

Edited by Bill Travis and Anne Watson Swager

## Synchronous oscillator converts audio, video to FM

Vasil Uzunoglu, Synchtrack, Gaithersburg, MD

**T**HE SYNCHRONOUS OSCILLATOR (SO) and the coherent phase-locked synchronous oscillator (CPSO) are universal multifunctional networks that track, synchronize, and amplify as much as 80 dB; improve SNR by as much as 70 dB; and modulate AM, FM, and FSK signals. You can also use these networks as ADCs, sampling networks, and dividers that divide by rational integer numbers, such as 3/4, 5/7, and 7/8. Suitable applications include wideband spread-spectrum communications and binary-phase-shift-keying (BPSK) and quadrature-phase-shift-keying (QPSK) generation. A CPSO retains all the properties of the SO and provides zero phase error.

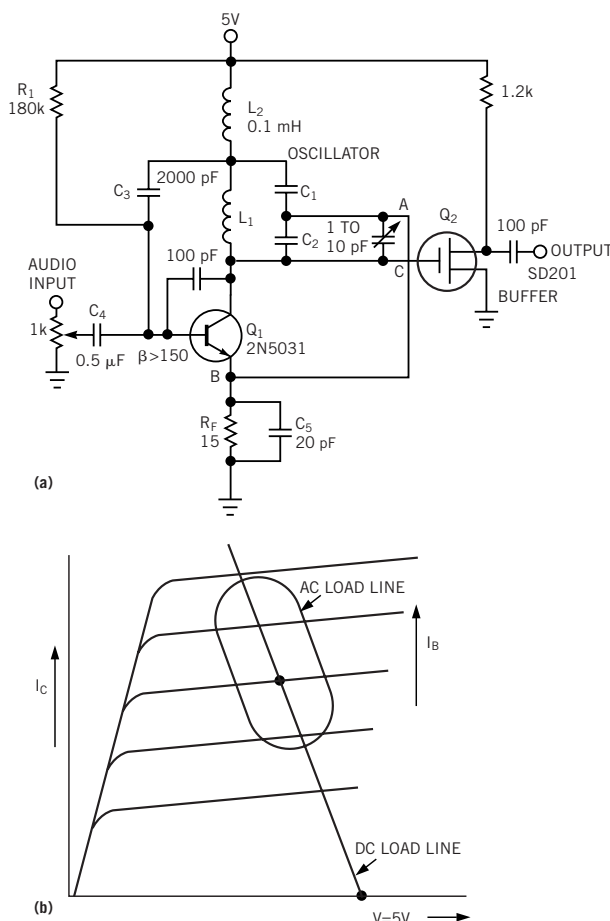
One SO application area is the conversion of audio and video to FM in a single process. **Figure 1a** shows a simple, one-stage SO oscillating at 94 MHz. This frequency is a good choice for testing the audio-to-FM conversion on an FM broadcast receiver. The audio or video input to this SO should not exceed -5 dBm; any signal above this level may induce amplitude modulation.  $C_4$ 's decoupling capacitor passes the audio or video to the SO and eliminates dc bias at the in-

put.  $R_1$  biases the SO. Because the oscillator's load is a combination of resistance, inductance, and capacitance, the load line is a combination of a straight line and an ellipse (**Figure 1b**). The linear load line indicates the dc bias, and you must locate the load line away from the nonlinear characteristics of the transistor.

$C_3$  and the connection between points A and B each provide positive feedback. High positive feedback is essential to the

optimum operation of the SO. The value of  $C_3$  should range from 2000 to 5000 pF.  $R_F$  in parallel with  $C_5$  has numerous functions. The presence of this network allows positive feedback within the SO, but looking from the input,  $R_F$  and  $C_5$  also provide negative feedback. This negative feedback adds frequency stability. Finally,  $R_F$  and  $C_5$  divert the input audio or video to the oscillator, not to ground.  $R_F$  and  $C_5$  are approximately  $15\Omega$  and 20

**Figure 1**



**A single-stage synchronous oscillator converts audio or video to FM (a). The load line is a combination of a straight line and an ellipse (b).**

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pF, respectively, and these values depend on the  $\beta$  of the transistor.  $L_2$  is an RF choke that diverts the collector feedback to the base of the transistor. You should tune the tank circuit, which comprises  $L_1$ ,  $C_1$ , and  $C_2$ , to approximately 95 MHz to be within the FM broadcast band. A 2N5031 for  $Q_1$  works well for this application because it has high gain and low noise; any other transistor with the same characteristics is suitable. A 2- or 3-in. long wire attaches to the output and acts as an antenna.

