Video multiplexer uses high-speed op amps
Bruce Carter, Texas Instruments, Dallas, TX

Video multiplexers route video from several sources to a single channel. Low-end consumer products use CMOS analog switches and multiplexers, such as the 4066 and 4051. Unfortunately, these devices have a series on-resistance that ranges from approximately 100\(\Omega\) to 1 k\(\Omega\), a resistance that is not constant with video level and that appears in series with the signal. The traditional way of solving this problem is by buffering the analog-switch outputs with transistor stages. With this approach, the characteristics of the CMOS switch and the buffer stage degrade video performance. However, if you forget the multiplexing action for a moment and consider just the buffer-amplifier function, you will see that a better approach exists. It must present high enough input impedance to the switch that a 1-k\(\Omega\) switch resistance is inconsequential and that variation in resistances does not affect video level and that appears in series with the signal. The traditional way of solving this problem is by buffering the analog-switch outputs with transistor stages. With this approach, the characteristics of the CMOS switch and the buffer stage degrade video performance. However, if you forget the multiplexing action for a moment and consider just the buffer-amplifier function, you will see that a better approach exists. It must present high enough input impedance to the switch that a 1-k\(\Omega\) switch resistance is inconsequential and that variation in resistances does not affect video level and that appears in series with the signal. The traditional way of solving this problem is by buffering the analog-switch outputs with transistor stages. With this approach, the characteristics of the CMOS switch and the buffer stage degrade video performance. However, if you forget the multiplexing action for a moment and consider just the buffer-amplifier function, you will see that a better approach exists. It must present high enough input impedance to the switch that a 1-k\(\Omega\) switch resistance is inconsequential and that variation in resistances}

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**Figure 1**

High-speed video op amps make ideal video multiplexers, devoid of video distortion or other artifacts.

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**Video multiplexer uses high-speed op amps**

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Any tunable, second-order, active RC-filter section requires at least two thoroughly matched variable resistors. But the lowpass implementation in Figure 1 provides for wide-range cutoff-frequency control using only a single variable resistor, R. In addition to the resistor, this filter comprises an operational amplifier, IC2, which serves as a unity-gain buffer; two capacitors, C1 and C2; and a single-pole, double-throw analog switch, IC1, driven by a periodic sequence of square-wave switching pulses applied to the SW input. Thanks to the high-frequency periodic switching, you can simultaneously control the time constants of both C1 and C2 in their recharging processes using only R. The approximate voltage-transfer function of the filter, assuming that the switching frequency is much higher than the filter's cutoff frequency, is:

\[
H(s) = \frac{1}{\left(\frac{s}{\omega_p}\right)^2 + \left(\frac{s}{\omega_Q}\right) + 1},
\]

where \( \omega_p = 1/(R\sqrt{C_1C_2(1-\Theta)}) \) is the pole frequency; \( Q = \Theta/(1-\Theta)C_1/C_2 \) is the quality factor; \( \Theta = \pi/T \) is the online time ratio (duty cycle); \( \pi \) is the pulse width; and \( T \) is the switching period.

Obviously, controlling \( R \) results only in variations of pole frequency and does not affect the quality factor. So, you can tune this filter over a wide frequency range, preserving its passband gain and ripple. You can achieve a stable value of \( \Theta \) by using a binary counter. A high-resolution, digitally programmable potentiometer is probably the most appropriate choice for \( R \) in this filter. Figure 2 shows the filter's frequency response, simulated in PSpice. This design tunes the cutoff frequency over 20 Hz to 20 kHz by varying the resistor value from 1.2 MΩ to 1.2 kΩ, with \( C_1 = 10 \text{ nF} \), \( C_2 = 1 \text{ nF} \), \( \Theta = 0.5 \), and a switching frequency of 500 kHz. Using this method, you can also implement a high-order lowpass filter by cascading second-order sections or by joining them to multiple-feedback structures.

