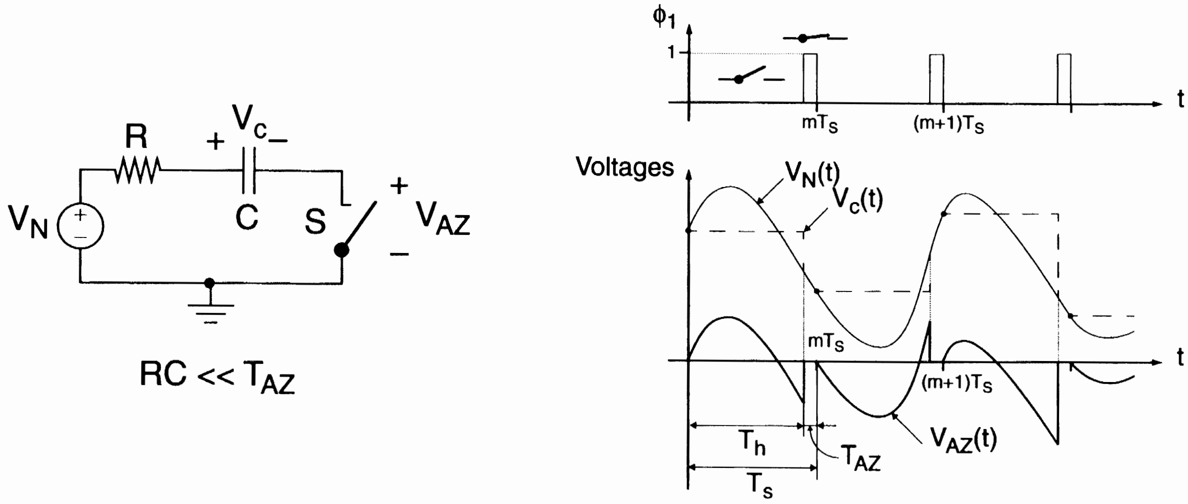
[1] This work is dedicated to Prof. Karoly Simonyi on his 80th birthday.



(a) (b)

Fig. 1. Basic autozeroed stages. (a) Analog offset control storage and (b) digital offset control storage.



(a) (b)

Fig. 2. (a) Basic AZ circuit and autozeroed signal: (b) shows voltages in (a).

presented. An example of the use of the CHS technique to realize a low-noise and low-offset micropower amplifier for intrumentation applications is presented in Section VI. Finally, a summary is given in Section VII, where the two techniques discussed in this paper are compared.

II. Autozeroing and Correlated Double Sampling Techniques

In this section, the principle of AZ and CDS techniques will be introduced and their effect on offset and noise analyzed.

A. Basic Principle

The basic idea of AZ is sampling the unwanted quantity (noise and offset) and then subtracting it from the instantaneous value of the contaminated signal either at the input or the output of the op-amp. This cancellation can also be done at some intermediate node between the input and the output of the op-amp, using an additional input port defined as the nulling input and identified with the letter N in the schematics of Fig. 1.

If the noise is constant over time (like a dc offset) it will be cancelled, as needed in a high-precision amplifier or high-resolution comparator. If the unwanted disturbance is low-frequency random noise (for example, $1/f$ noise), it will be high-pass filtered and thus strongly reduced at low frequencies but at the cost of an increased noise floor due to aliasing of the wideband noise inherent to the sampling process. The general principle of the AZ process will first be described considering only the input referred dc offset voltage V_{os} and will then be extended to the input referred random noise voltage V_N.

The AZ process requires at least two phases: a sampling phase (ϕ_1) during which the offset voltage V_{os} and the noise voltage V_N are sampled and stored, and a signal-processing phase (ϕ_2) during which the offset-free stage is available for operation. The two major categories of AZ are shown in Fig. 1. During the sampling phase (shown in Fig. 1), the amplifier is disconnected from the signal path, its inputs are short-circuited and set to an appropriate common-mode voltage. The offset is nulled using an auxiliary nulling input port N by means of an appropriate feedback configuration and/or a dedicated algorithm. The control quantity x_c is next sampled and stored, either in an analog form as a voltage using a S/H stage [Fig. 1(a)] or in a digital form, using for example a register [Fig. 1(b)]. The output V_{out} is forced to a small value in these particular configurations. The input terminals of the amplifier can afterwards be connected back to the signal source for amplification. If

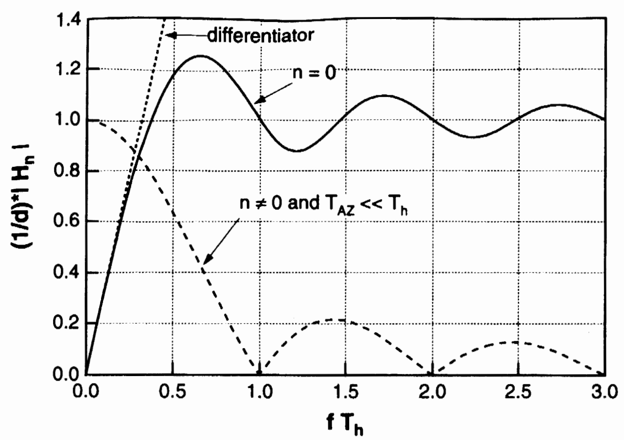


Fig. 3. Autozero baseband and foldover bands transfer functions.

it is used under the same conditions as during sampling, the amplifier will ideally be free from any unwanted offset.

