WITH CUMULATIVE INDEX

Rudolf F. Graf
&
William Sheets

Encyclopedia of ELECTRONIC CIRCUITS

Volume 7

TAB
The Electronics Authority

Volume 7 ; Partition #3
CH 32; Doubler Circuits to CH 49; Load Circuits

To Scott
With much love from Popsi

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Encyclopedia of

ELECTRONIC CIRCUITS

Volume 7

Rudolf F. Graf
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Contents

Preface

This latest volume of *The Encyclopedia of Electronic Circuits* contains approximately 1000 new electronic circuits that are arranged alphabetically into more than 100 basic circuit categories, ranging from "Active Antenna Circuits" to "Weather-Related Circuits." When taken together with the contents of the previously published six volumes, we provide instant access to more than 7000 circuits that are meticulously indexed and cross referenced. This represents, by far, the largest treasure trove of easy-to-find, practical electronic circuits available anywhere.

We wish to express our sincere gratitude and appreciation to the industry sources and publishers who graciously allowed us to use some of their material. Their names are shown with each entry and further details are given at the end of the book under "Sources."

Our thanks also go out to Ms. Tara Troxler, whose skill at the word processor and dedication to this project made it possible for us to deliver the manuscript to the publisher in a timely fashion.

Rudolf F. Graf and William Sheets
January 1998

Encyclopedia of

ELECTRONIC CIRCUITS

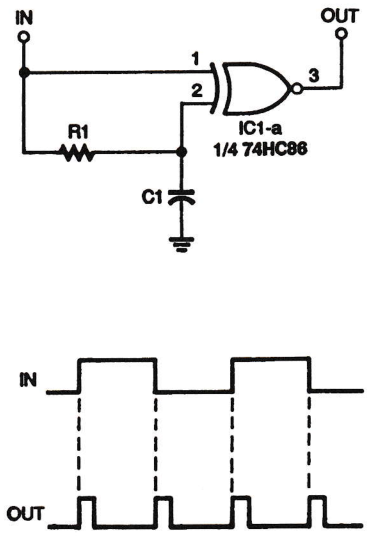
Volume 7

32

Doubler Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

SIMPLE FREQUENCY DOUBLER

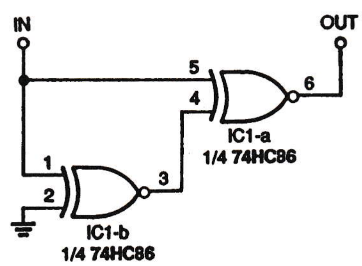


This circuit uses only a single exclusive-OR gate and a couple of passive components. The width of the output pulses is determined by the time constant of the RC network, and the maximum input frequency cannot exceed $\frac{1}{2}RC$.

ELECTRONICS NOW

Fig. 32-1

DIGITAL LOGIC FREQUENCY DOUBLER

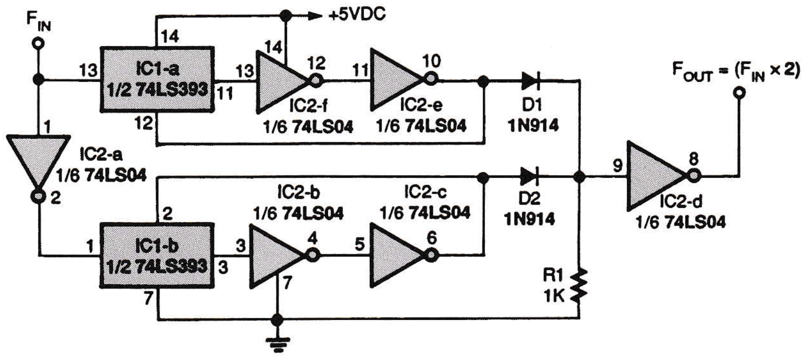


This circuit can be used if very high frequency operation is needed, or if very narrow output pulses can be tolerated. The output pulse width is determined by the propagation time delay of the exclusive-OR gate, and the input frequency cannot exceed $\frac{1}{2}$ (delay).

ELECTRONICS NOW

Fig. 32-2

DIGITAL FREQUENCY-DOUBLER CIRCUIT



ELECTRONICS NOW

Fig. 32-3

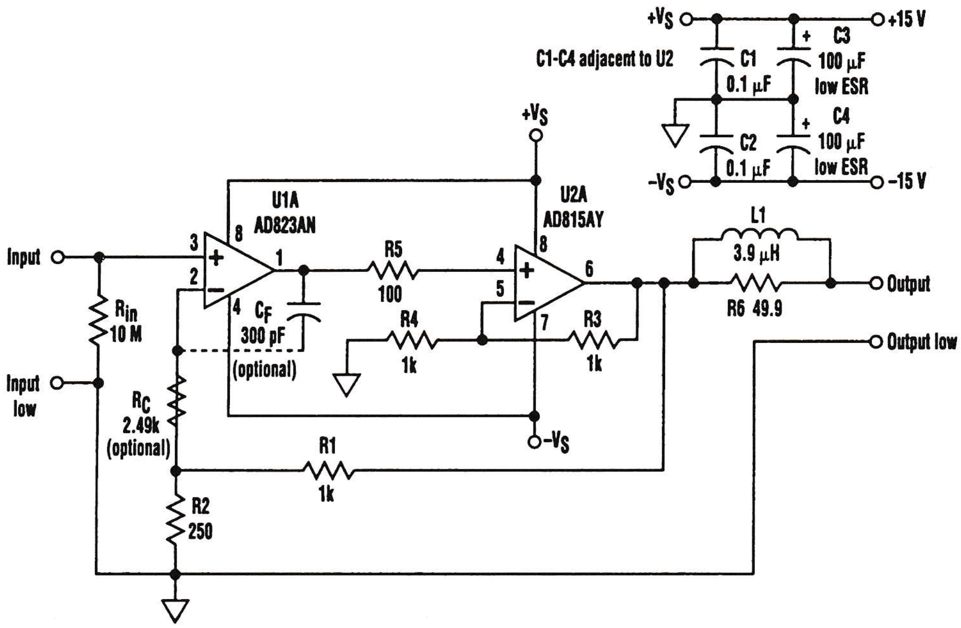
The 74LS393 is designed to advance one count on the positive-to-negative transition of the pulse at the clock input (pins 1 and 13). An inverter in front of the B section of the counter causes the B section to advance one count on the negative-to-positive transition of the input pulse. Each section of the counter is cleared with a positive-going pulse on the master reset input (pins 2 and 12). Positive-going output pulses from pins 6 and 10 of IC2 reset the respective counters. The duration of the pulses depends upon the propagation delay of the inverters. With the 74LS04 hex inverter, delay will probably be in the vicinity of 20 to 25 ns. The output pulses are also connected to the remaining inverter gate through switching diodes and a pull-down resistor, which configures the remaining inverter as a NOR gate. The output at pin 8 of IC2 represents the input frequency multiplied by 2.

