Reducing distortion from CMOS analog switches

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CMOS analog switches have become ubiquitous on the inputs and outputs of many electronic systems. They may be used to select between multiple input channels on data acquisition systems or to disable outputs during power-up or power-down events. In fact, analog switches have become so common that their operation is often taken for granted. But analog designers should be aware that semiconductor switches exhibit behavior quite unlike their mechanical cousins. For example, the resistance of a CMOS switch in the closed position, referred to as the on-resistance or R_{ON}, changes depending on the input voltage. This behavior is usually undesirable and can significantly distort the input signal in some applications.

To understand why CMOS switches behave in this manner, it is necessary to understand their basic construction and operation. A typical solid-state analog switch consists of two MOSFETs of opposite channel polarity and configured as a transmission gate as shown in Figure 1. The control voltages (C and $\overline{C}$) at the FET gates are dc voltages of opposite polarity. The switch is closed when the gate of the NMOS transistor is high and the gate of the PMOS transistor is low. Positive input voltages drive the V_{GS} of the PMOS more negative, decreasing the PMOS on-resistance. Therefore, the PMOS is the dominant current pathway for positive voltages. Conversely, negative voltages applied to the input terminal increase the gate-to-source voltage, V_{GS}, of the NMOS FET, decreasing its on-resistance and allowing current to flow through the NMOS pathway.

The basic switch architecture allows for both positive and negative voltages to be passed, but also causes the overall resistance of the switch to change with the input signal. Figure 2 is a plot of the on-resistance of the TS12A12511 switch versus the signal voltage range[1]. An R_{ON} "flatness" parameter may be included in the datasheet specification table to quantify the maximum deviation in the switch on-resistance over the signal range. For example, the R_{ON} flatness specification for the TS12A12511 is 1.6 Ω (typical).

A basic analog output circuit incorporating a CMOS switch is illustrated in Figure 3. Here the switch is used to disconnect the load from the output of an operational amplifier (op amp). Such applications of CMOS switches are very common in audio applications to suppress clicks and pops during the power-up or down of preceding circuitry.

Figure 1. CMOS transmission gate consisting of NMOS and PMOS transistors

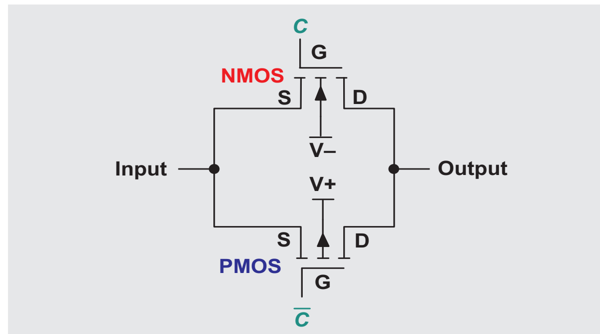


Figure 2. On-resistance variation of the TS12A12511 and example of R_{ON} flatness

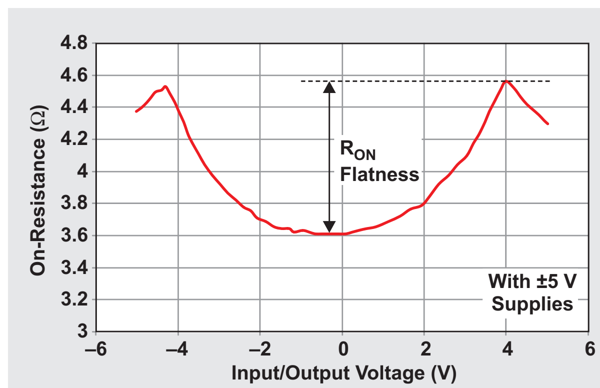
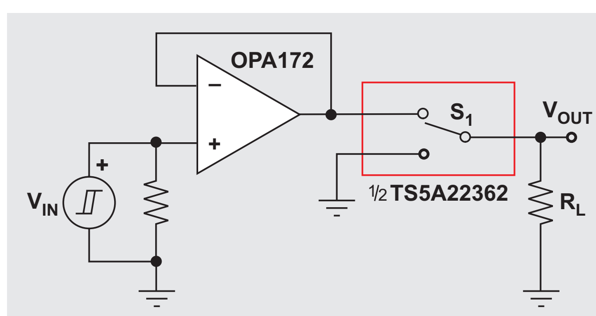


