Edited by Bill Travis and Anne Watson Swager

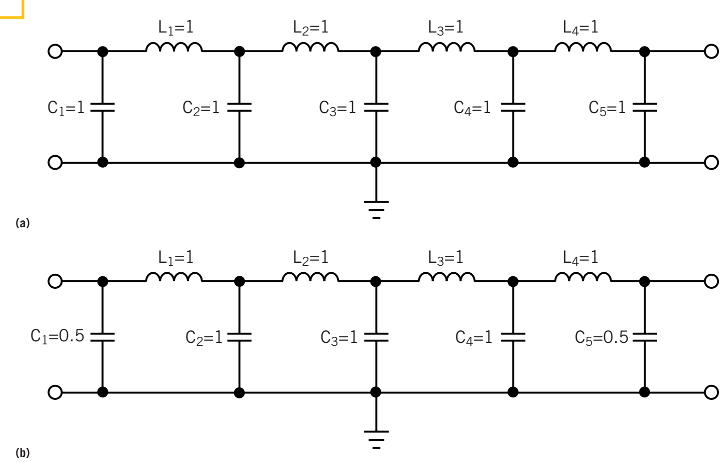
Equal-element filter improves passband performance

Richard M Kurzrok, RMK Consultants, Queens Village, NY

Designers originally conceived equal-element filters as all-pole microwave bandpass filters that provide minimum center-frequency insertion losses for specific values of resonator-unloaded Q (**Reference 1**). All resonators of the equal-element bandpass filter operate at the same loaded Q. For LC filters, the equal-element filter has another advantage. In the lowpass prototype, all inductors have the same value, and all capacitors have the same value. This minimum number of circuit elements provides design simplicity and reduces filter cost.

However, the equal-element filter's response shape has one severe shortcoming. Passband amplitude ripples, due to reflection, are unacceptable for some applications. In minimum-phase-shift filter circuits, group-delay ripples that preclude equalization accompany the amplitude ripples. At microwave frequencies, modifying the central resonator of a five-pole bandpass filter leads to improved performance (**Reference 2**).

Figure 1



The equal-element lowpass filter can feature unacceptable passband ripple for some applications (a). A modified filter realizes substantial improvement in passband performance with some reduction in stopband selectivity (b).

Figure 1a shows the schematic of a nine-pole, equal-element, lowpass-filter prototype. **Figure 1b** shows a comparable schematic of a modified lowpass-filter prototype. By altering the filter input and output capacitors, the modified filter realizes substantial improvement in passband performance with some reduction in stopband selectivity. In the modified equal-element design, all filter inductors are still equal in value, and you need only two capacitor values in a convenient 2-to-1 ratio. **Table 1** shows comparative theoretical amplitude responses for the nine-pole, equal-element lowpass filter and a nine-pole, modified, equal-element, lowpass filter with inductor-unloaded Qs of 100.

The reference frequency, at normalized frequency x=1.0, is not the 3-dB cutoff frequency for the filters in **Figure 1**. The 3-dB cutoff frequency occurs close to x=1.9. This feature differs from Butterworth and Chebyshev filters, for which x can equal 1.0 at 3-dB cutoff frequencies. You use this normalization for equal-element and modified equal-element designs to calculate the values of the circuit elements.

A nine-pole, modified, equal-element lowpass filter was designed at a reference frequency F_R of 4.681 MHz, for which

TABLE 1—THEORETICAL AMPLITUDE RESPONSES FOR NINE-POLE, EQUAL-ELEMENT AND MODIFIED, EQUAL-ELEMENT, LOWPASS FILTER

Normalized frequency	Frequency (MHz)	Insertion loss of equal-element lowpass filter (dB)	Insertion loss of modified equal-element lowpass filter (dB)
0	0	0.172	0.172
0.1	0.468	0.185	0.172
0.2	0.936	0.202	0.174
0.3	1.404	0.201	0.176
0.4	1.872	0.212	0.178
0.5	2.341	0.315	0.182
0.6	2.809	0.532	0.184
0.7	3.277	0.739	0.186
0.8	3.745	0.74	0.192
0.9	4.213	0.475	0.215
1	4.681	0.259	0.27
1.1	5.149	0.707	0.345
1.2	5.617	1.777	0.386
1.3	6.085	2.546	0.337
1.4	6.553	2.167	0.258
1.5	7.022	0.664	0.462
1.6	7.49	1.849	1.191
1.7	7.958	5.52	1.84
1.8	8.426	5.91	1.124
1.9	8.894	3.574	2.887
2	9.362	19.255	12.533
2.1	9.83	28.937	20.678
2.2	10.298	36.285	27.288
2.3	10.766	42.426	32.953
2.4	11.234	47.794	37.977
2.5	11.703	52.612	42.53
2.6	12.171	57.011	46.715
2.7	12.639	61.075	50.604

