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The Design of a Bootstrapped Sampling Circuit

Bootstrapped samplers serve as an integral component of analog-to-digital converters (ADCs). The bootstrapping action reduces the distortion and improves the speed with a minimal power penalty. In this article, we present the step-by-step design of this circuit to see how the analog mind goes about such a task. The reader is encouraged to review the sampler's detailed operation [1]–[4] before continuing with this article.

Basic Operation

As explained in [4], metal–oxide semiconductor (MOS) sampling switches exhibit two sources of distortion: their on-resistance and channel charge vary with their gate source voltage, V_{GS} , and hence with the analog input level. These effects can be minimized if V_{GS} is maintained constant in the sampling phase, e.g., by tying a battery between the gate and the source terminals [Figure 1(a)]. We say the gate is “bootstrapped” to the source, as the former's voltage varies in unison with the latter's.

The circuit in Figure 1(a) requires two modifications so it can act as a sampler. First, a means of disengaging V_B from M_1 is necessary so that M_1 can be turned off. Second, the battery can be approximated by a capacitor, C_B , but the charge on C_B must be periodically refreshed. We thus arrive at the basic topology shown in Figure 1(b). In the sampling mode (also called the *track mode*), M_2 and M_3 are on, reducing the cir-

cuit to that in Figure 1(a). In the hold mode, M_4 – M_6 are on, grounding the gate of M_1 and charging C_B to V_{DD} . The choice of PMOS and NMOS devices for these switches is described in [3] and [4].

Design Specifications

We wish to design a differential sampler for the front end of a Nyquist-rate ADC with a resolution of 10 b and a sampling rate of 5 GHz. Of the clock period of $T_{CK} = 200$ ps, we allocate one half to the sampling mode and the other half to the hold mode. The design proceeds in a 28-nm CMOS process in the slow–slow corner, at a temperature of 75°C and with a worst-case supply of $1\text{ V} - 5\% = 0.95\text{ V}$. We assume a single-ended input range from 0.25 to 0.75 V, which translates to a least-significant-bit (LSB) size of $1\text{ V}/1,024 \approx 1\text{ mV}$ for the differential ADC. The bootstrapped sampler accommodates greater swings, but the stage preceding it—typically, a buffer—may not. For an N -bit ADC sensing a signal of the form $A\cos\omega_{in}t$, the ideal signal-to-noise ratio (SNR) is given by $(A^2/2)/(\Delta^2/12)$, where

Δ is the LSB and equal to $2A/2^N$. It follows that $\text{SNR} = 6.02N + 1.76\text{ dB}$, which amounts to 62 dB for $N = 10$. In our differential system, $A = 0.5\text{ V}$.

The ADC specifications readily dictate certain values for the sampler. First, the harmonic distortion (HD) must remain well below -62 dB . Our ambition is to achieve an HD of roughly -65 dB in the worst case, namely, when $A = 0.5\text{ V}$ and $f_{in} = \omega_{in}/(2\pi)$ is near 2.5 GHz (i.e., at the Nyquist rate). Second, the noise introduced by the sampler must negligibly degrade the overall SNR. Thus, the kT/C noise associated with M_1 and C_1 in Figure 1(b) must be sufficiently small. We express the SNR of the sampler–ADC cascade as

$$\text{SNR} = \frac{\frac{A^2}{2}}{\frac{\Delta^2}{12} + \frac{2kT}{C_1}}, \quad (1)$$

where the factor of 2 in the denominator accounts for the noise in a differential sampling configuration. Let us target a 1-dB penalty due to kT/C noise, divide the SNR in (1) by the ideal SNR equation, take

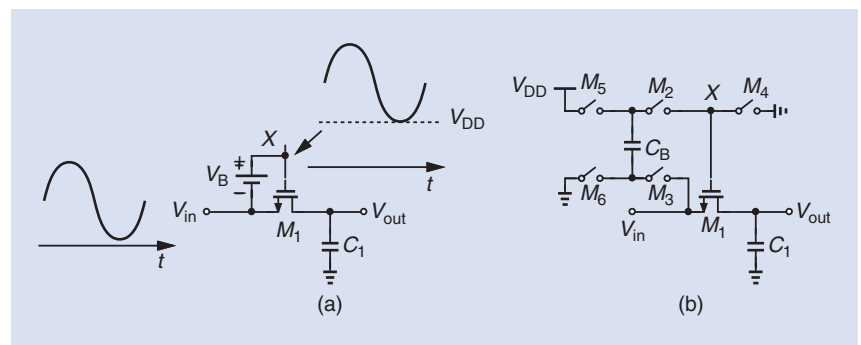


FIGURE 1: (a) A simple bootstrapped switch. (b) A more complete implementation.