33

Driver Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

LOW-DISTORTION LINE DRIVER

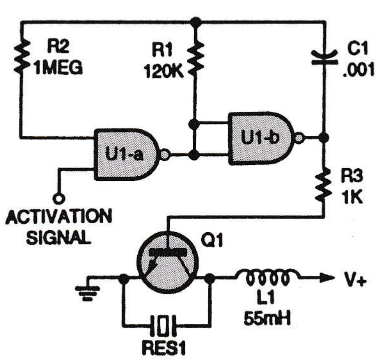


ELECTRONIC DESIGN ANALOG APPLICATIONS

Fig. 33-1

This low-distortion driver circuit delivers up to ±0.5 A and is suitable for loads of 10 Ω and up. Using a low-offset, low-bias-current input stage, the driver can be entirely direct-coupled. Gain is equal to $1+R_2/R_1$.

TWO-TERMINAL PIEZO DEVICE DRIVER

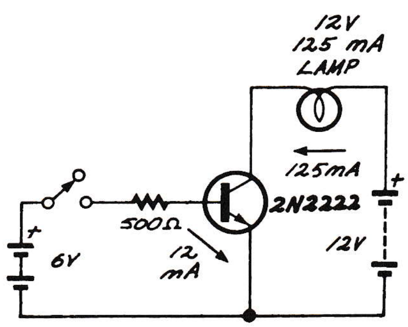


Two-terminal devices can be driven by two NAND gates. A booster coil is used to compensate for the sound-pressure attenuation caused by the case.

POPULAR ELECTRONICS

Fig. 33-2

SIMPLE LAMP DRIVER

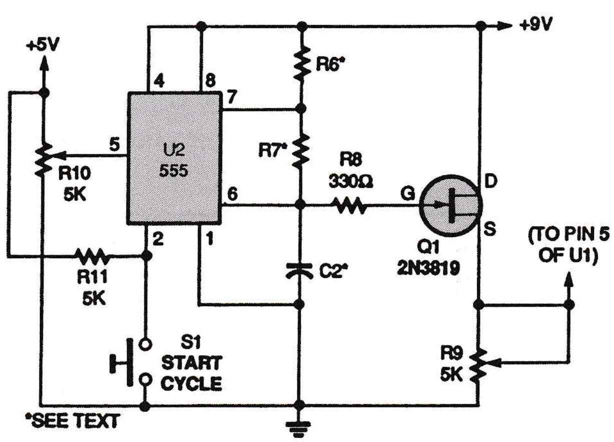


ELECTRONICS NOW

Fig. 33-3

Small current through the base controls large current through the collector. Close the switch, and the bulb lights.

R/C SERVO SWEEP DRIVER CIRCUIT

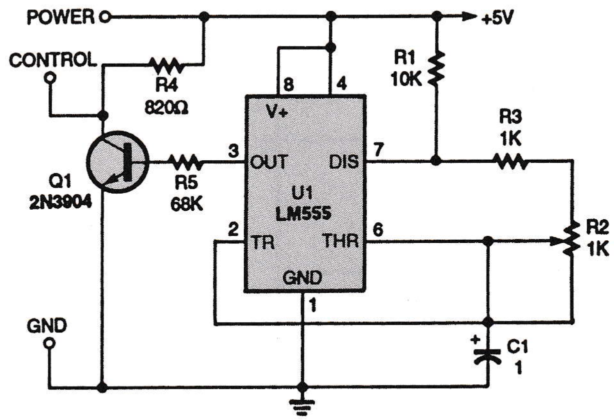


POPULAR ELECTRONICS

Fig. 33-4

The circuit shown produces a ramp that can be used to modulate the threshold of another 555 timer, which is configured to generate the pulse signal to drive the R/C servo. In this way, the servo can be slowly moved through a desired angle. Note that for long time constants, R6 and R7 will be large. Therefore, C2 should be a low-leakage capacitor, or a CMOS timer (7555) can be used to permit larger resistances and smaller capacitors to be used.

R/C SERVO DRIVER CIRCUIT

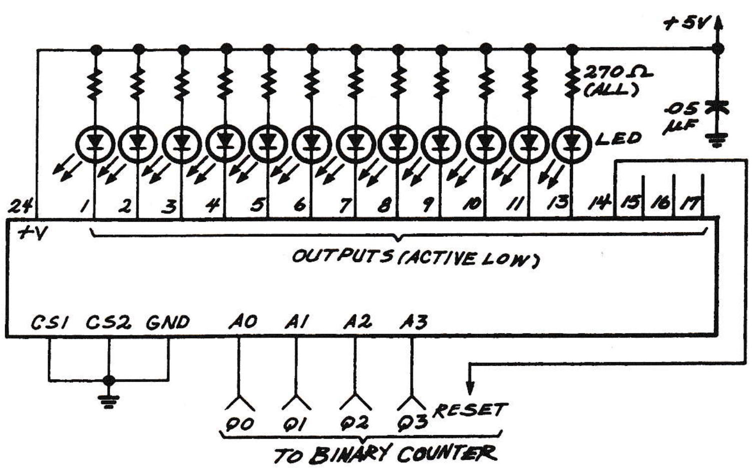


POPULAR ELECTRONICS

Fig. 33-5

The circuit shown produces the necessary pulse signal to drive R/C servos through a 90° range by adjustment of the pot R2.