Note: Data is for inductor-unloaded Qs of 100.

x=1. For 50Ω input and output impedances Z_0, you calculate the normalizing inductance, L_0, and normalizing capacitance, C_0, as follows:

$$L_0 = \frac{Z_0}{2\pi \bullet F_R} = 1.7\ \mu H;$$

$$C_0 = \frac{1\times10^{-6}}{2\pi \bullet F_R Z_0} = 680\ pF.$$

You then use these values of L_0 and C_0 to denormalize the filter to actual circuit-element values. Filter inductors L_1, L_2, L_3, and L_4 are equal to L_0=1.7 μH. Interior filter capacitors C_2, C_3, and C_4 are equal to C_0=680 pF. The filter input and output capacitors, C_1 and C_5, are equal to $0.5\times C_0$=340 pF. In the actual filter, the input and output capacitors are standard 330-pF values. The nine-pole, modified, equal-element, lowpass filter was constructed in a die-cast aluminum box with BNCs. The filter circuit was fabricated using vector board. All capacitors were 5%-tolerance polypropylene units. All inductors used 18 turns of number 26 magnet wire on Micro Metals' T37-2 toroids. **Table 2** shows the measured amplitude-response data. The measured data provides reasonable correlation between theory and experiment and shows substantial improvement in amplitude response over most of the filter passband with some degradation in stopband performance.

TABLE 2—MEASURED AMPLITUDE RESPONSE FOR NINE-POLE, MODIFIED, EQUAL-ELEMENT, LOWPASS FILTER

Frequency (MHz)	Insertion loss (dB)
2	0.1
4	0.15
5	0.2
5.5	0.25
6	0.3
6.5	0.4
7	0.6
7.25	0.85
7.5	1.15
7.75	1.4
8	1.5
8.1	1.1
8.2	0.8
8.3	1.5
8.4	1.1
8.5	0.8
8.6	1.25
8.7	0.75
8.8	1.25
8.9	2.4
9	4
9.25	9.2
9.5	13.7
10	21.9
11	34.6
12 to 30	Greater than 45

You can transform the modified, equal-element, lowpass prototype into useful highpass and bandpass filters with similar design features.

References

1. Taub, JJ, "Design of Minimum Loss Band-Pass Filters," *Microwave Journal*, Volume 6, pg 67, November 1963.
2. Bawer, R, and G Kefalas, "A Modified Equal-Element Band-Pass Filter," *IRE Trans MTT*, Volume MTT-5, pg 175, July 1957.

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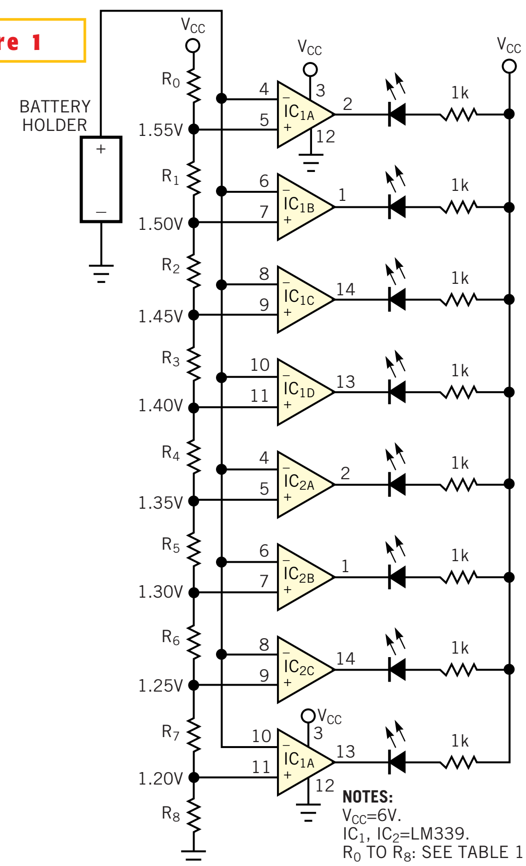
Test batteries without a voltmeter, part 2