The conversion from audio or video to FM takes place in the internal base-emitter junction capacitor by its parametric

action, and the changes of the junction capacitor due to audio-to-video variations modulates the oscillations to provide an audio- or video-to-FM conversion.

The SO also has high input-signal sensitivity and high noise rejection. It can detect signals as low as  $-100$  dBm and signals with SNRs as low as  $-40$  dB. For a PLL, these performance numbers are  $-25$  dBm and 3 dB, respectively. The SO is frequency-stable and has low phase jitter because the tank circuit has a high Q that can reach  $3 \times 10^6$ . The SO has three independent internal filters. With a 200-Hz noise-rejection filter, the SO exhibits

a data bandwidth or tracking range of several megahertz. For input signals with high noise, a two-transistor cascoded SO is preferable.

The output buffer,  $Q_2$ , provides protection between the oscillator and the output world. However, if you want to simplify the circuit, you can remove  $Q_2$  and its 1.2-k $\Omega$  pullup resistor and connect the 100-pF capacitor at the output directly to Point C. (DI #2379)

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## High-isolation converters use off-the-shelf magnetics

Mitchell Lee, Linear Technology Corp, Milpitas, CA

ISOLATED FLYBACK CONVERTERS usually evoke thoughts—or bitter memories—of custom transformers, slipped delivery schedules, and agency-approval problems. Off-the-shelf flyback trans-

formers carry isolation ratings of only 300 to 500V and rarely of as much as 1 kV. Gate-drive transformers are readily available from stock with high isolation ratings and low cost, but they are wound

on ungapped cores, have high inductance (500  $\mu$ H to 2 mH), and quickly saturate in a normal flyback-converter circuit. Thus, high isolation calls for an abnormal flyback converter (Figure 1).

Based on the uncoupled SEPIC (single-ended primary-inductance-converter) topology, the converter operates from a 12V battery-backed input supply and outputs 24V at 200 mA. The key feature is the second “coil,” which is not a coil but rather an off-the-shelf gate-drive transformer,  $T_1$ . This component offers 3750V rms isolation and full VDE approval; it functions flawlessly in SEPIC service. The output is completely isolated from the input.

The converter derives feedback from the primary winding through  $D_1$ . The transformer winding is 1-to-1.  $C_1$  peak detects a voltage roughly equal to the output. A minimum load of 3.6 k $\Omega$  prevents the output from rising uncontrollably at zero load (Figure 1). (DI #2380)

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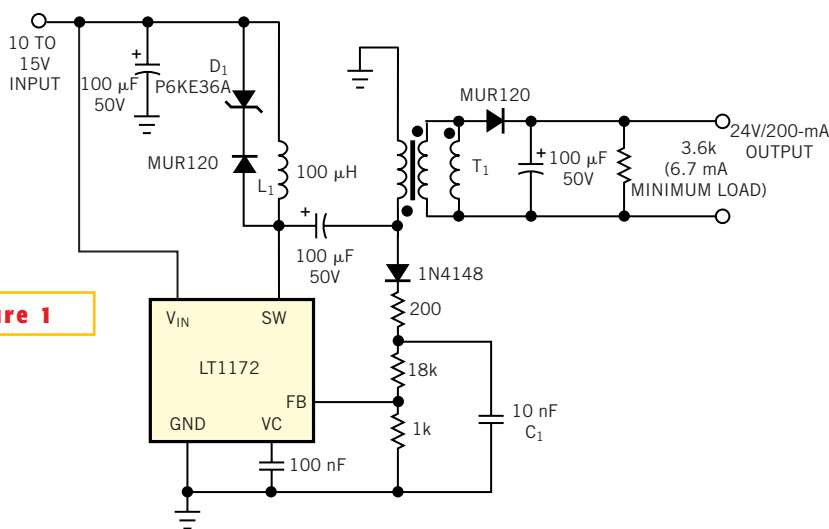


Figure 1

NOTES:  $L_1$ =PE53829 (PULSE ENGINEERING, WWW.PULSEENG.COM).  
 $T_1$ =PE-63387.