10log of the result, and equate it to -1 dB. With $\Delta = 1$ mV and $T = 75^\circ\text{C} = 348\text{K}$, we obtain $C_1 = 445$ fF, which we round up to 0.5 pF.

Third, the low-pass filter formed by M_1 and C_1 in Figure 1(b) must negligibly attenuate the input signal in the track mode. The output amplitude is equal to $A/\sqrt{1 + R_{\text{on}1}^2 C_1^2 \omega_{\text{in}}^2}$, where $R_{\text{on}1}$ denotes the on-resistance of M_1 . We aim for an attenuation less than 0.5 dB at the Nyquist rate, arriving at $R_{\text{on}} = 44.5\Omega$. We should select M_1 so that it is wide enough to ensure such a low resistance with $V_{\text{GS}1} \approx V_{\text{DD}} = 0.95\text{V}$. This constraint yields $W_1 \approx 10\mu\text{m}$ if $L_1 = 28\text{nm}$.

Design Procedure

To understand the effect of device nonidealities, we follow an incremental design procedure. Specifically, we proceed in five steps: 1) we keep the battery in Figure 1(a) and examine the distortion in only the track mode; 2) we allow the circuit to act as a sample-and-hold stage but, except for M_1 in Figure 1(b), use ideal switches; 3) we change M_2 and M_4 to MOS devices; 4) we change the remaining switches to MOSFETs; and 5) we replace the battery with a capacitor. In each case, we apply the full-scale input amplitude and measure the harmonic content of the differential output at a moderate input frequency and at the Nyquist rate. The incremental approach also permits us to optimize the design in every step. We prefer to select minimum widths and lengths for the transistors unless there is a compelling rea-

son not to do so. In this article, all channel lengths are equal to 28nm .

Simulation Issues

In the design of sampling circuits, the clock and the input frequencies must be carefully chosen. If f_{CK} is an integer multiple of f_{in} , then the circuit samples only certain points on the input waveform [Figure 2(a)], failing to reveal its true performance. Even if we select $f_{\text{CK}}/f_{\text{in}} = P/Q$, where P and Q are integers, we still have $QT_{\text{in}} = PT_{\text{CK}}$, where $T_{\text{in}} = 1/f_{\text{in}}$. This means that every Q cycles of the input exactly coincide with P cycles of the clock; thus, only Q values of the input are periodically sampled. For this reason, we should select f_{CK} and f_{in} such that the samples gradually slide along the input sinusoid and eventually assume all (or most) of the input values [Figure 2(b)]. This is accomplished if $f_{\text{CK}}/f_{\text{in}}$ is an irrational number. For simulations, we simply guarantee that P/Q yields a long periodicity. For example, with $f_{\text{CK}}/f_{\text{in}} = 5\text{GHz}/570\text{MHz} = 500/57$, the sampled points repeat themselves after every 500 clock cycles; i.e., 500 distinct points of the input voltage are collected.

Another issue relates to how we process the output time-domain waveform in Figure 1(b) to go to the frequency domain. Since the ADC following the sampler senses only the held values on C_1 , the time points taken by the fast Fourier transform (FFT) must be confined to the hold mode [Figure 3(a)]. That the ADC digitizes only the held values also sug-

gests that the sampler-ADC cascade equivalently multiplies the analog input by a train of impulses in the time domain [Figure 3(b)] and hence convolves the input spectrum with a train of impulses in the frequency domain [Figure 3(c)]. We call the range between $-f_{\text{CK}}/2$ and $+f_{\text{CK}}/2$ the first Nyquist zone and generally confine our inspection of unwanted components to this region.

If the FFT senses only the values at t_1, t_2 , and so on, the resulting spectrum is given by the convolution of the input spectrum and a train of impulses. This point proves useful in understanding the FFT results. For example, if the sampler introduces third-order distortion at $3f_{\text{in}}$ and this component lies above $f_{\text{CK}}/2$, then it is aliased to the first Nyquist zone. Depicted in Figure 3(d), this phenomenon causes the third harmonic to land at $f_{\text{CK}} - 3f_{\text{in}}$.