Harry Gibbens Jr, PowerStream Technology, Orem, UT

From a high-volume-production point of view of the Design Idea "Test batteries without a voltmeter" (*EDN*, Nov 9, 2000, pg 167), it is a time-consuming and laborious task to tweak the large number of potentiometers on each of the comparator input references. An alternative to this onerous adjustment chore is to replace all potentiometers with 1%-tolerance fixed resistors (**Figure 1**). Before calculating each voltage-reference resistance value, you should use a reasonable selected total resistance value, R_{TOT}, ranging from 100 kΩ to 1 MΩ. You can usually obtain each of the resistance-divider values from off-the-shelf fixed resistors. After calculating all the resistance-divider values (with R_{TOT}= 100 kΩ), if the closest reasonable stock values are unavailable, just increase the value of R_{TOT}. For the example in **Figure 1**, you calculate the values using a spreadsheet to obtain quick results. This example uses R_{TOT}=182 kΩ. All the calculations use Kirchoff's law:

Figure 1



Fixed, 1%-tolerance resistors eliminate the need to trim potentiometers in this battery-testing circuit.

$$V_X = V_{TOT}\left(\frac{R_X}{R_1+R_2}\right).$$

Rearranging the terms in the formula, you determine the resistance value, R_X:

$$R_X = R_{TOT}\left(\frac{V_X}{V_{TOT}}\right),$$

where V_X is each comparator's reference voltage, V_{REFX}; V_{TOT}=V_{CC}; and R_{TOT} is the manually selected value. The first step is to calculate the "Formulated R" in **Table 1**, using the rearranged formula. An example follows:

$$R_{FORM1} = 182\text{ k}\Omega\left(\frac{1.55\text{V}}{6\text{V}}\right)$$
$$= 182\text{ k}\Omega(0.2583)$$
$$= 47.01\text{ k}\Omega.$$

Next, calculate the first resistance value, R_0, by using the formula

$$R_X = R_{TOT} - R_{TOT}\left(\frac{V_X}{V_{TOT}}\right) = R_{TOT} - R_{FORMX}$$

$$R_0 = 182\text{ k}\Omega - 182\text{ k}\Omega\left(\frac{1.55\text{V}}{6\text{V}}\right)$$
$$= 182\text{ k}\Omega - 182\text{ k}\Omega(0.2583)$$
$$= 182\text{ k}\Omega - 47.01\text{ k}\Omega$$
$$= 134.99\text{ k}\Omega.$$

Calculate the rest of the resistor values, R_1 through R_7, by using the formula

$$R_{X+1} = R_{TOT}\left(\frac{V_X}{V_{TOT}}\right) - R_{TOT}\left(\frac{V_{X+1}}{V_{TOT}}\right) =$$
$$R_{FORMX} - R_{FORMX+1}.$$

For example, for R_1:

$$R_1 = 100\text{ k}\Omega\left(\frac{1.55\text{V}}{6\text{V}}\right) - 100\text{ k}\Omega\left(\frac{1.5\text{V}}{6\text{V}}\right)$$
$$= 1.51\text{ k}\Omega.$$

Finally, use the rearranged formula to

TABLE 1—CALCULATED VALUES FOR BATTERY TESTER

Reference voltage	Voltage (V)		Formulated R kΩ	RX	Calculated R kΩ	Fixed R kΩ
V_{TOT}	6			R_{TOT}	182	
V_{REF1}	1.55	0.26	47.01	R_0	134.99	134.5
V_{REF2}	1.5	0.25	45.5	R_1	1.51	1.5
V_{REF3}	1.45	0.24	43.99	R_2	1.51	1.5
V_{REF4}	1.4	0.23	42.46	R_3	1.53	1.54
V_{REF5}	1.35	0.23	40.95	R_4	1.51	1.5
V_{REF6}	1.3	0.22	39.44	R_5	1.51	1.5
V_{REF7}	1.25	0.21	37.91	R_6	1.53	1.54
V_{REF8}	1.2	0.2	36.4	R_7	1.51	1.5
				R_8	36.4	36.4

find the last resistance value, R_8.

$$R_8 = 182\ k\Omega\left(\frac{1.2V}{6V}\right) = R_{FORM8}$$
$$= 36.4\ k\Omega.$$

In **Table 1**, you match the rounded-off calculated resistance value (Calculated R) to the nearest available fixed-resistor value (Fixed R) as shown in the last two columns. Note that R_0 consists of two series-connected resistors. You can download the table of 1%-tolerance resistor values from *EDN*'s Web site, www.ednmag.com. Click on "Search Databases" and then enter the Software Center to download the .gif file for Design Idea #2654.