The second “coil” of this unusual flyback converter is not a coil but rather an off-the-shelf gate-drive transformer. This component offers 3750V rms isolation and full VDE approval.

# Autoranging circuit simplifies hardware and software

Dana Romero, Phonex, Midvale, UT

**A** NATURAL APPROACH TO autoranging is to use an inverting op amp and switch in one of N feedback resistors at a time for a gain of  $A = -R_F/R_O$  (Figure 1). However, the inverting configuration suffers from low input impedance of  $Z_{IN} = R_O$  because the negative input is at virtual ground. Also, another inverter is necessary to provide positive voltage ranges because of positive signals at the input.

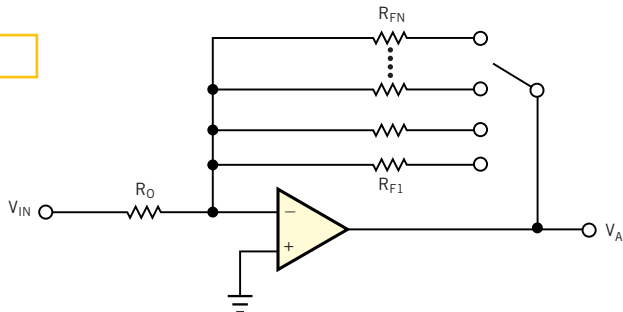
An alternative approach is to use a noninverting op amp that has a high input impedance and a gain equation of  $A = 1 + R_F/R_O$  (Figure 2). This circuit augments the typical 0 to 5V input range of a  $\mu C$  with ranges of 0 to 1V and 0 to 10V. The circuit in Figure 2 uses a 68HC16  $\mu C$ , but the results generally apply to any  $\mu C$  with an ADC. The unusual arrangement of circuit components provides three input ranges that use only a single op amp, two spdt relays, and resistors with standard values. Also, the circuit simplifies the autoranging software because after  $K_2$  switches, the status of  $K_1$  is irrelevant.

With the relays in the positions that Figure 2 indicates,  $R_F$  equals  $R_1$ , which in turn equals  $R_O$ . Thus, the amplifier gain is 2. Combined with the voltage divider of  $R_4/R_5$ , the overall circuit gain is 1/2, which is the gain necessary for a 0 to 10V input at  $V_{IN}$  to provide 0 to 5V at  $V_{AD}$ .

When  $K_1$  switches,  $R_F$  becomes  $R_2 + R_1 = 2R_O + R_O = 3R_O$  for a gain of  $1 + 3R_O/R_O = 4$ . Dividing by 4 gives an overall gain of 1, which is suitable for a 0 to 5V range at  $V_{IN}$ . At this point, it becomes obvious why the voltage divider sits before the amplifier: With  $V_{IN}$  near 5V, multiplying by 4 would saturate the op amp.

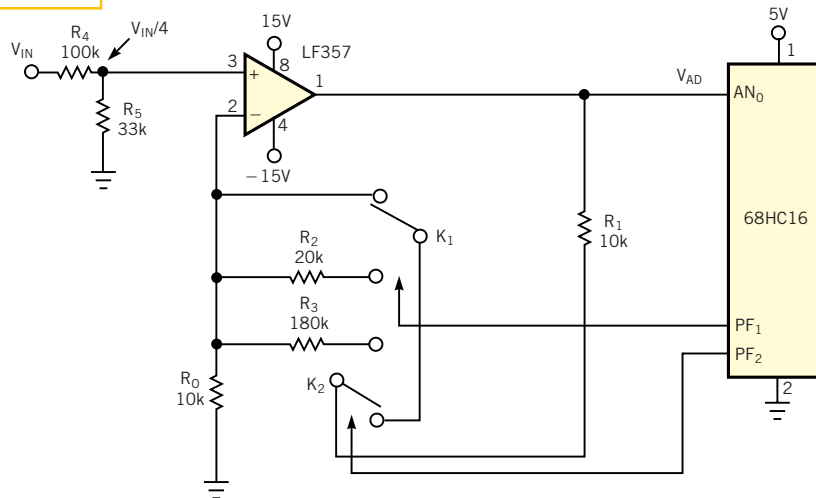
When  $K_2$  switches, the position of  $K_1$  is irrelevant, and  $R_F$  equals  $R_3 + R_1 = 18R_O + R_O = 19R_O$ . Thus,  $A = 1 + 19R_O/R_O = 20$ . Now, the overall gain is  $20/4 = 5$ , or exactly the value necessary for a  $V_{IN}$  range of

**Figure 1**



A single noninverting op amp, two spdt relays, and standard-value resistors can produce a 0 to 5V output from input ranges of 0 to 10V, 0 to 5V, and 0 to 1V.