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**Simple circuit provides precision ADC interface**

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Real-world measurement requires the extraction of weak signals from noisy sources. High common-mode voltages are often present even in differential measurements. The usual approach to this problem is to use an op amp or an instrumentation amplifier and then perform some type of low-pass-filtering to reduce the background noise level. The problems with this traditional approach are that a discrete op-amp circuit has poor common-mode rejection, and its input voltage range is always lower than the power-supply voltage. When you use a differential signal source with an instrumentation-amplifier circuit, using a monolithic IC can provide precision drive and lowpass filtering for an ADC input.
greatly improve common-mode rejection. However, a standard instrumentation amplifier cannot handle sources greater than the power-supply voltage or signals riding on high common-mode voltages. Instrumentation amplifiers using a single external gain resistor also suffer from gain drift. In addition, lowpass filtering requires the use of a separate op amp along with several external components. This approach uses up valuable board space. The circuit of Figure 1 overcomes all of these performance limitations on one μSOIC.

An AD628 precision-gain-block IC is configured as a differential-input amplifier and a two-pole lowpass filter. This circuit can extract weak signals riding on common-mode voltages as high as ±120V. The precision-gain block directly drives an ADC. A separate \( V_{\text{REF}} \) pin is available for offsetting the AD628 output signal so that it is centered in the middle of the ADC’s input range. Although Figure 1 indicates ±15V, the circuit can operate with ±2.25 to ±18V dual supplies. The \( V_{\text{REF}} \) pin can also allow single-supply operation; for this purpose, you simply bias \( V_{\text{REF}} \) at \( V_{\text{supply}}/2 \). The gain block has two internal amplifiers: A1 and A2. Pin 3 connects to ground, thus operating amplifier A1 at a gain of 0.1. The output of A1 directly drives the positive input of amplifier A2.

The first pole of the lowpass filter is a function of the internal 10-kΩ resistor at the output of A1, and an external capacitor, \( C_1 \). The gain of A2 is a function of external resistors \( R_1 \) and \( R_2 \). An external RC time constant in the feedback of A2 creates the second pole. This time constant comprises capacitor \( C_2 \) across resistor \( R_2 \). Note that this second pole provides a more rapid roll-off of frequencies above its RC “corner” frequency (1/(2\( \pi R_2 C_2 \)) than does a single-pole lowpass filter. However, as the input frequency increases, the gain of amplifier A2 eventually drops to unity and does not decrease. So, the ratio of \( R_2/R_1 \) sets the voltage gain of amplifier A2 at frequencies below its −3-dB corner and unity gain at higher frequencies.

Figure 2 is a graph of the filter’s output versus frequency using components to provide a 200-Hz, −3-dB corner frequency. Note the sharp roll-off between the corner frequency and approximately 10 times the corner frequency. Above this point, the second pole starts to become less effective, and the rate of attenuation is close to that of a single-pole response. Tables 1 and 2 provide typical component values for various −3-dB corner frequencies and two full-scale input ranges. The values are rounded off to match standard resistor and capacitor values. Capacitors \( C_1 \) and \( C_2 \) must be high-Q, low-drift units; avoid low-grade disc ceramics. High-quality NP0 ceramic, Mylar, or polyester-film capacitors offer the best drift characteristics and settling time.

### TABLE 1—COMPONENT VALUES FOR 10V P-P FULL-SCALE INPUT FOR A TWO-POLE LOWPASS FILTER

<table>
<thead>
<tr>
<th>Capacitor C1 (µF)</th>
<th>200 Hz</th>
<th>−3-dB corner frequency 1 kHz</th>
<th>5 kHz</th>
<th>10 kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1</td>
<td>0.047</td>
<td>0.002 µF</td>
<td>590 pF</td>
<td>220 pF</td>
</tr>
<tr>
<td>C2</td>
<td>0.01</td>
<td>0.001 µF</td>
<td>600 pF</td>
<td>600 pF</td>
</tr>
</tbody>
</table>

Note: Output is 5V p-p; \( R_1=49.9 \) kΩ, and \( R_2=12.4 \) kΩ.

### TABLE 2—COMPONENT VALUES FOR 20V P-P FULL-SCALE INPUT FOR A TWO-POLE LOWPASS FILTER

<table>
<thead>
<tr>
<th>Capacitor C1 (µF)</th>
<th>200 Hz</th>
<th>−3-dB corner frequency 1 kHz</th>
<th>5 kHz</th>
<th>10 kHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1</td>
<td>0.02</td>
<td>0.0039 µF</td>
<td>820 pF</td>
<td>390 pF</td>
</tr>
<tr>
<td>C2</td>
<td>0.047</td>
<td>0.001 µF</td>
<td>600 pF</td>
<td>600 pF</td>
</tr>
</tbody>
</table>

Note: Output is 5V p-p; \( R_1=24.3 \) kΩ, and \( R_2=16.2 \) kΩ.