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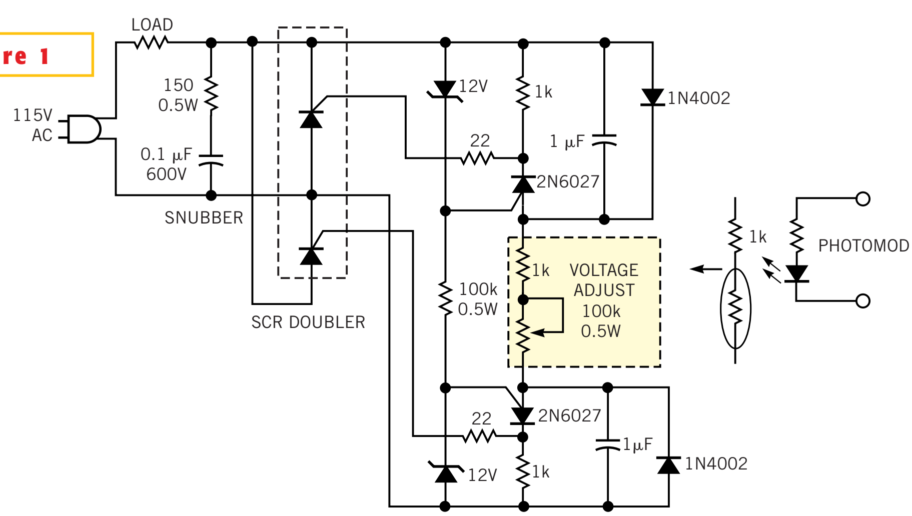
SCR phase control yields solid-state switch

James Keith, York, PA

SCRs (silicon-controlled rectifiers), or thyristors, have higher current and voltage ratings, lower conduction losses, and more robustness than triacs. For these reasons, SCRs are better suited to high-power applications. For example, you can use two SCRs to configure a 100 or 200A, 460V control circuit. **Table 1** lists some of the advantages of SCRs over triacs. The main challenge is driving the SCRs: You now have two gates, rather than one, to drive. Furthermore, the gates are referenced to opposite polarities with a significant voltage difference. The circuit in **Figure 1** solves the problem with two PUTs (programmable unijunction transistors), one connected to each SCR. Performance is good because the circuit does not "fold back" as inexpensive triac phase controls (dimmers) do. The PUT fires when the timing-capacitor voltage exceeds the PUT reference voltage by one diode drop and dumps the capacitor's charge into the SCR gate.

You adjust the phase delay by varying the charge rate of the capacitor via the potentiometer. You can enhance the balance by matching both the zener-diode voltages and the values of the charging capacitance. You can obtain 230V operation by using 1.5- to 2-µF timing capacitors. You can achieve 460V-ac operation by using 1200V SCRs; 3-µF timing capacitors; and a 100-kΩ, 25W potentiometer. Using the circuit as a solid-state switch is practical; simply replace the voltage-adjust block in the broken lines in **Figure 1** with a *photomod*. A photomod is a unique type of photocoupler that has an LED or incandescent-light source and a cadmium-sulphide photocell. Unique properties of these photocells are their high dark resistance and high voltage ratings.

Figure 1



This SCR phase-control circuit uses one potentiometer and no pulse transformers.

TABLE 1—SCR-VERSUS-TRIAC COMPARISON

Device	SCR	Triac
Current rating	>>50A	<40A
Conduction voltage drop	~1V	~1.5V
Junction temperature	125°C	115°C
Thermal resistance	Low	Medium
Maxium voltage	1400V	600V
Surge current	High	Limited
Robustness	High	Low
Package	Doubler or "hockey puck"	¼-in. stud
Isolation	Isolated	Nonisolated
Number required	Two	One

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Circuit facilitates video fading

JM Terrade, Clermont-Ferrand, France

When you're copying videotapes, it's sometimes desirable to suppress some passages. Using the pause control of the recorder does not yield satisfactory results. Another method produces better results (**Figure 1**). The video source connects to the video-in plug, and the recorder connects to the video-out plug. Turning potentiometer P_1 adjusts the image brightness from normal video to a black image. With the P_2 potentiometer ganged to P_1, the sound also varies accordingly. The objectives in building this circuit are to use inexpensive, readily available components and to obtain batteryless operation. The video signal follows two paths (**Figure 2**). In the first path, the signal undergoes amplification by a factor of two and connects to one end of a potentiometer. In the second path, the synchronization pulse, separated from the input signal, connects to the other end of the potentiometer. The wiper of the potentiometer connects to the second video amplifier, which provides the video output.