**Figure 2**



An obvious autoranging circuit suffers from low input impedance and needs an inverter to produce positive output-voltage ranges.

0 to 1V to produce an output of 0 to 5V.

Although the gain calculation works out exactly with standard 5% resistor values, for greater accuracy, you can switch the resistors to the nearest 1% values. Or, even better, sort through a batch of 5% resistors and use a good multimeter to

find resistors that are within 1% of nominal. (DI #2381)

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# Simple technique speeds Microstrip breadboarding

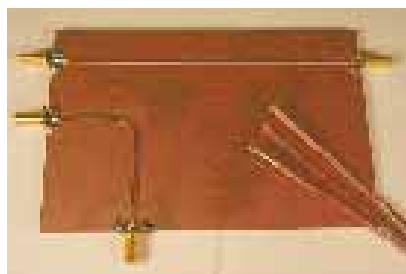
Steve Hageman, Hewlett-Packard Co, Santa Rosa, CA

**S**UPERGLUE AND SOME THIN, sheared pieces of Teflon copper-clad material are all you need to easily produce Microstrip-type structures for breadboarding high-frequency analog or RF circuits. By gluing properly sheared widths of copper-material to the pc-board material, you can build an RF-circuit prototype in less than one-tenth of the time it takes to carve and clean a Microstrip on the top surface of a prototype pc board. Also, this method allows you to easily move or add Microstrip lines as prototyping progresses.

High-frequency prototyping of analog circuits demands that you build everything on a ground plane to limit the inductance in the ground returns. Furthermore, RF-circuit design demands that all interconnections be  $50\Omega$  transmission lines, or  $75\Omega$  transmission lines if you are a TV type. The most common way to make a  $50\Omega$  transmission line on a pc board is to use a Microstrip structure. A Microstrip transmission line is a trace on the top of the board, and the ground return is the backside copper of the board. This type of structure is easy to make when you are laying out a board for production, but it presents formidable problems when you are prototyping. For one thing, it is inconvenient to use the backside of the prototype pc board as the ground plane. For every node that must connect to ground, you must drill a via through the board to reach the ground. Also, making the Microstrip line requires that you use a razor or power tool to carve out the top-side copper. This technique is messy and error-prone, especially if you are building the circuit piece by piece.

The sheared-copper method exhibits the necessary  $50\Omega$ -impedance characteristics and eliminates the ground-plane-attachment problems for the other components. The key is to get your hands on a 0.026-in.-thick Teflon pc-board material called RT/duroid 5870 (Rogers Corp, [www.rogers-corp.com](http://www.rogers-corp.com)).

This material has a relative dielectric constant ( $\epsilon_r$ ) of about 2.33. Using any electrical-engineering text, you can find equations that state that for an  $\epsilon_r$  of 2.33, the width-to-height ratio of a Microstrip structure is about 3.3 to 1 for a  $50\Omega$  impedance. So, when using 0.026-in.-thick pc-board material, you should shear the