Track-Mode Distortion

We begin by simulating two instances of Figure 1(a) with differential inputs and $f_{\text{in}} = 570\text{MHz}$. As mentioned, $W_1 = 10\mu\text{m}$ for now. We surmise that this test yields a lower bound for the distortion because higher input frequencies and the sample-and-hold action tend to introduce greater nonlinearity. Figure 4 plots the FFT of the differential output. As expected, even harmonics are absent. The third and fifth harmonics are 79 and 97 dB below the fundamental, respectively. We denote the relative level of the former by HD_3 and that of the latter by HD_5 . Next, we increase f_{in} to 2.47GHz and repeat the simulation. Figure 5(a) shows the time-domain outputs, revealing a peak-to-peak single-ended swing of 483mV , which satisfies our attenuation constraint of 0.5 dB. The output spectrum is plotted in Figure 5(b), exhibiting a third harmonic at -56 dB. This high level of distortion occurs because $R_{\text{on}1}$ varies significantly due to the body effect. As explained in [4], the output phase then varies considerably with the input voltage. We must therefore reduce $R_{\text{on}1}$. Changing W_1 from 10 to

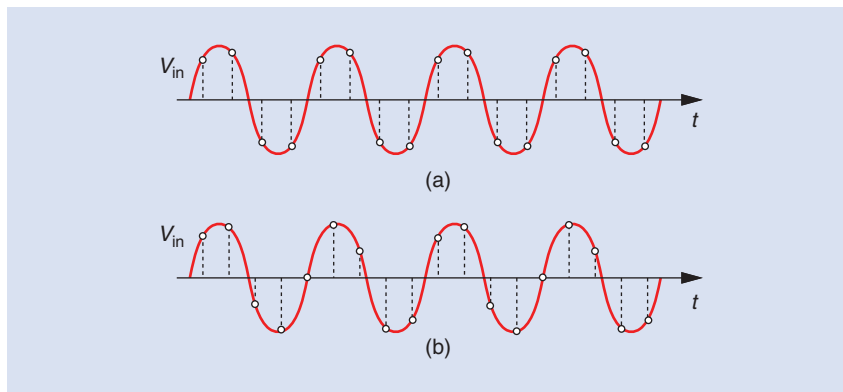


FIGURE 2: Sampling the input with (a) an integer $f_{\text{CK}}/f_{\text{in}}$ ratio and (b) an irrational $f_{\text{CK}}/f_{\text{in}}$ ratio.

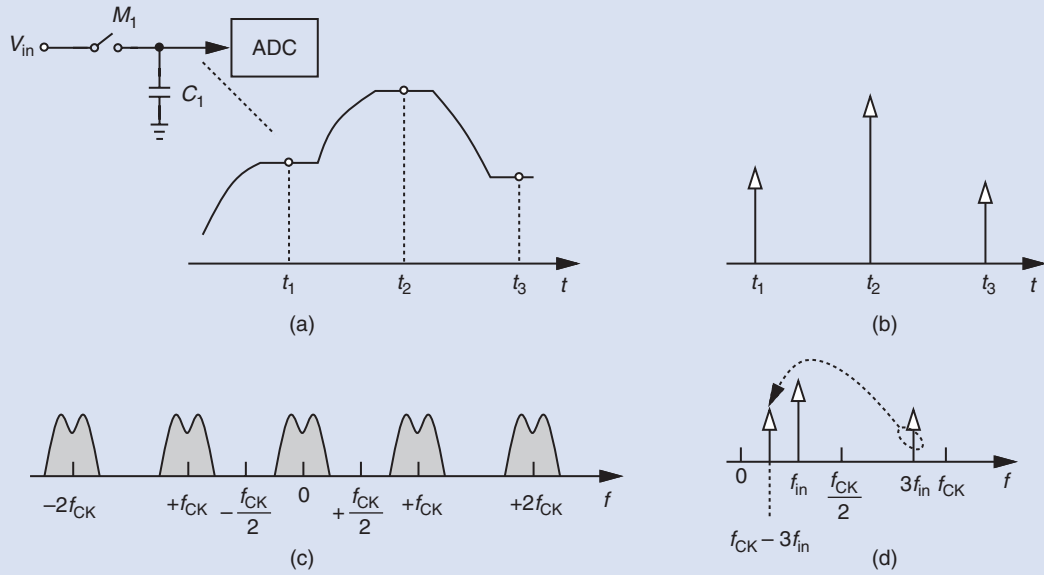


FIGURE 3: (a) Sensing the held values via an ADC. (b) The equivalent ADC output in the time domain. (c) The corresponding spectrum. (d) The aliasing of the third input harmonic.