When you adjust P_1's wiper from one end to the other, the video image disappears and fades to a black screen. Because P_1 and P_2 are ganged, the sound follows the image brightness. The circuit could have used triple integrated video amplifiers, such as an AD813, and a video sync separator, such as an LM1881. However, these ICs are expensive (approximately $25) compared with the six standard transistors shown in **Figure 2**. R_1 sets the input impedance at 75Ω. Q_1, Q_2, and associated components form a video amplifier with an approximate gain of two. R_2, R_3, and D_1 set the dc voltage, and C_1 blocks any dc voltage from the source. The amplified video signal connects

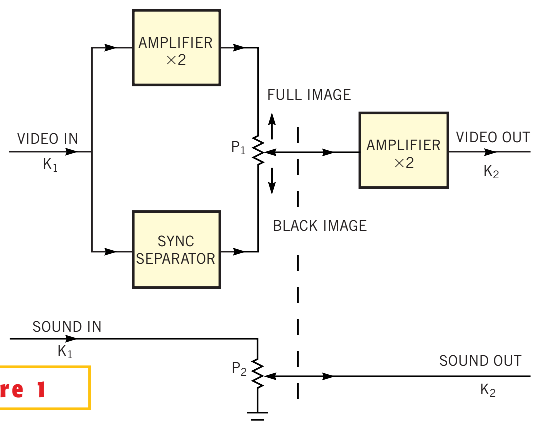
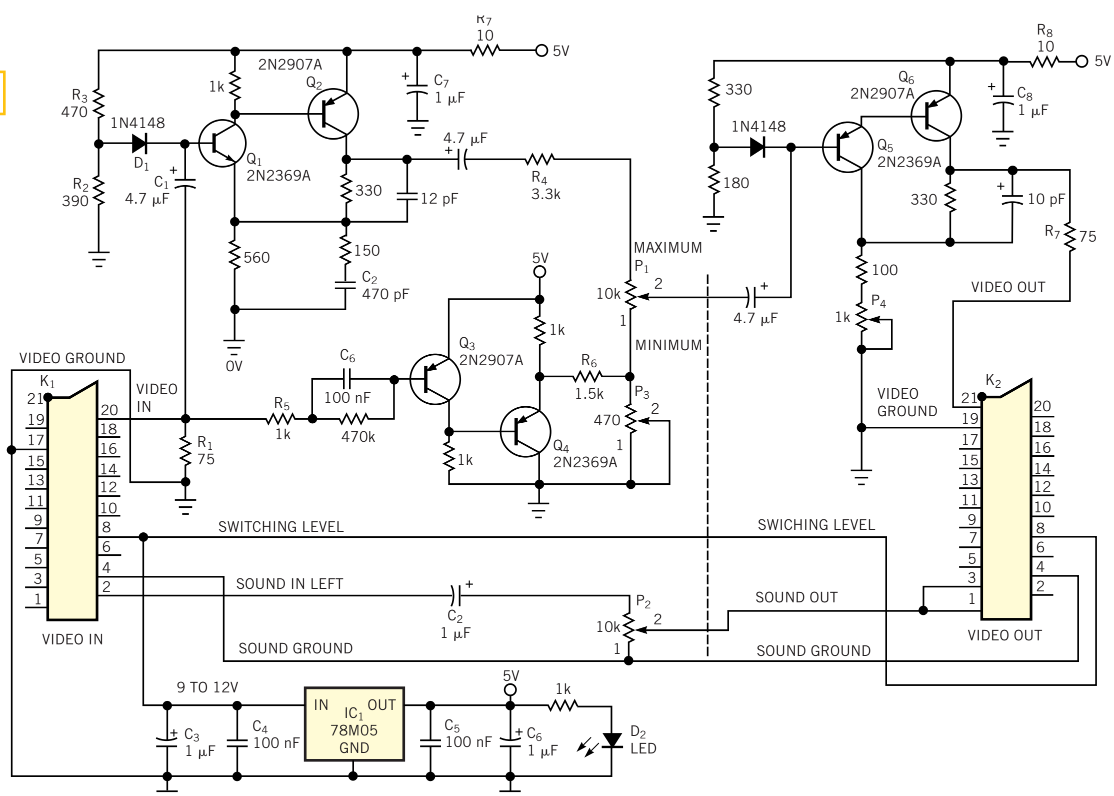


Figure 1

A simple circuit provides for effective video and audio fading when you're recording source material.