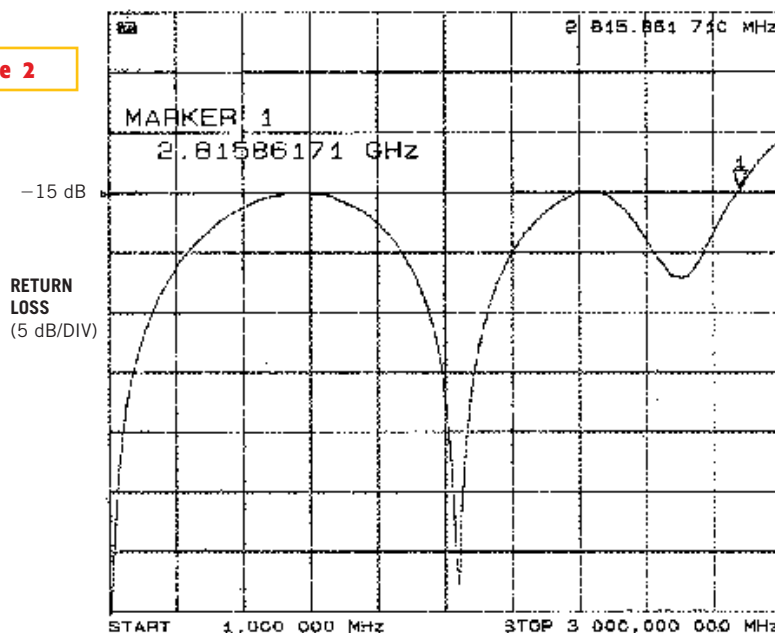
**Figure 1** Gluing strips of RT/duroid material on top of normal FR4 pc-board material makes a very quick Microstrip structure. Make sure that the line lies flat and is firmly glued in place against the board to maximize the return loss of the line.

RT/duroid board in approximately 0.086-in.-wide strips.

You can attach the strips to the copper-clad prototype board using dabs of cyanoacrylate, or instant, adhesive along the length of the strips to form the Microstrip structure. The RT/duroid material is soft and makes it easy to place and bend the strips as necessary; it exhibits an excellent  $50\Omega$  match to around 3 GHz or higher. Also, you can easily cut the resulting Microstrip structure to allow for placement of series-matching or passive-lumped elements for dc decoupling. This method also makes it easy to place parallel components and to place components from the Microstrip line to the ground plane; you simply solder from the top of the Microstrip line to the adjacent ground plane without drilling vias.

For measurement purposes, the photo in Figure 1 shows two Microstrip lines on a prototype pc board. You glue the upper line to the prototype board to form a straight transmission line. The lower

**Figure 2**



A return-loss (S11) measurement of the bent Microstrip-line in Figure 1 exhibits an excellent match— $-15$  dB or VSWR $<1.4$  to 1—to nearly 3 GHz.

line, which easily bends around corners, is bent 90°. Measurements of the bent line show that the return loss of the line exceeds 15 dB to almost 3 GHz and that the straight Microstrip line has a better than 20-dB return loss all the way to 3 GHz (Figure 2).

Alternative methods of breadboarding yield poorer results. The use of common

wire-wrap wire placed above a ground plane yields an impedance of about 100Ω and is usable only to the 100-MHz region for short lengths. The use of miniature coax is not only time consuming, but the shield pigtailed at the ends of the coax, which are necessary to access the center conductor, add enough inductance to cause power holes of 5 to 10 dB in the

range of 100 MHz to 3 GHz. Also, both of these methods preclude the easy attachment of components anywhere along the transmission line. (DI #2382)

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## Comparator has independent trip voltages

Gregory Billiard, Fike Corp, Blue Springs, MO

**F**OR AN INVERTING AMPLIFIER with hysteresis, resistors  $R_A$ ,  $R_B$ , and  $R_F$  determine the crossover voltages (Figure 1). Unfortunately, using three resistors to set the upper and lower trip voltages creates a dependence between the two trip voltages: It's impossible to set one voltage without affecting the other. However, you can achieve the same output-voltage response with two independent trip voltages by using a comparator with an open-collector output, such as the LM393 or LM339. In Figure 2, the resistor divider of  $R_1$ ,  $R_2$ , and  $R_3$  determines the upper trip voltage,  $V_1$ .  $R_4$

and  $R_5$  determine the lower trip voltage,  $V_2$ .