12-LED SEQUENTIAL DRIVER

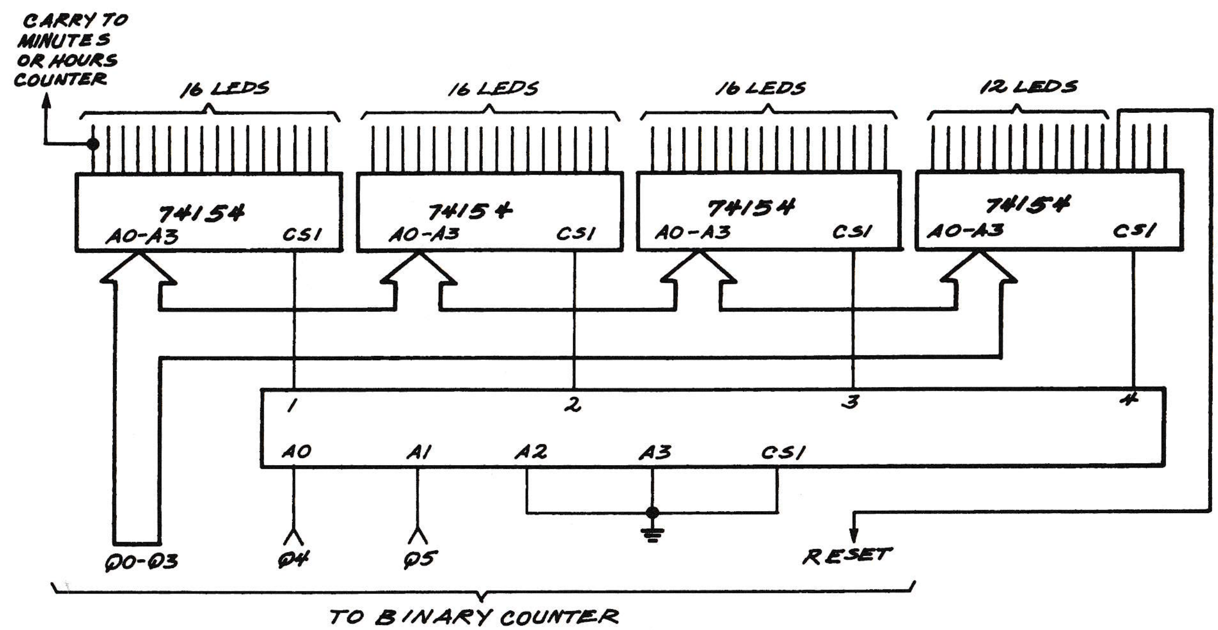


ELECTRONICS NOW

Fig. 33-6

A demultiplexer selects a different LED for each binary number at its input. The counter resets after a count of 12.

60-LED SEQUENTIAL DRIVER

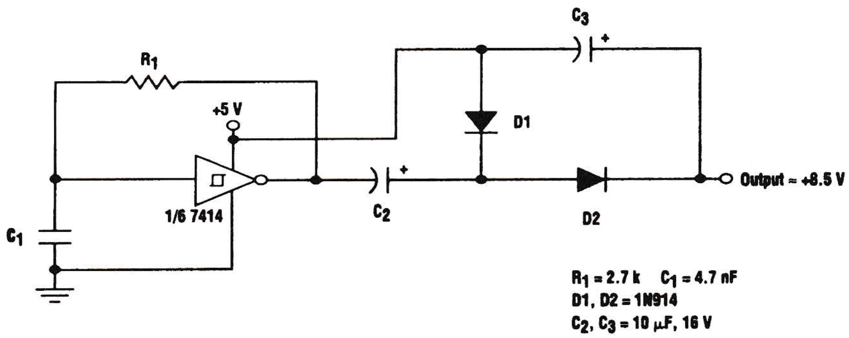


ELECTRONICS NOW

Fig. 33-7

This circuit is useful as the minute hand for an LED clock or as a general-purpose sequential driver.

MOSFET GATE DRIVER



ELECTRONIC DESIGN

Fig. 33-8

This ratiometric 20-kHz voltage-to-frequency converter (VFC) provides superior performance with strain gauges and other ratio-responding transducers, even with noisy, unregulated excitation voltages. Feedback isn't used to achieve the excellent 4-Hz linearity, so there is very low frequency jitter—period measurements can be used to get several digits of resolution even when operating at a fraction of full scale. An operational synchronizing transistor starts the VFC with zero charge at the beginning of each count cycle, eliminating the characteristic digit jumping often encountered with VFC designs. Good linearity is attained by making the comparator's reference voltage vary with the input voltage, which precisely compensates for the finite capacitor reset time:

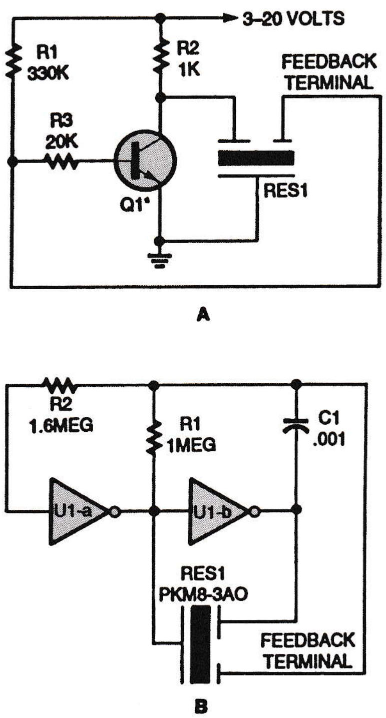
$$
\begin{aligned}
Period &= t_1 + t_2 \\
&= t_1 + (V_{CC} - V_{ref})/AV_{in} \\
&= (t_1 AV_{1n} + V_{CC} - V_{ref})/AV_{in}
\end{aligned}
$$

where $AV_{in} = \Delta V/\Delta t$. If V_{ref} is made to include the amount $t_1 AV_{in}$, then the effect of t_1 is eliminated:

$$Period = [t_1 AV_{in} + V_{CC} - (t_1 AV_{in} + V_{ref})]/AV_{in}$$

The MPSA-18 is a remarkably high-gain transistor, even at low currents, giving good current-source linearity down to 0 Hz. In addition, bipolar transistors work well with the low collector voltages encountered in this single-supply, 10-V design. Moreover, most single-supply op amps will work in place of the LM10. But the LM10 also has a reference amplifier that could be used to construct a 10-V excitation regulator. The LM311 propagation delay gives a reset pulse width near 400 ns, which gives the transistor time to discharge the capacitor. Also, the 311's bias current gives a small negative offset that ensures a 0-Hz output for 0 V in.