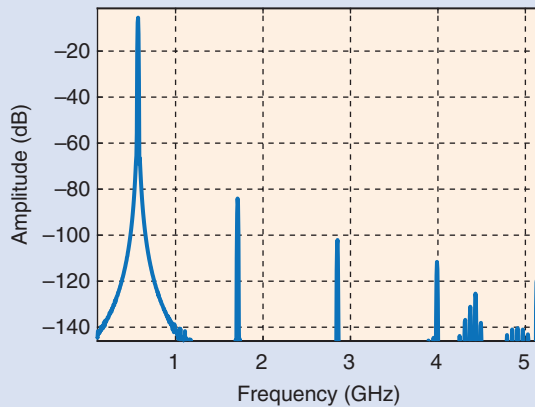


FIGURE 4: The track-mode output spectrum for $f_{in} = 570$ MHz.

20 μm yields the spectrum shown in Figure 5(c), achieving an HD_3 of -66 dB and an HD_5 far below this level. We hereafter assume $W_1 = 20$ μm .

Basic Sampler

In the next step of our design effort, we add two ideal switches to Figure 1(a) so that M_1 is turned on and off by the clock [Figure 6(a)]. To create some similarity to the actual circuit, we define an on-resistance of 50 Ω for M_2 and M_4 . Figure 6(b) plots the output spectrum for $f_{in} = 570$ MHz. The third harmonic is at 1.71 GHz and has a relative level of -83 dB. The fifth harmonic, on the other hand,

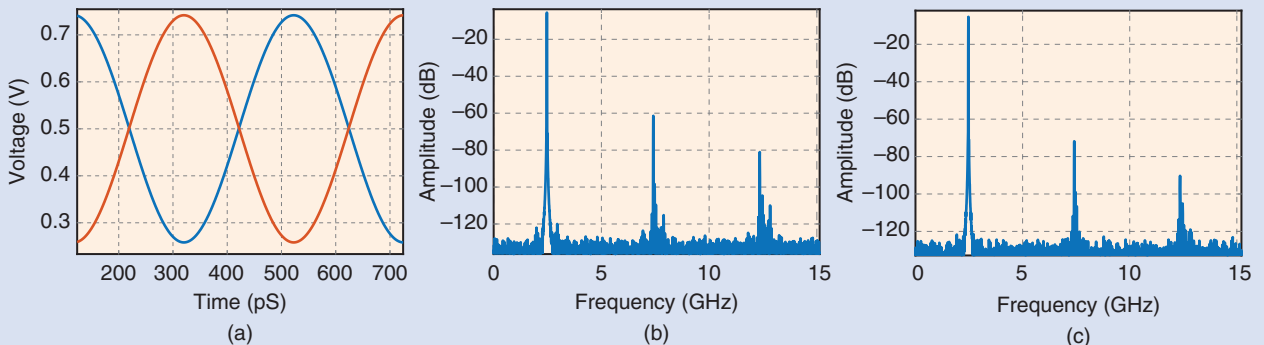


FIGURE 5: (a) The track-mode output waveforms for $f_{in} = 2.47$ GHz. (b) The corresponding spectrum. (c) The reduced distortion as a result of doubling W_1 .

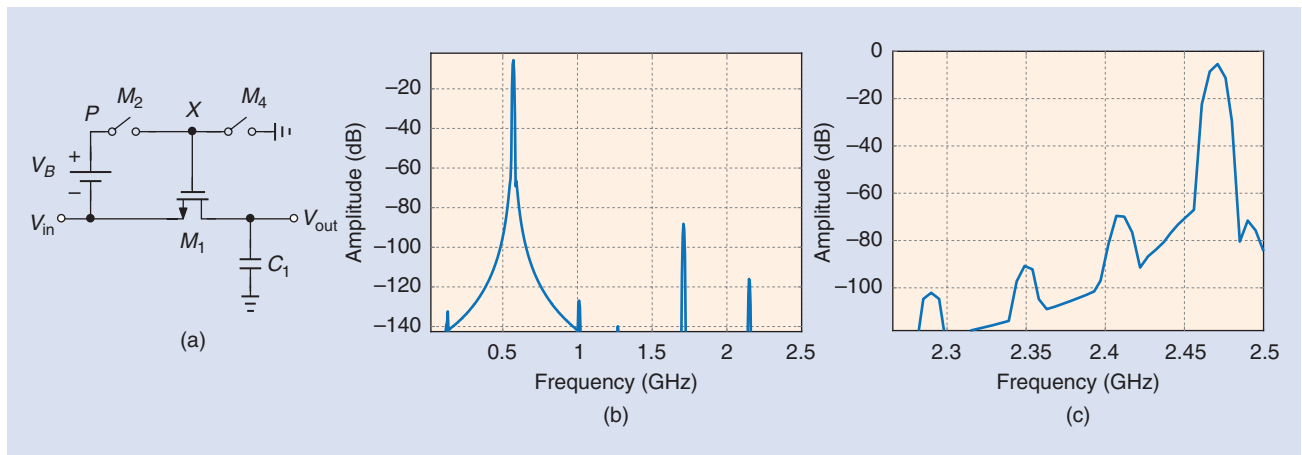


FIGURE 6: (a) A sampler using ideal switches M_2 and M_4 . (b) The output spectrum for $f_{in} = 570$ MHz. (c) The output spectrum for $f_{in} = 2.47$ GHz.