Figure 2



A handful of transistors and associated components yields a professional-quality video fader.

to P_1 through R_4 and the C_1 dc-blocking capacitor. R_5, Q_3, Q_4, and associated components form a sync separator. The sync pulse connects to the bottom of P_1 through R_6 and P_3, an adjustable voltage divider. The wiper of P_1 connects to the second video amplifier comprising Q_5, Q_6, and associated components. You can adjust the amplification with P_4. R_7 sets the output impedance at 75Ω.

P_2, ganged to P_1, is a simple voltage divider, using C_2 to block dc voltages. The sound input uses the left channel, and the output goes to both the left and right channels. With a video source connected to video in, a dc voltage of 9 to 12V appears at Pin 8 of the video plug. IC_1 and the C_2 through C_6 capacitors derive power from this pin and provide a stable 5V for the circuit. D_2 is a high-brightness LED that indicates that a video source is present. R_7, C_7, R_8, and C_8 provide decoupled supplies for the amplifiers. Video cables are often of poor quality. For that reason, the circuit in **Figure 2** provides for amplification of the video signal. Also, compensation of the first amplifier provides amplification of color burst with a concomitant improvement of video quality. To adjust the circuit, first turn P_1 fully clockwise and then adjust P_4 for a good video image. Then, turn P_1 fully counterclockwise and adjust P_3 to obtain a stable black image.

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Time-delay relay uses proximity control

Dennis Eichenberg, Parma Heights, OH

You can build a circuit that allows a passerby to briefly operate model trains in a display window (**Figure 1**). The design uses a proximity detector rather than a pushbutton switch to eliminate the need to mount and wire any equipment outdoors. The circuit worked well in this application and other applications. The heart of the circuit is the quad CMOS NAND gate, IC_1. A human hand near the sensor induces 60-Hz power-line noise into IC_{1A}, and this IC triggers IC_2. IC_2 is configured as a monostable multivibrator, with a period equal to $1.1(R_3+R_4)(C_3)$. The period is adjustable from approximately 0.5 to 50 sec. The sensitivity of the circuit depends on the size of the sensor. A 10-in.-sq piece of screen mounted inside the display window works well for this application, because it permits complete visibility through the sensor. The circuit triggers several inches away from the window in this application.

You can manually or automatically operate the circuit through the single-pole, double-throw, center-off switch, S_1. In automatic operation, receptacle J_1 is active when IC_2 drives relay K_1 on. Lamp LP_1 indicates when the load is active. Lamp LP_2 indicates when the circuit is in automatic mode. J_1 is active whenever S_1 is in the on position, as LP_1 indicates. V_{CC} for the circuit comes from transformer T_1, rectifier D_1, and filter capacitors C_1 and C_2. The load rating for the circuit depends on the selection of fuse F_1, switch S_1, relay K_1, and receptacle J_1.

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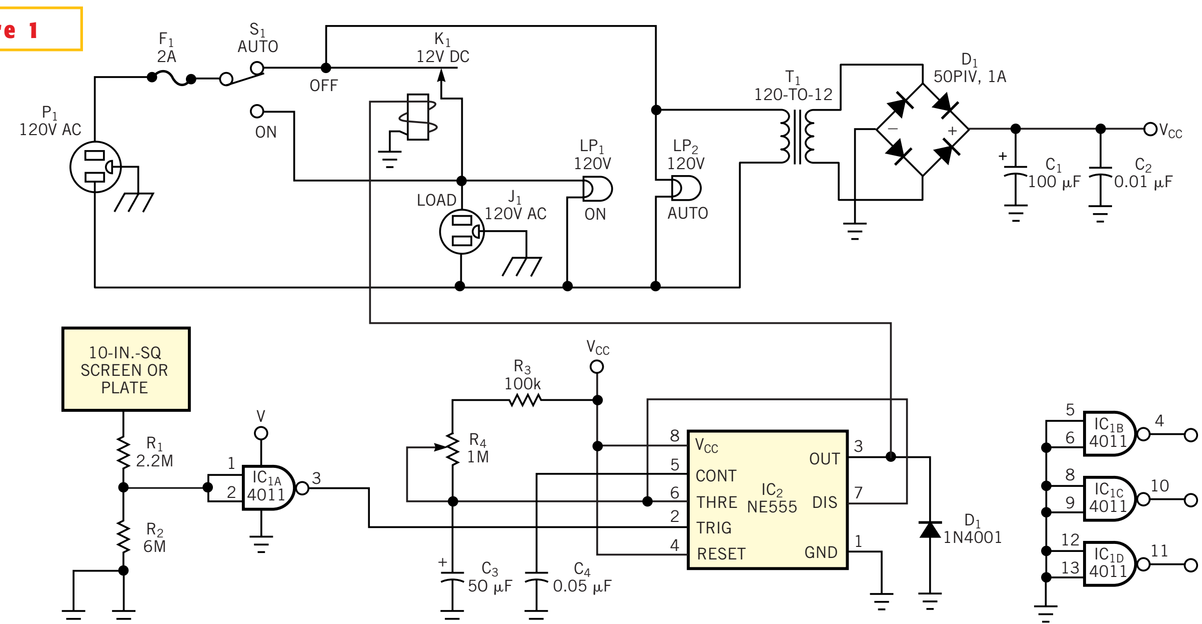
Figure 1