Figure 3. Signal distortion in a typical CMOS switch application



The switch on-resistance forms a voltage divider with the load resistance R_L and the output voltage is:

$$V_{OUT} = V_{IN}\frac{R_L}{R_{ON(S1)}+R_L} = \frac{V_{IN}}{\frac{R_{ON(S1)}}{R_L}+1} \quad (1)$$

In reality, the value of $R_{ON(S1)}$ is not a constant, but is a function of V_{IN}. As an example, assume that $R_{ON(S1)}$ is a linear function of the input voltage:

$$R_{ON(S1)}(V_{IN}) = \Delta R \times V_{IN} + R_O \quad (2)$$

In Equation 2, ΔR represents the change in switch on-resistance with input voltage and R_O is the resistance for an input signal of 0 V. In reality, the relationship between R_{ON} and V_{IN} is more complex, but assuming a linear relationship simplifies the analysis while still revealing the distortion mechanism.

Inserting Equation 2 for on-resistance back into Equation 1 for output voltage gives a new equation:

$$V_{OUT} = \frac{V_{IN}}{\frac{\Delta R}{R_L}V_{IN}+\frac{R_O}{R_L}+1} \quad (3)$$

For simplicity, a generic form of Equation 3 can be generated by substituting the constants A and B for terms in the above equation:

$$\text{Let } x = V_{IN},\ A = \frac{\Delta R}{R_L} \text{ and } B = \frac{R_0}{R_L}+1,$$

$$\text{then } V_{OUT} = f(x) = \frac{x}{Ax+B} \quad (4)$$

To show the introduction of distortion, this more generic equation can be written instead as its equivalent Maclaurin series (shown here to 5 terms):

$$f(x) = \frac{1}{B}x - \frac{A}{B^2}x^2 + \frac{A^2}{B^3}x^3 - \frac{A^3}{B^4}x^4 + \frac{A^4}{B^5}x^5 \ldots \quad (5)$$

Now a sine wave is inserted as the input signal with $x = \sin(2\pi ft)$:

$$f[\sin(2\pi ft)] = \frac{\sin(2\pi ft)}{B} - \frac{A}{B^2}\sin(2\pi ft)^2 + \frac{A^2}{B^3}\sin(2\pi ft)^3 - \frac{A^3}{B^4}\sin(2\pi ft)^4 + \frac{A^4}{B^5}\sin(2\pi ft)^5 \ldots \quad (6)$$

Using the power reduction rules for trigonometric functions and simplifying the equation, the individual terms for each harmonic can be grouped together as shown in Table 1. The Maclaurin series was abbreviated to five terms so the amplitude for the harmonics are approximations.

Although the on-resistance of a CMOS switch is almost never linearly related to the input voltage, this example provides some useful rules for reducing the distortion from analog switches. Looking at the equations for the individual harmonics, a reduction in distortion requires that either the value of A must be very small, or B must be very large. The latter option is very un-attractive in most applications. Recalling the equation for B:

$$B = \frac{R_0}{R_L}+1 \quad (7)$$

For B to be large, R_O must be much greater than R_L. Now the majority of the signal voltage is dropped across the switch, rather than the load resistor. The net effect is that the output signal is attenuated.

In most systems, it is more practical to reduce the value of A:

$$A = \frac{\Delta R}{R_L} \quad (8)$$

Examining the equation for A, it can be seen that if $\Delta R = 0$, the harmonic terms will be eliminated. Although this metric is constantly being improved in analog switches, the on-resistance is never completely independent of input voltage. An alternate, and more common solution, is to select a load resistance value that is much larger than the variations in the on-resistance. This solution is commonly used on analog inputs, where R_L is the input impedance of data acquisition circuitry and is typically very large.