is aliased to $f_{CK} - 5f_{in} = 2.15$ GHz, exhibiting a normalized level of -110 dB. Next, we raise f_{in} to 2.47 GHz and note that its third harmonic appears at $f_{CK} - 3f_{in} = 2.41$ GHz. Depicted in Figure 6(c), the output spectrum yields $HD_3 = -64.4$ dB, 1.3 dB higher than the track-mode distortion found in the preceding. This can

be attributed to the channel charge injected by M_1 onto C_1 in Figure 6(a); even though V_{GS1} is constant, the transistor's threshold is not, modulating the charge as V_{in} varies.

Sampler With MOS Switches

At this point, we wish to implement M_2 and M_4 in Figure 6(a) by using MOS devices. We recognize that M_4 must be an NMOS device because it pulls X to the ground, whereas M_2 must be a PMOS device as it ties X to a high potential (Figure 7). But we must also decide to which node the n-well of M_2 should be connected. If attached to X , the n-well is drawn toward the ground by M_4 when M_2 is off, thereby forward-biasing the junctions between the n-well and the source and the drain of M_2 . To avoid this issue, we tie the n-well to node P .

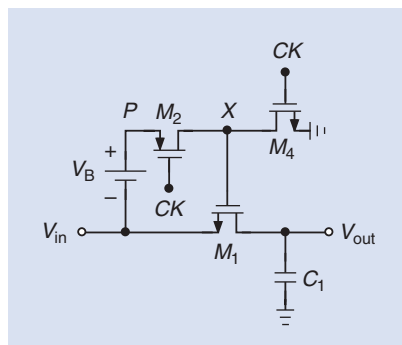


FIGURE 7: The realization of M_2 and M_4 .

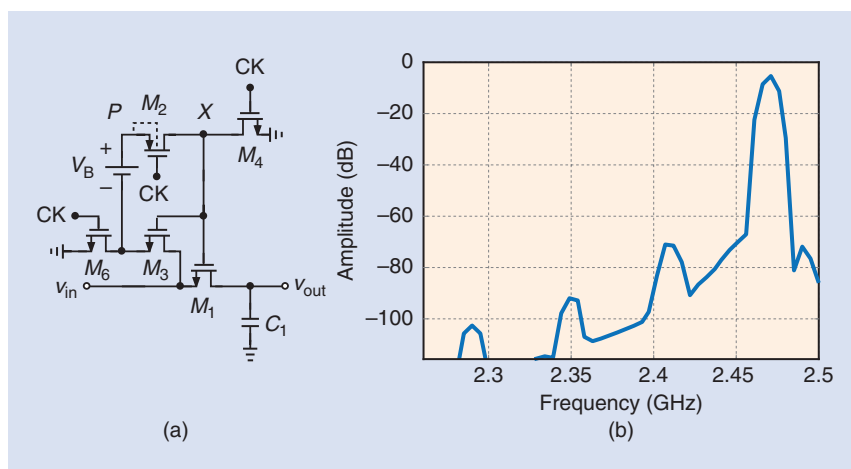


FIGURE 8: (a) The addition of switches to disengage V_B in the hold mode. (b) The output spectrum for $f_{in} = 2.47$ GHz.

The minimum widths of M_2 and M_4 are dictated by the following constraints. We note that R_{on2} and the total capacitance at X —primarily, the gate source capacitance of M_2 —form a low-pass filter, attenuating the input swing as it reaches X . Thus, R_{on2} must be chosen so that it provides a bandwidth far beyond 2.5 GHz. This is readily possible with a W_2 of a few microns. We select $W_2 = 2.5 \mu\text{m}$. Moreover, M_4 must pull X down at a high slew rate so as to rapidly turn off M_1 . If it does not, the on-resistance of M_1 increases slowly, causing distortion. We select $W_4 = 2.5 \mu\text{m}$ for now. For the clock driving these two switches, we choose 10-ps rise and fall times, as a representative in 28-nm technology.