B. The Effect of AZ on the Noise

The autozero principle can be used not only to cancel the amplifier offset but also to reduce its low-frequency noise, for example $1/f$ noise. But unlike the offset voltage, which can be considered constant, the amplifier's noise and particularly its wideband thermal noise component is time-varying and random. The efficiency of the AZ process for the low-frequency noise reduction will thus strongly depend on the correlation between the noise sample and the instantaneous noise value from which this sample is subtracted. The autocorrelation between two samples of $1/f$ noise separated by a time interval τ decreases much slower with increasing τ than it does for white noise, assuming they have the same bandwidth. The AZ process is thus efficient for reducing the $1/f$ noise but not the broadband white noise.

Another way of looking at the effect of AZ is to note that it is equivalent to subtracting from the time-varying noise a recent sample of the same noise. For dc or very low-frequency noise this results in a cancellation. This indicates that AZ effectively high-pass filters the noise.

In addition to this basic high-pass filtering process, since AZ is a sampling technique, the wideband noise is aliased down to the baseband, increasing the resulting in-band power spectral density (PSD) unless the system is already a sampled-data one.

The effects of AZ on the amplifier's noise can be better understood by analyzing the simple circuit shown in Fig. 2, where source V_N may represent the noise at the output of the amplifier in the autozero phase [see, i.e., Fig. 21(a)]. Each time switch S is closed, the output voltage V_{AZ} is reset to zero and the noise source voltage V_N appears across resistor R and capacitor C. Assuming $RC \ll T_{\mathrm{AZ}}$, at the end of the sampling phase (when switch S opens) the noise voltage V_N is sampled onto capacitor C. The output voltage becomes equal to the difference between the instantaneous voltage V_N and the voltage V_c stored on capacitor C. This eliminates the dc component of V_N, but not its time-varying part. It can be shown [8] that if source voltage $V_N(t)$ corresponds to a stationary random noise with a PSD $S_N(f)$, the PSD of the autozero voltage across the switch can be decomposed into two components: one caused by the baseband noise (which is reduced by the AZ process) and the other by the foldover components introduced by aliasing. Thus

$$S_{\mathrm{AZ}}(f) = \underbrace{|H_0(f)|^2 S_N(f)}_{\text{baseband}} + \underbrace{S_{fold}(f)}_{\text{foldover}} \tag{1}$$

where

$$S_{\text{fold}}(f) \equiv \sum_{\substack{n=-\infty \\ n\neq 0}}^{+\infty} |H_n(f)|^2 S_N\left(f - \frac{n}{T_s}\right). \tag{2}$$

The foldover component results from the replicas of the original spectrum shifted by the integer multiples of the sampling frequency. The baseband transfer function $|H_0(f)|^2$ is given by (see (3) at the bottom of the page) where $d \equiv T_h/T_s$ is the duty cycle of the clock signal [Fig. 2(b)]. The magnitude of $H_0(f)$ normalized to the duty cycle d is plotted as a function of fT_h in Fig. 3, which shows its high-pass characteristic. Note that for $\pi f T_h \ll 1$, $H_0(f)$ acts like a differentiator

$$|H_0(f)| \cong \pi f T_h. \tag{4}$$

It imposes a zero at the origin of frequency axis that cancels out any dc component present in $V_N(t)$. The other transfer functions $|H_n(f)|^2$ for $n \neq 0$ are derived in the Appendix. Their shape depends on the duty cycle d, but they all merge to a common function in the case the AZ time T_{AZ} can be considered much smaller than the hold time ($T_{\mathrm{AZ}} \ll T_h$)

$$|H_n(f)|^2 \cong [d \cdot \mathrm{sinc}\,(\pi f T_h)]^2 \quad \text{for } n \neq 0 \text{ and } T_{\mathrm{AZ}} \ll T_h \tag{5}$$

where $\mathrm{sinc}\,(x) \equiv \sin(x)/x$. $|H_n(f)|$ is plotted in Fig. 3.

The PSD at the output of the AZ circuit clearly depends on the PSD of the source which is autozeroed. The low-frequency input-referred noise PSD of an amplifier generally contains both a white and a $1/f$ noise component. It can be written in the following convenient form:

$$S_N(f) = S_0\left(1 + \frac{f_k}{|f|}\right) \tag{6}$$

$$|H_0(f)|^2 = d^2\left\{\left[1 - \frac{\sin(2\pi f T_h)}{2\pi f T_h}\right]^2 + \left[\frac{1-\cos(2\pi f T_h)}{2\pi f T_h}\right]^2\right\} \tag{3}$$

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