PIEZOELECTRIC DRIVER CIRCUIT



POPULAR ELECTRONICS

Fig. 33-9

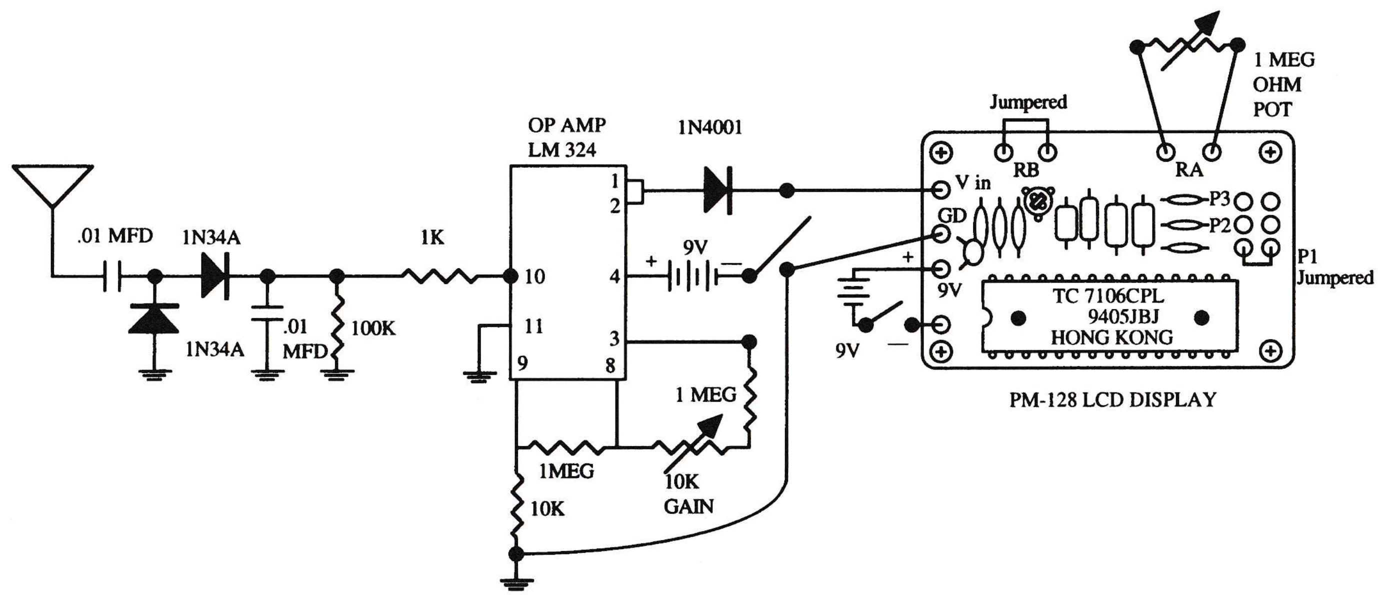
Three-terminal piezoelectric elements are typically driven by transistor circuits (A) or logic gates (B).

34

Field-Strength-Measuring Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

DIGITAL FIELD-STRENGTH METER

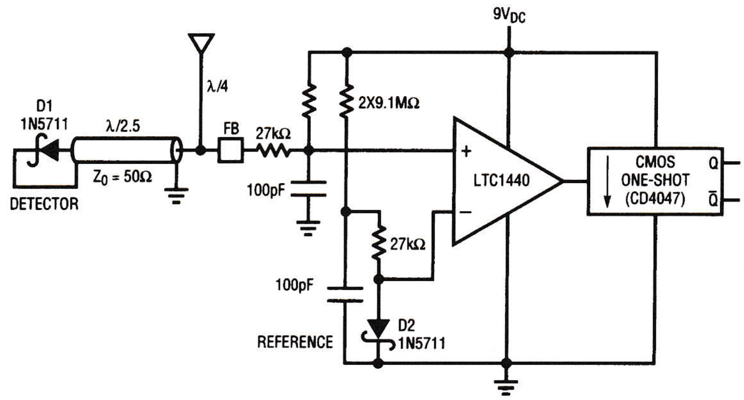


73 AMATEUR RADIO TODAY

Fig. 34-1

This field-strength meter uses a surplus LCD panel meter display, but any suitable unit (1 V full scale, etc.) can be used. RF is detected by a pair of 1N34 diodes and fed to an LM324 op amp acting as a dc amplifier whose gain is set via a 1-MΩ pot. The op-amp output drives the LCD panel meter.

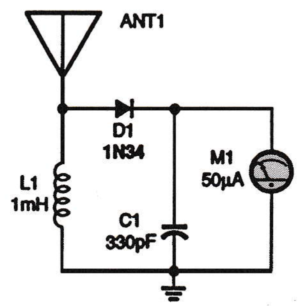
MICROPOWER FIELD DETECTOR FOR 470 MHz



LINEAR TECHNOLOGY *Fig. 34-2*

This circuit, which was tested at 470 MHz, contains a couple of improvements over the standard L/C-with-whip field-strength meter. The 0.4-wavelength section presents an efficient, low-impedance match to the base of the quarter-wave whip, but transforms the received energy to a relatively high voltage at the diode for good sensitivity. Biasing the detector diode improves the sensitivity by an additional 10 dB. The detector diode's bias point is monitored by an LTC1440 ultra-low-power comparator and by a second diode, which serves as a reference. Schottky diode D1 rectifies the incoming carrier and creates a negative-going bias shift at the noninverting input of the comparator. Note that the bias shift is sensed at the base of the antenna, where the impedance is low, rather than at the Schottky, where the impedance is high. This introduces less disturbance into the tuned antenna and transmission-line system. The falling edge of the comparator triggers a one-shot, which temporarily enables answer-back and other pulsed functions. Total current consumption is approximately 5 μA. Alternatively, a discrete one-shot constructed from a quad NAND gate draws negligible power. Sensitivity is excellent. The finished circuit can detect 200 mW radiated from a reference dipole at 100 ft. Range, of course, depends on operating frequency, antenna orientation, and surrounding obstacles; in the clear, a more reasonable distance, such as 10 ft, can be covered at 470 MHz with only a few milliwatts. All selectivity is provided by the antenna itself. Add a quarter-wave stub (shorted with a capacitor) to the base of the antenna for better selectivity and improved rejection of low-frequency signals.

FIELD-STRENGTH METER

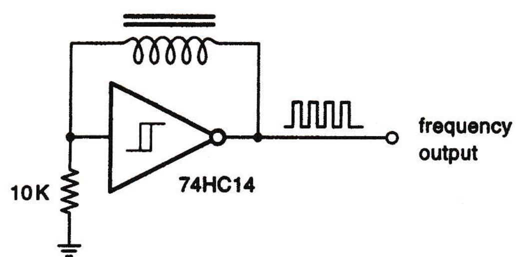


This simple field-strength meter is great for tuning transmitters or antennas. It's a tool no ham should be without. The antenna is a 24-in whip.

POPULAR ELECTRONICS

Fig. 34-3

EARTH'S FIELD DETECTOR

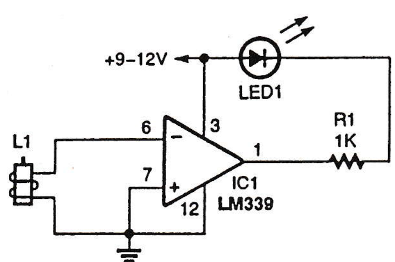


POPULAR ELECTRONICS

Fig. 34-4

This elegantly simple earth's field detector uses a special variable permeability cored coil. The output frequency varies with orientation.