Upon simulating the topology of Figure 7 with $f_{in} = 570$ MHz and 2.47 GHz, we obtain an HD_3 of -68 and -58 dB, respectively. Why is the linearity degraded by such a large amount? The reason is that M_2 fails to turn off for part of the input swing: if $V_{in} > 0.5$ V, then $V_P > 1.5$ V, keeping M_2 on, even if its gate is raised to 1 V. Consequently, M_4 cannot pull X to zero, and M_1 does not completely turn off.

To resolve this issue, we must remove the bootstrapping action in the hold mode. That is, the battery must be disconnected from the analog input. We thus add switches M_3 and M_6 [Figure 8(a)]. Now, when CK is high, V_B generates $V_P = V_{DD}$, allowing M_2 to turn off. The gate of M_3 could not be driven by the rail-to-rail clock because

the on-resistance of M_3 would vary considerably in the track mode. For this reason, we bootstrap its gate to V_{in} in a manner similar to M_1 . The on-resistance of M_3 should be comparable to that of M_2 and not limit the bandwidth at node X . We tentatively choose $W_3 = 2.5\mu\text{m}$ and $W_6 = 1\mu\text{m}$. Simulations indicate an HD_3 of -84 and -60 dB for $f_{in} = 570$ MHz and $f_{in} = 2.47$ GHz, respectively. The latter case is shown in Figure 8(b). The circuit performs satisfactorily at this stage of the design.

In the last step, we replace the bootstrapping battery with a capacitor, as indicated in Figure 1(b). This requires that M_5 be added so that C_B can be charged to V_{DD} in the hold mode. Several questions arise. First, to which node should the gate of M_5 be connected? If driven as in Figure 9(a), M_5 fails to turn off when \overline{CK} is high because $V_P = V_{in} + V_{DD}$ can reach 1.75 V. We therefore surmise that the gate of M_5 must be bootstrapped to V_{in} as well [Figure 9(b)]. Second, to which node should the n-well of M_5 be connected? For the same reason mentioned for M_2 , this n-well must be tied to node P .

The third question relates to the minimum acceptable value of C_B . Two effects play a role here. When C_B is switched to X , it experiences charge sharing with the total parasitic capacitance at this node, thereby providing a bootstrapping voltage that is less than

V_{DD} . The parasitics at X include the gate capacitances of M_1, M_3 , and M_5 (the first two conducting and the third remaining off) and the drain capacitances of M_2, M_4 , and M_5 . Thus, C_B must be large enough to minimize the voltage loss. For C_B to fully charge to V_{DD} during the hold mode, the series combination of M_5, C_B , and M_6 must exhibit a time constant that is less than half of the clock period. As an example, with $T_{CK}/2 = 100$ ps and $C_B = 0.25$ pF, the total on-resistance of M_5 and M_6 must remain lower than 400Ω .

We select $C_B = 0.25$ pF and $W_5 = 2.5\mu\text{m}$ in Figure 9(b) and increase W_6 from 1 to $2.5\mu\text{m}$. The simulation yields an HD_3 of -63 dB at $f_{in} = 2.47$ GHz. To assess the bootstrapping ability of C_B , we plot its voltage as a function of time [Figure 9(c)]. We observe that C_B charges to only 0.895 V (rather than to $V_{DD} = 0.95$ V) due to the long time constant. Moreover, the bootstrapping voltage is less than 0.83 V due to charge sharing. From the difference between 0.895 and 0.83 V, we estimate a parasitic capacitance of 20 fF at node X .

To improve the charge replenishment of C_B , we must further widen M_5 and M_6 , and to remedy the charge sharing, we must increase the value of C_B . Of course, an excessively wide M_5 raises the parasitic at X and exacerbates the problem of charge sharing. Let us then choose $C_B = 500$ fF, $W_5 = 5\mu\text{m}$, and $W_6 = 5\mu\text{m}$. Simula-

tions still suggest an HD_3 of -63 dB at $f_{in} = 2.47$ GHz. We face diminishing returns, partially because the large parasitic at X now creates a long time constant with M_4 , slowing down the turn-off transition of M_1 . This is alleviated by increasing W_4 to $5\mu\text{m}$. The circuit now exhibits an HD_3 of -65 dB at 2.47 GHz.

Device Stress Issues

The topology of Figure 9(b) applies a drain source or gate source voltage well above 1 V to two of the MOSFETs in the track mode. Called *device stress*, this effect degrades the transistors' performance over time. We recognize that, when it is off, M_4 experiences a peak drain source voltage of $V_{in} + V_{DD} \approx 1.75$ V. Similarly, M_2 sees the same voltage difference between its gate and its source. To protect the former, we place a cascode device in series with it. As depicted in Figure 10(a), M_8 shields M_4 , ensuring that $V_{DS4} < V_{DD}$; M_8 itself sustains a maximum V_{DS} of $1.75\text{ V} - (V_{DD} - V_{TH4}) \approx 1.05$ V. The series action of these two devices demands that they both be $10\mu\text{m}$ wide.