Unfortunately, other applications that use analog switches do not have the luxury of specifying the load impedance. An example is switching the outputs to high-fidelity headphones. Furthermore, the distortion caused by even minute variations in switch on-resistance represents a surprising amount of distortion. The total harmonic distortion and noise (THD+N) of the circuit in Figure 3 was measured with a 2-V_{PP} signal and load resistances of 100 kΩ and 600 Ω. According to the TS5A22362 analog-switch datasheet, the on-resistance at 0 V (room temperature) is about 0.37 Ω. The on-resistance will vary approximately 0.115 Ω over the range of the 2-V_{PP} input signal.

Table 1: Approximate amplitude of fundamental and distortion harmonics for the example calculation

Fundamental	$\left(\frac{1}{B}+\frac{3A^2}{4B^3}+\frac{5A^4}{8B^5}\right)\sin(2\pi ft)$
2nd Harmonic	$\left(\frac{A}{2B^2}+\frac{A^3}{2B^4}\right)\sin\left(4\pi f+\frac{\pi}{2}\right)$
3rd Harmonic	$\left(\frac{A^2}{4B^3}+\frac{5A^4}{16B^5}\right)\sin(6\pi f+\pi)$
4th Harmonic	$\left(\frac{A^3}{8B^4}\right)\sin\left(8\pi f+\frac{3\pi}{2}\right)$
5th Harmonic	$\frac{A^4}{16B^5}\sin(10\pi f)$

The measured THD+N over frequency is given in Figure 4 for two load impedances. With the 100-kΩ load impedance, the THD+N is extremely low. In this case, the measurement is determined by the noise floor of the instrument, roughly 0.0005%. However, decreasing the load impedance to 600 Ω increases the distortion by an order of magnitude to 0.005%. This level of distortion may not be acceptable in many high-precision analog systems.

The distortion contribution from the switch is constant over frequency because the voltage-drop across the switch does not change over the measured bandwidth.

An FFT of the output signal at 1 kHz into a 600-Ω load (Figure 5) shows that the 2nd harmonic is dominant, but spurs are visible above the noise floor up to the 5th harmonic. The harmonics are due to the R_{ON} variations of the switch.

Conceivably, enclosing the switch inside the feedback loop of an amplifier allows for the additional distortion to be corrected, but this is not as simple as it may seem. The amplifier's feedback loop must still be closed when the switch is open, otherwise the amplifier output would saturate to one of the power supply rails. Closing the switch while the amplifier output is saturated could cause an undesirable transient voltage at the load.

One solution to this problem is shown in Figure 6. In this circuit topology, two switches are used. One switch, S_1, is the signal path for the load. The second switch, S_2, allows the op amp feedback loop to be closed around the first switch. S_2 contributes negligible additional distortion in the system because the op amp inverting input is a very high impedance.

With both switches configured as shown in Figure 6, resistor R_1 is in parallel with the pathway through S_1 and S_2. For minimal distortion, the dominant feedback pathway should be through the switches and not through R_1. Therefore, the on resistance of the switches should be much less than R_1:

$$R_1 \gg R_{ON(S1)} + R_{ON(S2)} \tag{9}$$

Considering the 0.37-Ω on-resistance of an analog switch such as the TS5A22362, this requirement is easily accomplished. But other switch parts, such as the extremely popular CD4066B, have typical on-resistances greater than 100 Ω.