MAGNETIC FIELD DETECTOR

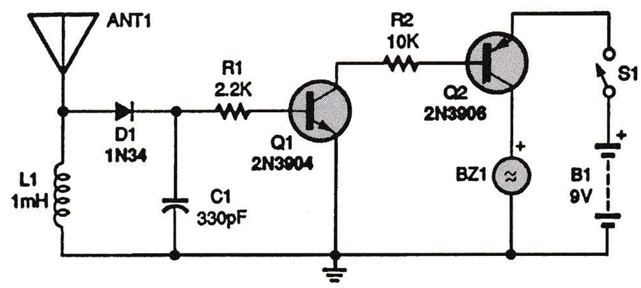


A relay coil makes a great magnetic-field detector.

POPULAR ELECTRONICS

Fig. 34-5

AMPLIFIED FIELD-STRENGTH METER

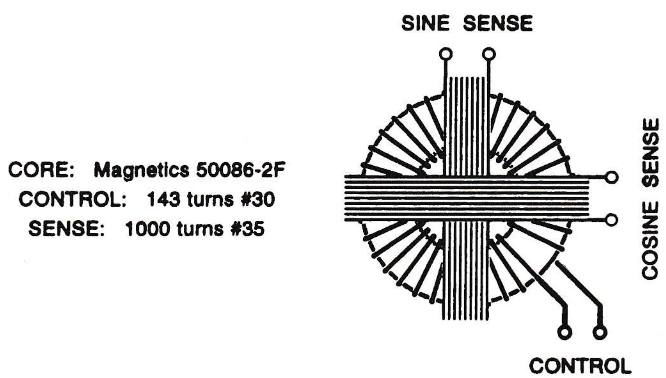


POPULAR ELECTRONICS

Fig. 34-6

This field-strength meter uses Q1 and Q2 to amplify the dc voltage produced by detector D1. A piezo sound BZ1 is used as an audible indicator, rather than using a meter. This circuit could be of use for the visually handicapped. The antenna is a 24-in whip.

ANALOG FLUXGATE MAGNETOMETER ASSEMBLY



ELECTRONICS NOW

Fig. 34-7

An audio input drives the sine-wave control input, switching (or "gating") the core in and out of saturation and drawing in or releasing an external magnetic field. Weak signals at the sense outputs are proportional to field strength.

MAGNETIC FIELD METER

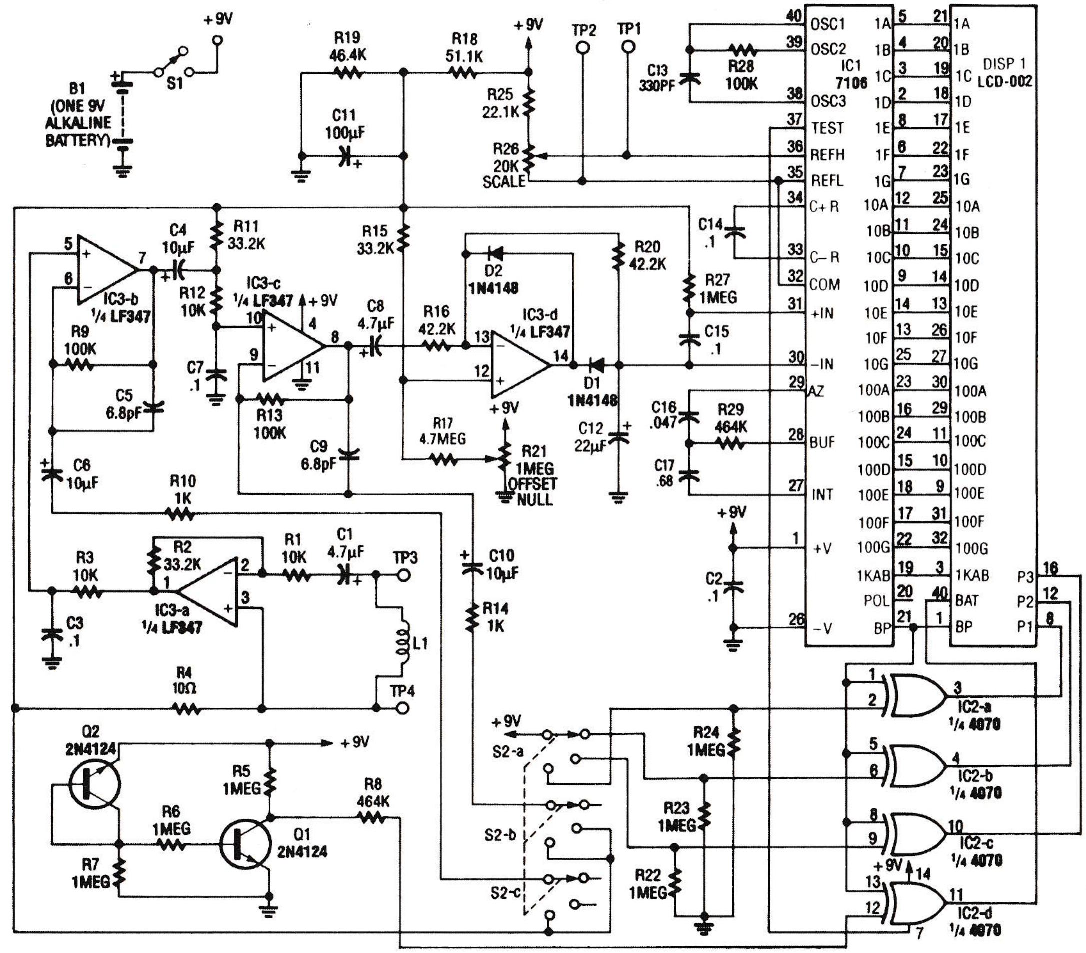
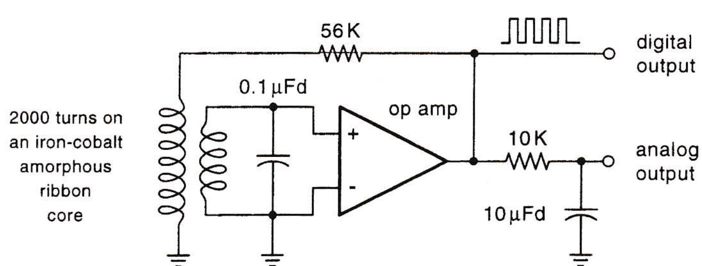