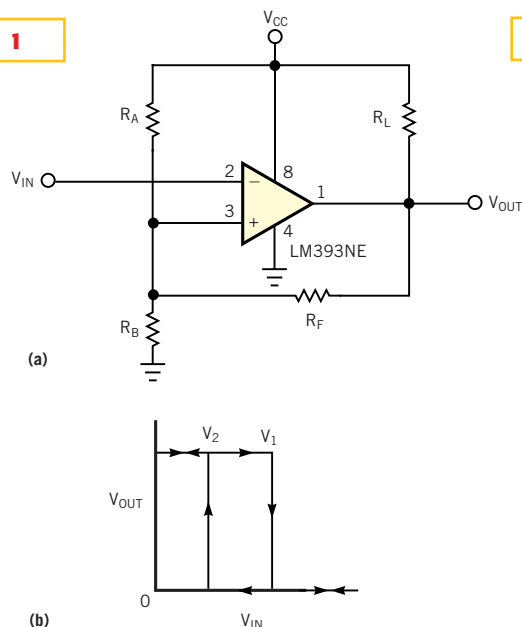
When  $V_{IN}$  is between 0 and  $V_1$ ,  $V_{OUT}$  is high and prohibits any current flowing through  $R_4$  and  $R_5$ , which sets  $V_2$  to  $V_{CC}$  and thus keeps the output of Comparator B high. When the output of Comparator B is high,  $V_1 = V_{CC} \times R_3 / (R_1 + R_2 + R_3)$ . When  $V_{IN}$  exceeds  $V_1$ ,  $V_{OUT}$  goes low, to 0.7V, allowing current to flow through the resistor divider.  $V_2$  then changes from  $V_{CC}$  to  $0.7V + V_{CC} \times R_5 / (R_4 + R_5)$ . On this change, the output of Comparator B also goes low, bringing  $V_1$  to  $0.7 \times R_3 / (R_2 + R_3)$ .  $V_1$  is now lower than

$V_{IN}$ , so  $V_{OUT}$  goes low.  $V_{OUT}$  stays low until  $V_{IN}$  drops below  $V_2$ . When  $V_{IN}$  drops below  $V_2$ ,  $V_{OUT}$  goes high again;  $V_2 = V_{CC}$ ; and  $R_1$ ,  $R_2$ , and  $R_3$  set  $V_1$ .

You can easily set either trip voltage without affecting the other. However,  $V_1$  must be greater than  $V_2$  or both trip voltages will be equal to  $V_1$ . A disadvantage of this circuit is that it requires two comparators, but comparators usually come in dual packages. (DI #2383)

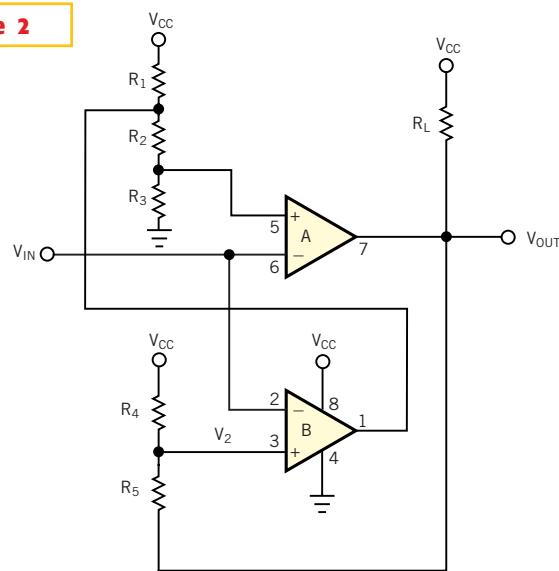
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**Figure 1**



Three resistors determine the trip voltages of an inverting comparator (a) with hysteresis (b), but the trip voltages are not independent.

**Figure 2**



Using a dual comparator with open-collector outputs allows you to set either trip voltage without affecting the other.