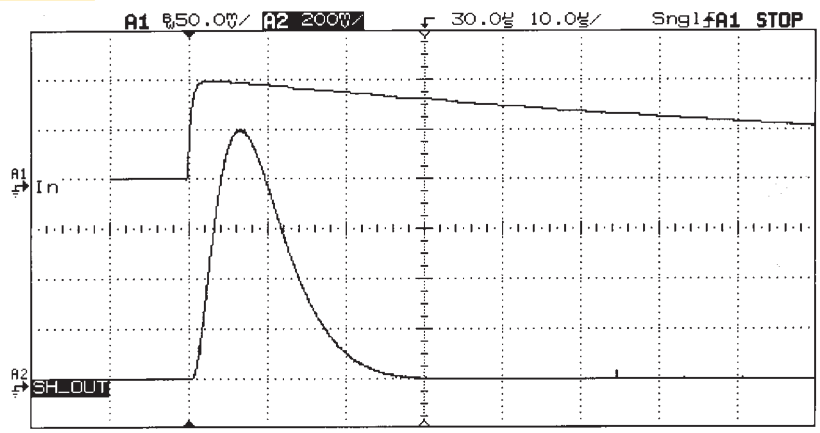
A proximity sensor turns a load on when a human hand comes near the sensor screen.

Circuit forms shaping amp and amplitude detector

Elio Rossi, ITESRE-CNR, Bologna, Italy

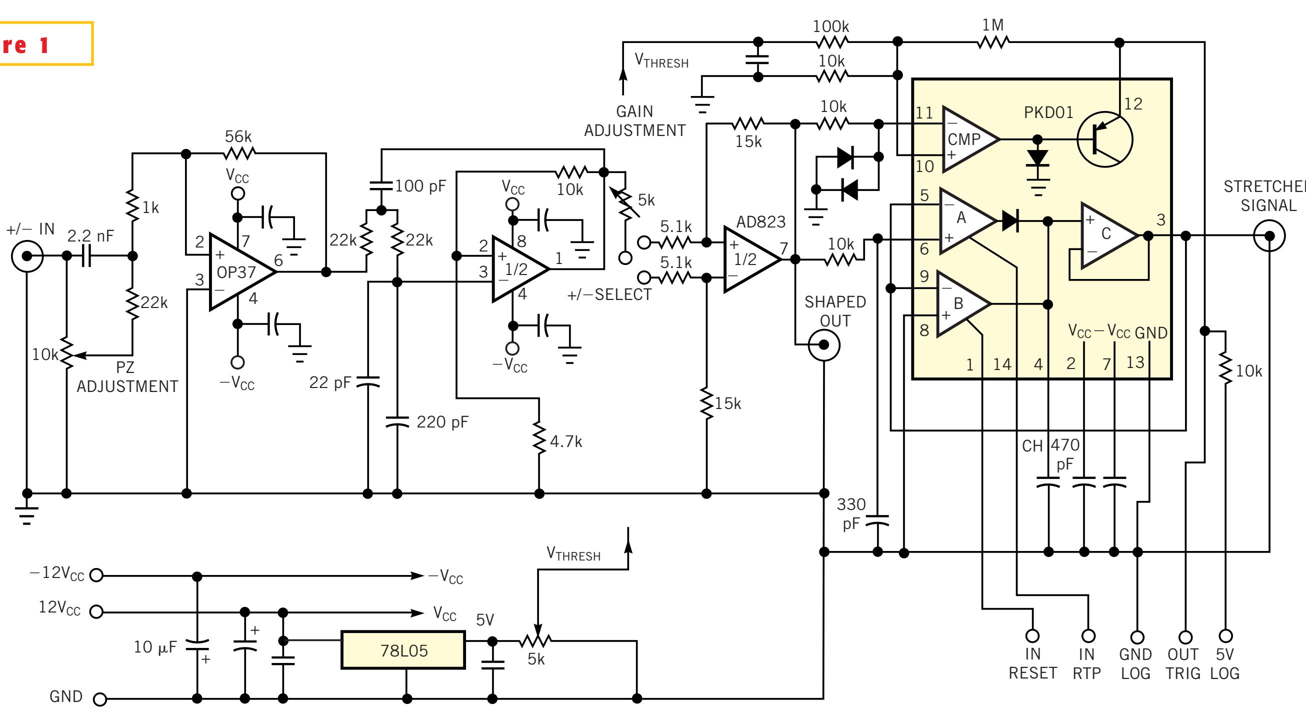
The use of solid-state detectors connected to charge amplifiers requires appropriate conditioning of the output signals, because of the signals' long decay time. Moreover, you must "stretch" the peaks of the shaped pulses for a period sufficient for A/D conversion. For a single detector, you can use a relatively expensive module. For a large detector array, you need to develop an ASIC. For a moderate number of detectors, you can use an inexpensive circuit that handles an array of 19 drift diodes connected to a scintillating crystal of cesium-iodide (**Figure 1**). The two ICs, an OP37 and an AD823, provide the correct gain and the semi-gaussian conditioning of the input pulses. The conditioning involves a differentiating input with pole-zero adjust in the gain-of-35 first stage, a lowpass Sallen-