Fig. 34-8

The meter's 12-turn field pickup is integrated into the unit's circuit board. For remote sensing, an external field coil probe can be used. The magnetic field picked up by the coil appears as a voltage, which is proportional to field strength and frequency at the input of a cascaded amplifier IC3-a, IC3-b, and IC3-c. With a first-stage amplifier gain of 3.3 set by R12 to R10, the overall sensitivity is 100 μV/μT, or 100 mV/mT. The meter sensitivity is nominally 2 V full scale, leading to the lowest-level sensitivity of 20 mT full scale. Op amp IC3-a amplifies the signal to a normalized level of 100 μV/μT. The voltage is further amplified by 1, 100, or 10,000 by IC3-b and IC3-c. The three amplifier stages provide the ranges of 2 mT, 200 μT, and 2 μT (full scale). Components R3 and C3 and R12 and C7 establish a frequency rolloff characteristic that compensates for the frequency-proportional sensitivity of the pickup coil, and set the 20-kHz cutoff point. IC3-d is a precision rectifier and peak detector. Its output drives IC1, a combination analog-to-digital (A/D) converter and LCD driver. Components R25 to R29 and C13 to C17 are used by IC1 to set display-update times, clock generation, and reference voltages. The decimal points are driven by IC2, as determined by range-select switch S2. Transistors Q1 and Q2 serve as a low-battery detector, and turn on the battery annunciator in the LCD when the battery voltage drops below 7 V.

RESONANT FLUXGATE MAGNETOMETER



ELECTRONICS NOW *Fig. 34-9*

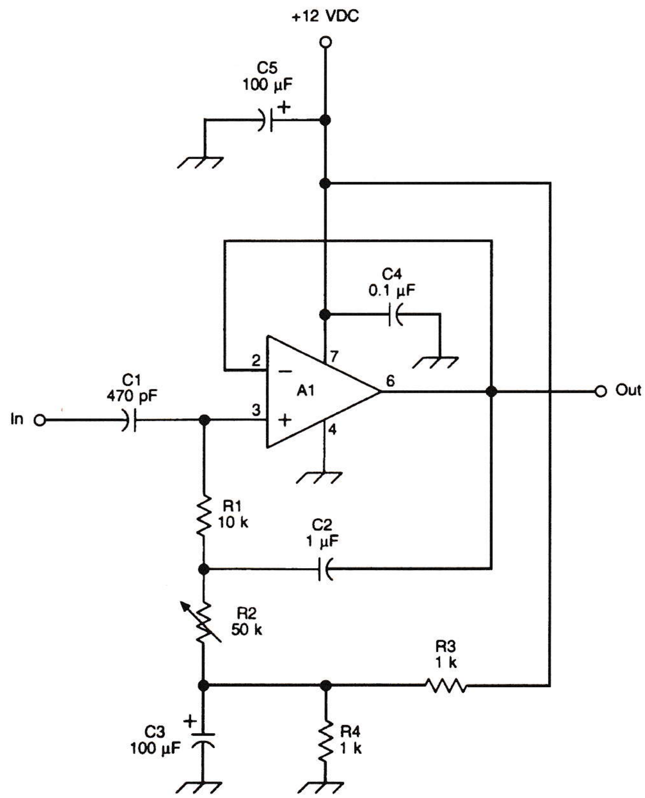
The core will get switched in and out of saturation. The output duty cycle varies in proportion to the single-axis field strength and direction. The high-level output square wave is easy to interface to a PIC or other microcontroller.

35

Filter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

PEAKING FILTER CIRCUIT



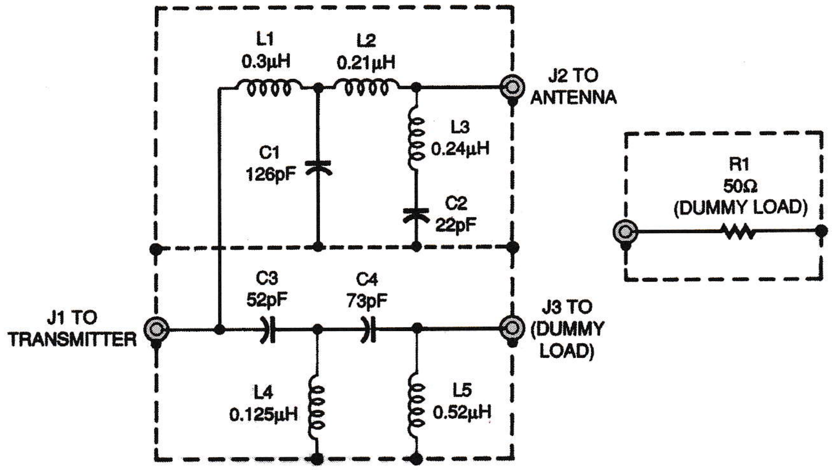
73 AMATEUR RADIO TODAY ***Fig. 35-1***

This circuit is basically a noninverting amplifier with positive feedback. The negative feedback is to unity by a direct connection between the output terminal of the operational amplifier and the inverting (−) input. A frequency-selective RC network connects the output to the noninverting (+) input, providing some positive feedback. The center frequency is

$$F_C = \tfrac{1}{2}\pi R_1 (C_1 C_2)^{1/2}$$

where F_C is the center frequency in hertz, R_1 is in ohms, and C_1 and C_2 are in farads.

AMATEUR TRANSMITTER ABSORPTIVE FILTER

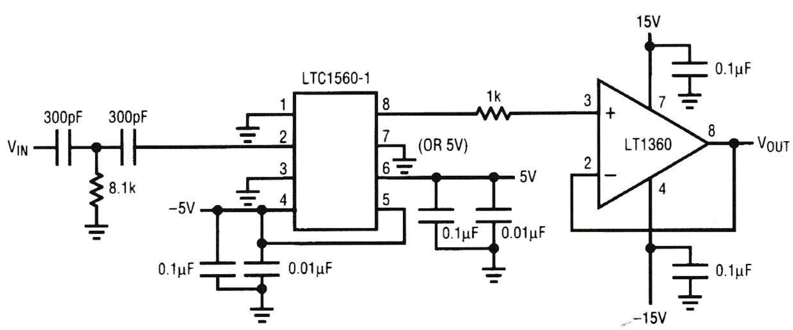


POPULAR ELECTRONICS

Fig. 35-2

This filter diverts harmonic energy above 30 MHz to a dummy load and dissipates it as heat instead of reflecting it back to the transmitter. The absorptive filter is based on the classic design by Weinreich and Carroll from 1968. The high-pass and low-pass sections are shielded from each other (even though they are in the same shielded box) to prevent interaction. If you want to try building this filter, wind the coils on high-powered toroids. Alternatively, you could wind an air-core coil using the nomograms in *The ARRL Handbook for Radio Amateurs.*