# Modem-access adapter reduces interference

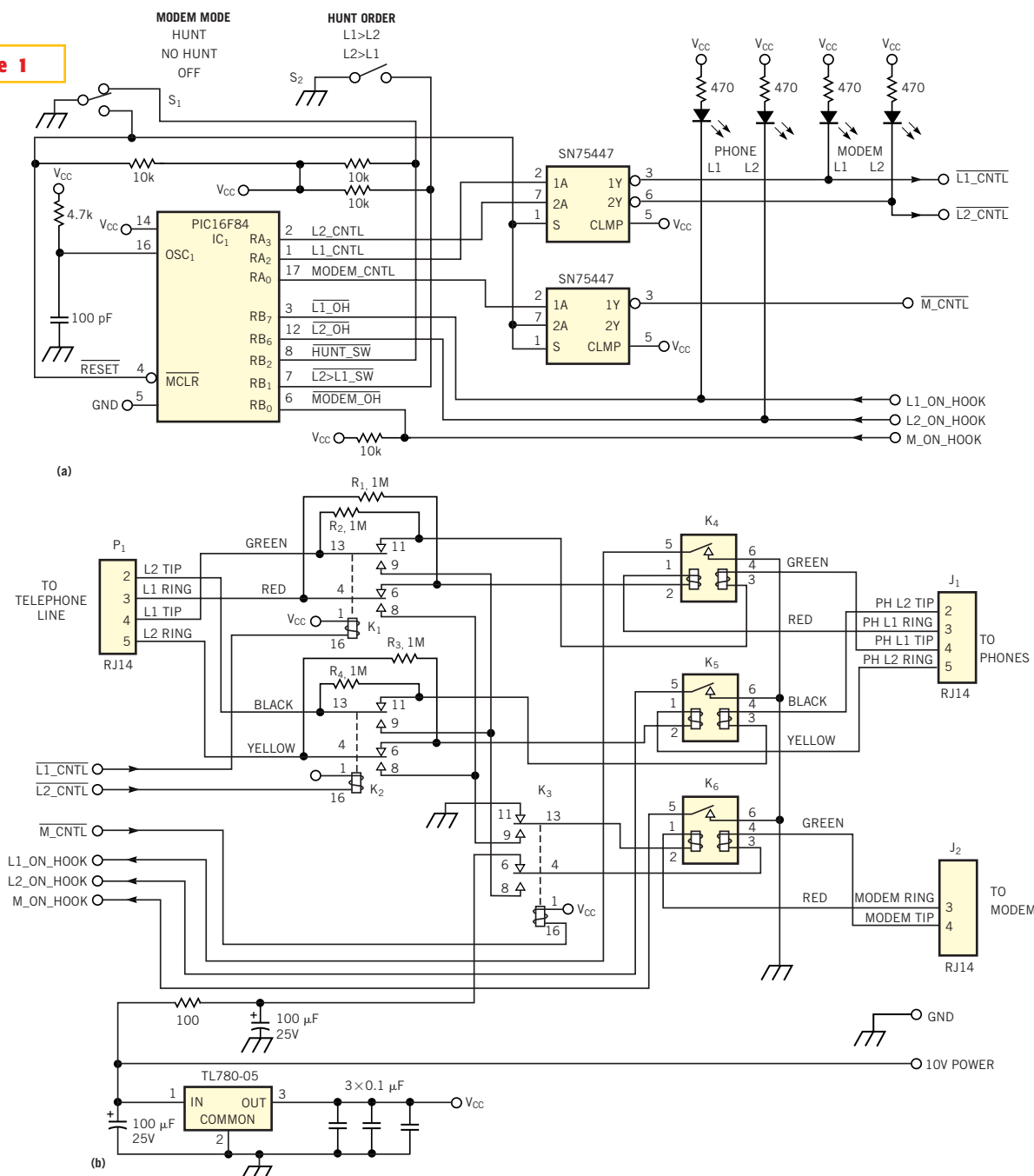
Thomas Schmidt, Schmidt Consulting, Milford, NH

**T**HE MODEM-ACCESS ADAPTER (MAA) in Figure 1 has features that reduce problems when modems and phones

share the same line. The design also overcomes the shortcomings of modem interference protectors. The design elimi-

nates mutual interference between phones and modems, is compatible with the line-usage indicators on multiline

**Figure 1**



A low-cost  $\mu$ C with flash memory controls the connection of phone lines L1 and L2 to the modem (a). Electromechanical relays switch the phone lines while providing an isolated data path (b).



phones, can search for an idle phone line in multiline installations, draws minimum power, and requires no changes to a PC modem's hardware or software.

Most residential PC users connect to the Internet using a dial-up modem. In most cases, the modem and voice calls must share one phone line. This sharing presents a problem because you can't use the modem and the phone at the same time; picking up an extension disconnects the modem. Many sites have more than one phone line. A computer has a better chance of connecting to a phone line if it can search multiple lines to find one that is idle.

Vendors of modem interference protectors claim the devices eliminate interference between extension phones and computer modems. The devices function by monitoring the voltage between the phone line's tip and ring leads. When all phones are on their hooks, this voltage is typically >48V. Picking up a phone closes the tip-and-ring circuit, which drops the voltage to approximately 7V. The protector monitors the line voltage. When the voltage drops below a certain threshold, the protector disconnects the phones that are connected to it. This action prevents an active phone from interference when you pick up an extension.