Figure 2



At the AD823 output in Figure 1, the signal peaks in 6 msec. A peak detector then "stretches" the signal.

Figure 1

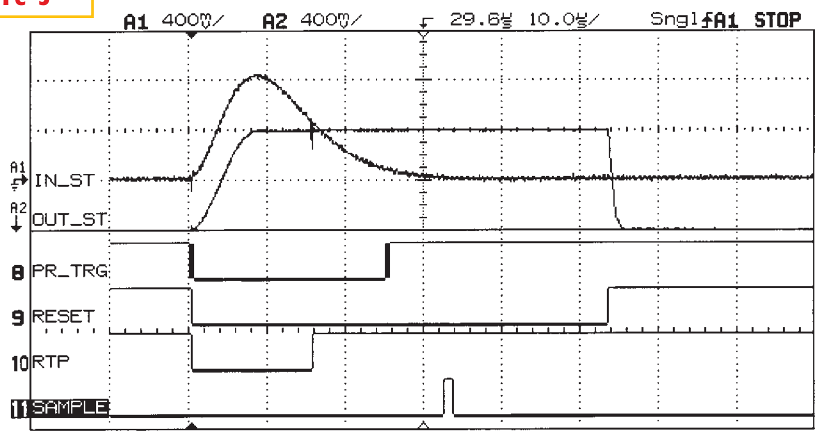


A few linear ICs form the basis of a shaping amplifier and peak-amplitude detector.

Key filter in the unity-gain second stage, and a gain-adjustable third stage. **Figure 2** shows a peaking time of 6 μsec with a gain of 50 to 100V/V. The circuit can use either a positive or a negative input signal.

The third IC, a PKD01, acts as a "stretcher" circuit with a built-in trigger discriminator. To measure the performance of the circuit, the design uses four external one-shot multivibrators, a precision pulse generator, and a multichannel analyzer in sample mode. The discriminator output triggers the sequence of the one-shots, which in turn open the in reset and close the in rise-time-protection, gated amplifiers, A and B (**figures 1** and **3**). This action allows the hold capacitor to reach the peak of the integrated (via the 10-kΩ, 330-pF network) input pulse, so that an A/D converter can begin its conversion. The circuit must maintain the stretched signal via the reset pulse until the end of conversion. Input integration is necessary to generate a delay between the aperture of the reset command and the peak of the input pulses to maintain the linearity of the low-level signal. Also, the approximately 0.7V/μsec slew rate of the stretched output requires at least 8 μsec in the peaking time of the input signal to obtain a stretched signal output greater than 5V.

Figure 3



In the second waveform down, the peak detector provides a sample/hold function and preserves the peak value of the signal.

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