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**Buck regulator operates without a dedicated clock**

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Most switching regulators rely on a dedicated clock oscillator to determine the switching frequency of operation. A dedicated oscillator circuit within the power controller usually generates the clock signal. A class of hysteretic switching regulators can actually operate at a relatively fixed frequency without a clock, even with changing input-line and output-loading conditions. Figure 1 shows a simplified buck regulator operating in continuous-conduction mode. (The inductor current always remains positive.) The output voltage, \( V_{\text{OUT}} \), is equal to \( DV_{\text{IN}} \), where D is the duty-cycle ratio of buck switch Q1, and \( V_{\text{IN}} \) is the input voltage. The duty cycle, D, at fixed-frequency operation is \( T_{\text{ON}}/T_s \), where \( T_{\text{ON}} \) is the on-time of \( Q_1 \) and \( T_s \) is the switching-frequency period, \( 1/F_s \). Some rearranging and substitution leads to the expression \( D=V_{\text{OUT}}/V_{\text{IN}}=T_{\text{ON}}/(1/F_s)=T_{\text{ON}}/T_s \).

Now, look at a regulator circuit, which, rather than using a fixed clock and a PWM, uses a circuit that turns on \( Q_1 \) for a time, \( T_{\text{ON}} \), that’s inversely proportional...
to the input voltage, \( V_{IN} \). **Figure 2** shows a regulator based on this principle. This regulator does not contain a clock oscillator, yet it remains at a fixed operating frequency even while the input voltage varies from 14 to 75V. The two main regulation blocks within this regulator are the on-timer and the regulation comparator. The comparator monitors the output voltage. If the output voltage is lower than the target value, the comparator enables the output switch, \( Q_1 \), for a period of time that the on-timer determines. The time period of the on-timer is \( T_{ON} = K R_{ON} / V_{IN} \), where \( K \) is a constant (1.3 \times 10^{-10}) \( R \) is a configuration resistor, and \( V_{IN} \) is the input voltage. If you now substitute \( T_{ON} \) in the previous buck-regulator equations, an interesting result occurs: \( V_{OUT} / V_{IN} = F_S K R_{ON} / V_{IN} \). If you solve for \( F_S \), you obtain \( F_S = V_{OUT} / K R_{ON} \). Because \( V_{OUT} \) remains regulated and the \( K \) and \( R_{ON} \) terms are constants, the switching frequency also remains constant.

The constant-frequency relationship holds true provided that the inductor current remains continuous. At lighter loading, the current in the inductor becomes discontinuous. (The inductor current is zero for some portion of the switching cycle.) At the onset of discontinuous operation, the switching frequency begins to decrease. This reduction is a desirable feature to maintain high efficiency as the load decreases, because switching losses greatly decrease at lower switching frequencies. You derive the switching frequency in discontinuous mode as follows: The peak inductor current \( I_L = V_{IN} T_{ON} / L = V_{IN} K R_{ON} / V_{IN} = K R_{ON} / L \), where \( L \) is the output-inductor value. The output power is \( P_{OUT} = V_{OUT}^2 / R_{OUT} = L I_L^2 / 2 = K^2 R_{ON}^2 / L \). Solving for \( F_S \): \( F_S = (V_{OUT}^2 / L) / (R_{OUT} K^2 R_{ON}^2) \). As you can see, the switching frequency varies inversely with the output resistance, \( R_{OUT} \).