When the switches are moved to their alternate position in order to disconnect the load from the amplifier output, R_1 closes the feedback loop of the op amp. Stability must always be considered when placing a resistor in the feedback path of an op amp. The feedback resistor interacts with the input capacitance to degrade the feedback-loop

Figure 4: THD+N measurement of the circuit in Figure 3

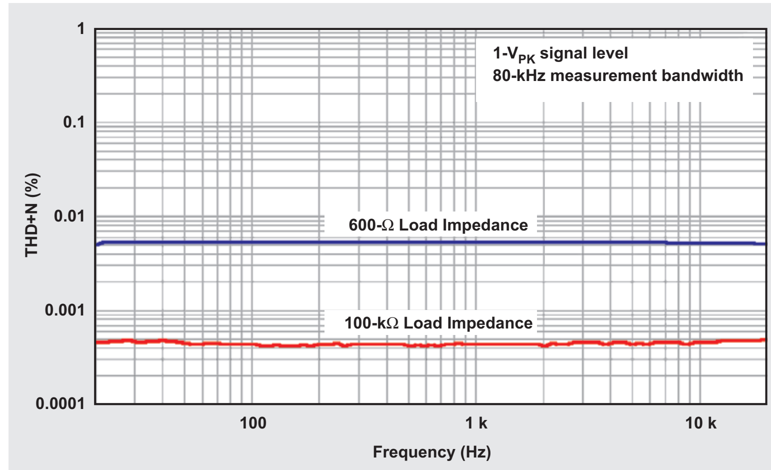


Figure 5. Spectrum of a 1-V_{PK}, 1-kHz sine wave at the output of the circuit in Figure 3

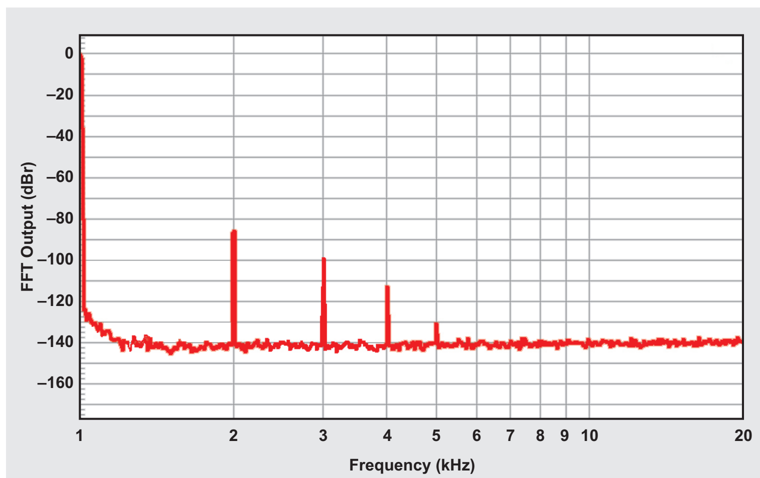
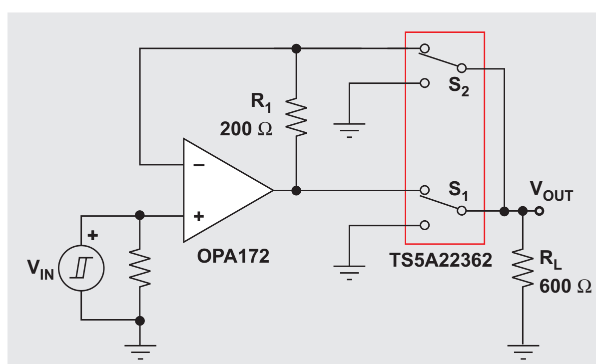


Figure 6: A dual-switch solution to close the amplifier feedback loop



phase margin. A conservative rule of thumb is given in Equation 10, the derivation of which is given in Reference 2.

$$\frac{1}{20\pi f_{GBW}(C_{CM} \parallel C_{DM})} \geq R_1 \qquad (10)$$

Where f_{GBW} is the op amp gain-bandwidth product and C_{CM} and C_{DM} are the op amp common-mode and differential input capacitances, respectively. Inserting the appropriate values for the OPA172, a precision op amp, gives a maximum value for R_1 of 198.9 Ω. A 200-Ω resistor is reasonably close to the calculated value to avoid stability concerns.

$$\frac{1}{20\pi(10\text{ MHz})(8\text{ pF})} = 198.9 \geq R_1 \qquad (11)$$

The circuit in Figure 6 was tested in the previously described manner and the results are given in Figures 7 and 8. By enclosing the switch inside the feedback loop of the op amp, the additional distortion from the R_{ON} variation has been effectively eliminated. The THD+N measurement over frequency for both load impedances (600 Ω and 100 kΩ) are identical, and at the noise floor of the measurement instrument.