HIGH-PASS / LOW-PASS FILTER

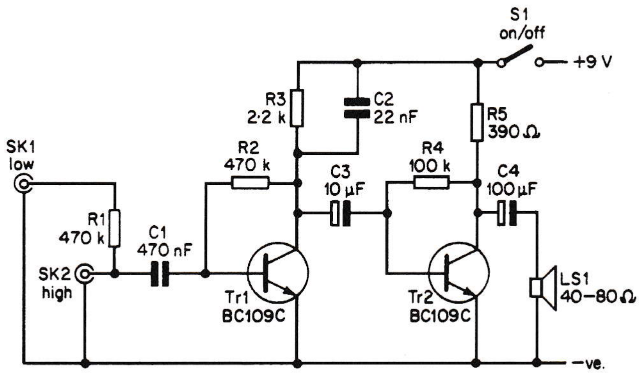


LINEAR TECHNOLOGY

Fig. 35-3

This filter uses a Linear Technology P/N LTC1560-1 and an LT1360 to implement a high-pass / low-pass filter.

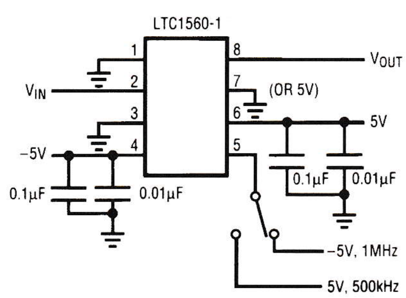
SW RECEIVER CW FILTER



ELECTRONIC EXPERIMENTERS HANDBOOK ***Fig. 35-4***

The CW filter is connected between the headphone or loudspeaker socket or terminal strip of your receiver and either the headphones or an external loudspeaker. The output of the receiver should have an impedance of 8 Ω or more. As the circuit provides unity gain at pass frequencies and a low-impedance output, there should be no problems with mismatching when the filter is in use. The frequency response of the circuit peaks at approximately 800 Hz, and the −6-dB bandwidth is about 300 Hz or so. The 0-dB points occur at about 350 Hz and 2 kHz. This is sufficient to normally give a substantial reduction in adjacent-channel interference, but the response is not so narrow and peaky that using the receiver with the filter in the circuit becomes difficult, as the wanted signal tends to drift out of the passband and become lost.

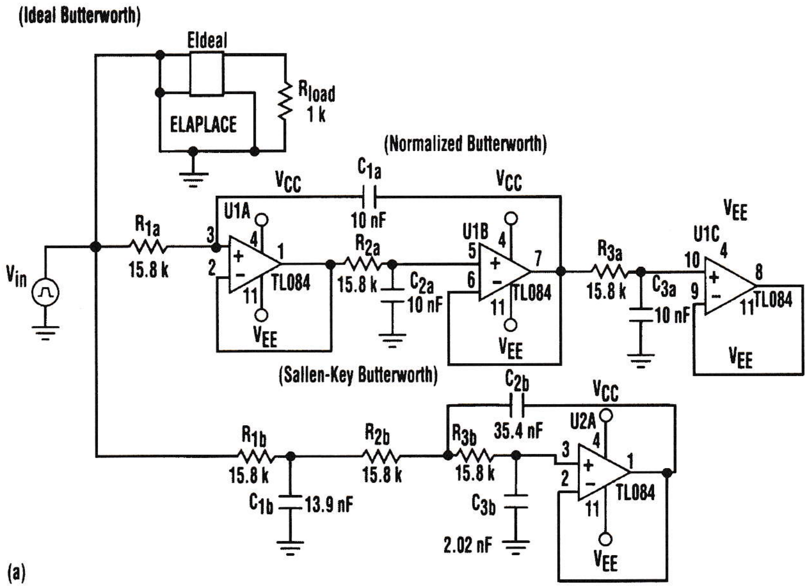
1-MHz OR 500-kHz ELLIPTIC LP FILTER



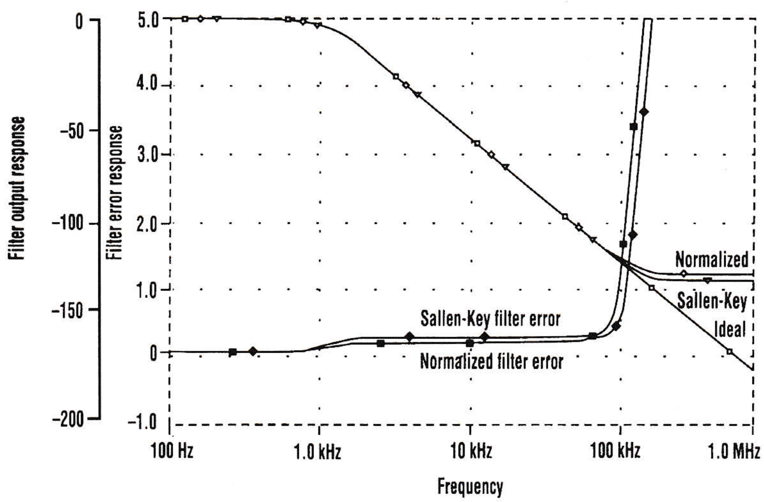
This filter uses a Linear Technology P/N LTC1560-1 to implement a switchable LP filter.

LINEAR TECHNOLOGY ***Fig. 35-5***

ACTIVE FILTER



(a)



(b)

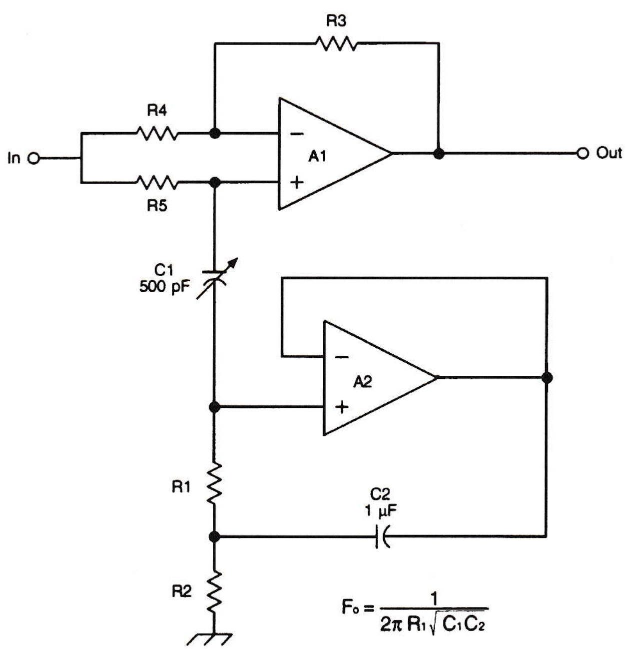
ELECTRONIC DESIGN

Fig. 35-6

ACTIVE FILTER *(Cont.)*

Equal-value components can be quite an advantage in filter designs when the total costs associated with the procurement, stocking, and assembly of the filter are considered. For instance, the Butterworth active third-order low-pass filter (middle) uses equal-value resistors and capacitors. This feature normalizes the filter's 3-dB corner frequency to $1/RC$ (in radians) for both low-pass and high-pass designs. The graphs in (*b*) are plots of the ideal, normalized, and Sallen-Key low-pass filter frequency-domain magnitude and error responses. Note how both the normalized and Sallen-Key filters follow the ideal response well into the stopband. The error plots were created by plotting the difference between the real and ideal filter responses. The plots indicate that the normalized filter achieves performance results that are equal to those of the Sallen-Key low-pass filter.