The solution for reducing the stress on M_2 in Figure 9(b) is more complex. We note that the only possibility is to allow the gate voltage of M_2 to change in unison with V_X and hence with V_{in} . That is, this gate must be connected to V_{in} in the track mode and to V_{DD} in the hold mode. Figure 10(b) presents an

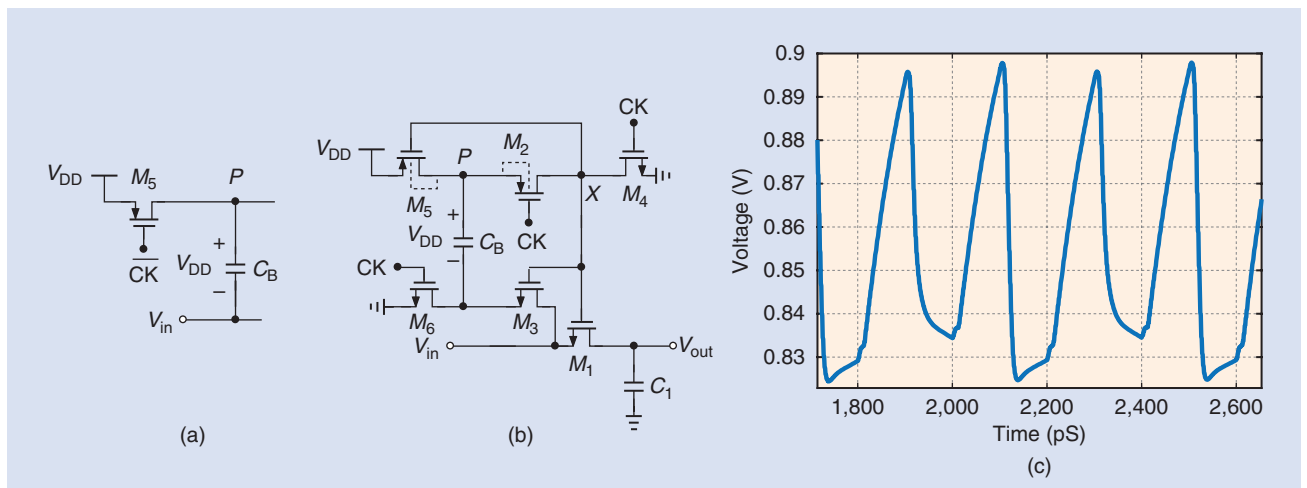


FIGURE 9: (a) A PMOS pull-up device driven by \overline{CK} . (b) Bootstrapping the gate of M_5 to V_{in} through V_X . (c) The voltage across C_B .

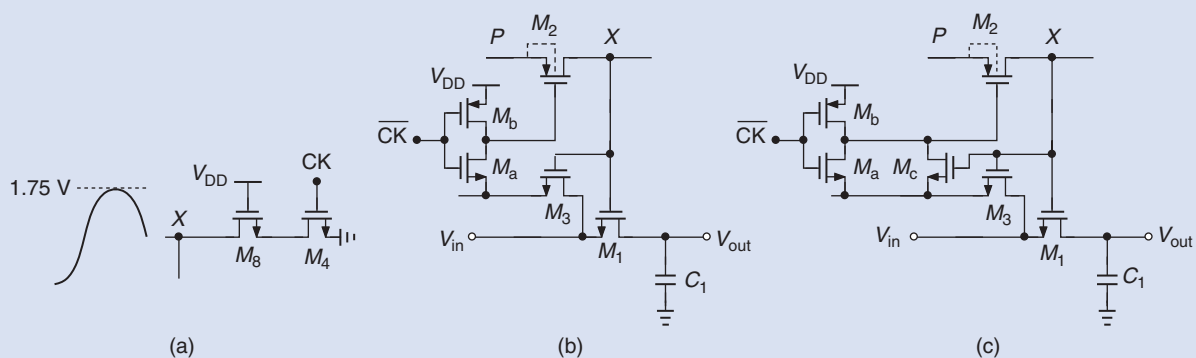


FIGURE 10: (a) The use of M_8 to shield M_4 . (b) The addition of M_a and M_b to bootstrap the gate of M_2 . (c) The addition of M_c to provide a low resistance in series with the gate of M_2 .