Unfortunately, these protectors have two serious shortcomings. First, the device needs to be in series with the phones and modems it is protecting, which requires one device per phone. When you use the device in this manner, you cannot use more than one extension at a time. This situation prevents you from placing a call on hold and picking it up on another extension. Second, you can split the wiring into two circuits: one for phones and one for the modem. This split wiring minimizes the number of protectors you need and allows multiple extensions to pick up the same call. Unfortunately, regardless of how you wire the protectors, the voltage drop across the protector confuses the line-busy indicators on most multiline phones. Multiline

phones no longer indicate when a line is in use.

The MAA in **Figure 1** does not suffer from these problems. You must connect the MAA in series between the telephone-company central office and the phones and modem the adapter is protecting. The most convenient location for the MAA is in proximity to the phone-company network's interface demarcation point. Simply disconnect the inside phone wiring from the demarcation point and plug this wiring into the MAA. Then run a dedicated phone line between the MAA and computer modem.

The MAA consists of control (**Figure 1a**) and data-path (**Figure 1b**) subsystems. Phone lines source high voltages and are balanced with respect to ground. The design uses electromechanical relays to switch the phone lines while providing an isolated control path.

The heart of the control circuit is IC<sub>1</sub>, a PIC16F84 (Microchip Technology, [www.microchip.com](http://www.microchip.com)) low-cost RISC  $\mu$ C with flash memory. Flash memory makes this part ideal for development. The MAA program is approximately 100 words long. The accompanying source code is available for downloading from EDN's Web site, [www.ednmag.com](http://www.ednmag.com). Click on "Search Databases" and then enter the Software Center to download the file for Design Idea #2389.

When the modem is idle, relay K<sub>3</sub> supplies power to the modem local loop (**Figure 1b**). When the modem goes "off hook" to place a call, relay K<sub>6</sub> senses current flow and wakes the CPU. The CPU reads the state of the hunt-order switch, S<sub>2</sub>, which is a physical switch that allows the user to select the phone-line search order.

Relay K<sub>4</sub> monitors Line 1 (L1), and K<sub>5</sub> monitors Line 2 (L2). When a phone is off hook, loop current causes the appropriate line-sense relay contact to close. The CPU uses this information to determine if the line is in use. Assuming the hunt-order switch is in the search-L1-before-L2 position, the CPU reads the state of L1. If L1 is idle, the CPU energizes K<sub>1</sub>

using the L1\_CNTL signal. Energizing K<sub>1</sub> disconnects the extension phones from L1 and routes L1 to the normally open contacts on K<sub>3</sub>. Then the CPU energizes K<sub>3</sub>, which connects the modem to L1.

If L1 is busy, the CPU reads the modem-mode switch. If the setting of S<sub>1</sub> enables the hunt mode, the CPU checks the other line for availability. If L2 is idle, the CPU energizes K<sub>2</sub> using the L2\_CNTL signal. Energizing K<sub>2</sub> disconnects the extension phones from L2 and routes L2 to the contacts on K<sub>3</sub>. The CPU then energizes K<sub>3</sub> to connect the modem to L2. If both lines are busy, the modem never gets a dial tone and eventually disconnects.

The telephone central office typically disconnects the phone line for a short time at the beginning of the dial tone. After the circuit establishes the connection, the CPU enters a long delay loop. During this time, the CPU ignores changes in the state of the line. At the end of the delay, the CPU checks to see if the line is in the expected condition. If it is, the CPU goes back to sleep. If it is not, the CPU processes the new state.

When the modem hangs up, relay K<sub>6</sub> deenergizes, which also wakes the CPU. When the CPU detects hang-up, it tears down the modem connection and returns the lines to their default state for voice.

The bleeder resistors R<sub>1</sub> through R<sub>4</sub> discharge the phone capacitance and impress the line voltage on all of the extension phones to ensure that the telephone-busy indicators function normally. Relays K<sub>4</sub> through K<sub>6</sub> (Teltone, [www.teltone.com](http://www.teltone.com)) are specially designed to monitor telephone loop current.

The MAA device connects directly to the phone line and therefore must meet FCC CFR 47 Part 68 for protection of the public switched telephone network. Residential devices must meet FCC CFR 47 Part 15 Class B for limitation on unintended radiation. (DI #2389)

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