Fixed-frequency operation without an oscillator offers a low-cost, easy-to-implement step-down regulator. You needn’t worry about any loop-compensation or stability issues. The transient response is fast because the circuit has no bandwidth-limiting feedback components. Depending on the inductor value and loading, the operating frequency remains constant for most of the output-power range. A desirable reduction in operating frequency occurs at low loading levels. □

In this buck regulator, the switching frequency remains constant over a wide range of input voltages.
LED driver combines high speed, precision
Richard Cappels, Mesa, AZ

Applications such as turbidity measurement and densitometry require cleanly pulsed light sources with stable amplitudes. The LED driver in Figure 1 illuminates retinal neurons in a biological experiment that has similar requirements. For a given LED at a given current, the intensity is stable, so switching a stable current is a simple and effective way to obtain the needed function. The circuit provides current pulses to the LED with rise and fall times lower than 500 nsec and overshoot lower than 7%. You can make the current computer-programmable by replacing the potentiometer with a DAC. The circuit comprises an adjustable, regulated current source (IC1 and Q2), an overdriven differential amplifier (Q3 and Q4) acting as a switch, and a level shifter (Q1) to shift the TTL input signal to levels needed to drive the differential pair.

Voltage at the wiper of R6 results in an equal voltage across R9 because of feedback to the op amp. Because transistor Q2 has a high alpha, most of the emitter current that produces the voltage across R9 comes from the collector of Q2. Because alpha varies little with temperature, this current remains stable. Transistors Q3 and Q4 constitute a differential pair. Depending on which transistor is conducting, the emitter of one or the other sources current to the collector of Q3. When the base of Q3 becomes several hundred millivolts more positive than the base of Q4, current from Q4 shunts to the 5V power supply. No current flows through Q3, so the LED is off. When the base of Q4 is less positive than that of Q3, current from Q4 passes through the LED. The all-or-nothing switching action results from the large differential voltage across the bases. Similar to the case of Q3, the collector current in Q4, when it is conducting, is a high and stable percentage of the emitter current. The constant load the emitters of Q3 and Q4 present to the current source enables the current source to operate continuously, allowing the use of a low-bandwidth op amp.

Q3 is a common-base amplifier connected in a manner essentially the same as a TTL-input stage with the exception of C4, the 1000-pF capacitor across base resistor R4. When the input signal is greater than 2V, the base of Q3 remains at 2.5V, and the collector of Q3 rises enough to ensure that Q4 and the LED conduct no current. When the input is below 0.4V, Q3’s emitter voltage is low enough and the base current through R4 is high enough to saturate Q3. This action holds the base of Q4 low enough to ensure that all the collector current from Q4 passes through Q4 and the LED. When the input signal swings positive again, the energy stored in C3 develops a reverse bias across Q4’s emitter-base junction to quickly deplete the stored charge, resulting in a rapid turn-off. Make sure that you don’t exceed the power rating of Q4. Take the current and collector-to-emitter voltage into account. Using transistors in TO-92 packages and an LED that drops 2V at 50 mA, the circuit in Figure 1 operates at temperatures greater than 55°C with a jumper in place of R12. If you need higher currents or use smaller transistor packages, you may find it necessary to use a finite resistor for R12 to lower the dissipation in Q4 to a safe level.□

Figure 1

This circuit delivers a stable, precision dose of current to an LED.
The problem: I couldn’t use my Heathkit oscilloscope in a house I lived in during the 1960s because my lab was too far from the power-line input to the house, and the line drop through the house was substantial. Depending on the time of day, the screen would shrink to perhaps half the normal display size. I checked the line voltage, and it was down to just approximately 100V. I lacked the funds to buy a high-wattage Variac to deal with the problem.

The solution: I had a couple of 12.6V filament transformers, rated at 3 or 4A. I simply connected one of these in my lab, with the primary winding across the ac line (Figure 1). Then, I connected the secondary winding such that one side connected to the ac line, and the other side provided the new, boosted ac line. Because the transformer had a center tap, I could adjust the line voltage in 6.3V steps. The beauty of this approach is that the transformer handles only the incremental power from the slight boost in voltage. And the technique uses less space and is less expensive than using a Variac.

Note that, by changing the polarity of the filament transformer’s output, you can decrease rather than increase the ac output. This fact could come in handy in situations in which the line voltage is too high, causing incandescent-lamp burnout. Reduction in lamp life is a function of approximately the 13th power of the overvoltage (Reference 1). For the long, skinny, and expensive European incandescent lamps that some bathrooms use as a vertical light source, the lamp-life reduction can be significant. You can buck, or subtract, the line voltage to increase the lamps’ life. Even at nominal line voltage, you can use the method to drop the voltage to an expensive or particularly inaccessible incandescent lamp.

Reference