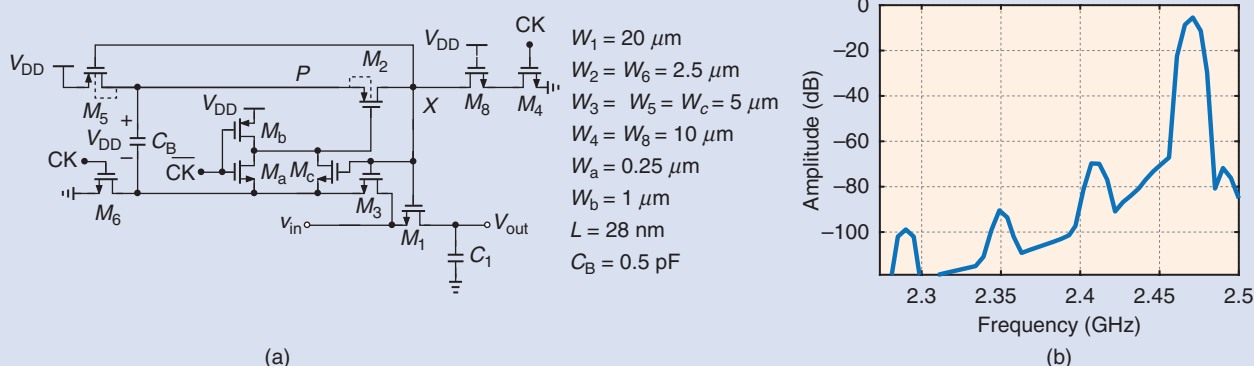


FIGURE 11: (a) The final design. (b) The final design's output spectrum for $f_{in} = 2.47$ GHz.

implementation example, where M_a and M_b , respectively, perform these tasks. However, if \overline{CK} reaches only V_{DD} , then M_a turns off as V_{in} approaches its peak value of 0.75 V. This issue is resolved as depicted in Figure 10(c), where M_c , with its gate voltage bootstrapped to V_{in} , provides a low resistance in series with the gate of M_2 . While appearing redundant, M_a is still necessary for the proper start-up of the circuit. We select $W_a = 0.25\mu\text{m}$, $W_b = 1\mu\text{m}$, and $W_c = 2.5\mu\text{m}$, the last one wide enough to avoid limiting the bandwidth at the gate of M_2 .

Simulation of the modified circuit yields an HD_3 of -63.4 dB. We surmise that the bandwidth at the gate of M_2 is still limited by the on-resistance of M_c . We then double W_c . The final design appears in Figure 11(a), and its output spectrum is in Figure 11(b). The HD_3 is

-64.3 dB, the HD_5 is approximately 20 dB lower, and the total power consumption is roughly 1 mW. The principal factor limiting the linearity is that the clock pulses arriving at node X are only some 80 ps wide, providing insufficient track time.

Input Current Issues

Recall that we initially selected an on-resistance of roughly 45Ω for the sampling switch in Figure 1(a) to ensure an attenuation less than 0.5 dB at $f_{in} = 2.5$ GHz. We later doubled the width of M_1 to meet the linearity requirement. With $R_{on1} \approx 23\Omega$, the circuit draws a large input current when operating near the Nyquist rate. This occurs because the previous sample on C_1 and the present input value at the beginning of the track mode can differ by as much as the entire single-ended swing, about 0.5 V. Thus, M_1 carries an initial cur-

rent of $0.5\text{ V}/R_{on1} \approx 22$ mA. This current must be provided by the stage preceding the sampler, typically, a buffer. From another perspective, if the buffer must not limit the sampling speed, its output impedance must be well below 23Ω , a great challenge in circuit design.

References

- [1] M. Dessouky and A. Kaiser, "Input switch configuration suitable for rail-to-rail operation of switched op amp circuits," *Electron. Lett.*, vol. 35, no. 1, pp. 8–10, Jan. 1999. doi: 10.1049/el:19990028.
- [2] A. M. Abo and P. R. Gray, "A 1.5-V 10-bit, 14.3 MS/s CMOS pipeline analog-to-digital converter," *IEEE J. Solid-State Circuits*, vol. 34, no. 5, pp. 599–606, May 1999. doi: 10.1109/4.760369.
- [3] M. Dessouky and A. Kaiser, "Very low-voltage digital audio $\Delta\Sigma$ modulator with 88-dB dynamic range using local switch bootstrapping," *IEEE J. Solid-State Circuits*, vol. 36, pp. 349–355, Mar. 2001. doi: 10.1109/4.910473.
- [4] B. Razavi, "The bootstrapped switch," *IEEE Solid State Circuits Mag.*, vol. 7, no. 3, pp. 12–15, Summer 2015. doi: 10.1109/MSSC.2015.2449714.