Examining the FFT of the output signal (Figure 8) shows that the additional harmonics from the TS5A22362 are now below the noise floor of the measurement instrument.

For high-performance analog systems where harmonic distortion must be minimized, enclosing a CMOS analog switch inside the feedback loop of an op amp can greatly improve performance. The circuit topology shown in Figure 6 reduces harmonic distortion from the switch and also allows the amplifier output to be completely disconnected from the load. The feedback loop of the op amp is closed regardless of the switch configuration, preventing the amplifier output from saturating and causing unwanted voltage transients when the switch is closed. Furthermore, a CMOS analog switch with extremely low R_{ON} variation is no longer absolutely crucial, which can potentially reduce system costs.

Acknowledgements

The author wishes to acknowledge John Xu, TI analog field applications engineer, whose idea was the initial inspiration for this work.

Figure 7: THD+N measurement of the circuit in Figure 6

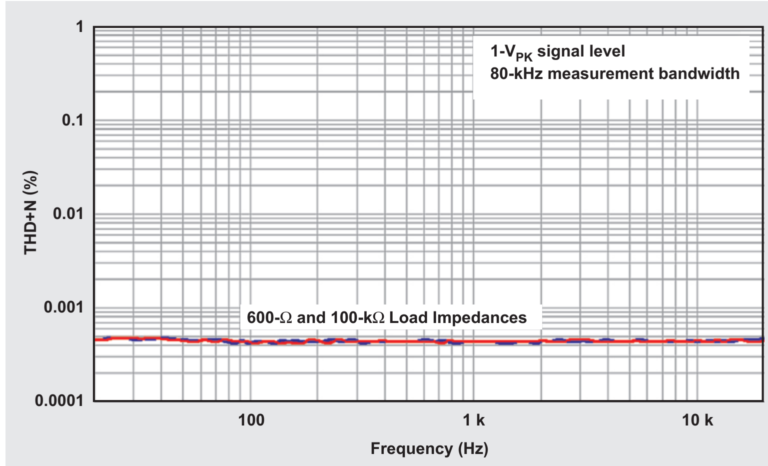
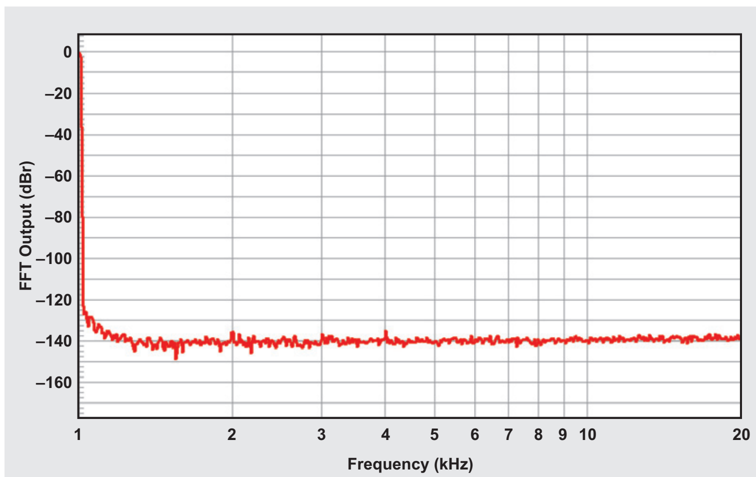


Figure 8: Spectrum of a 1-V$_{PK}$, 1-kHz sine wave at the output of the circuit in Figure 6



Reference

1. TS12A12511 SPDT Analog Switch datasheet, Texas Instruments, 2015. Available: **www.ti.com/1q15-TS12A12511**

Related Web sites

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www.ti.com/1q15-CD4066B

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