NOTCH FILTER CIRCUIT



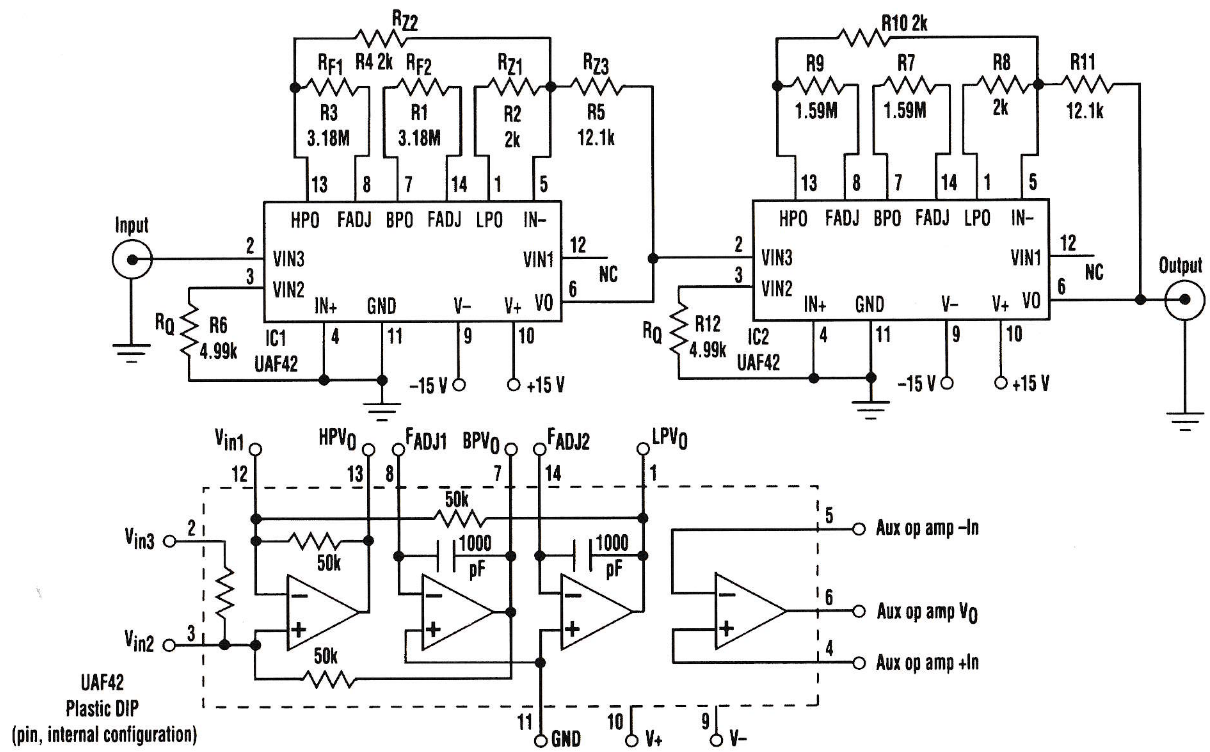
73 AMATEUR RADIO TODAY ***Fig. 35-7***

This circuit is basically a noninverting amplifier with positive feedback. The negative feedback path is set to unity by a direct connection between the output terminal of the operational amplifier and the inverting (−) input. A frequency-selective RC network connects the output to the noninverting (+) input, providing some positive feedback. The center frequency is

$$F_C = \tfrac{1}{2}\pi R_1 (C_1 C_2)^{1/2}$$

where F_C is the center frequency in hertz, R_1 is in ohms, and C_1 and C_2 are in farads.

CAPACITORLESS NOTCH FILTER



ELECTRONIC DESIGN

Fig. 35-8

The notch frequency for the filter is set by

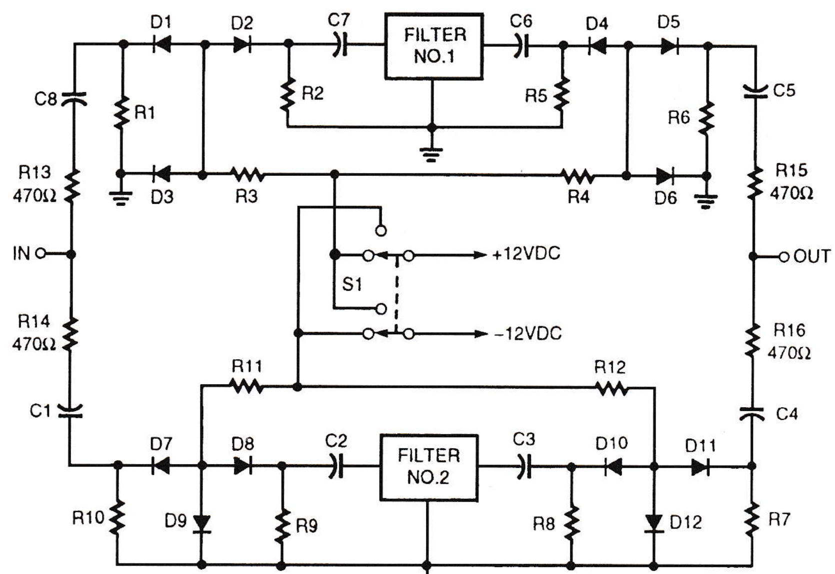
$$f_{notch} + [(A_{LP}/A_{HP}) \times (R_{Z2}/R_{Z1})]^{1/2} \times f_0$$

where A_{LP} and A_{HP} are the gain from input to low-pass output at $f=0$ Hz, and the gain from input to high-pass output at $f \ll f_0$, respectively. Typically $(A_{LP}/A_{HP}) \times (R_{Z2}/R_{Z1}) = 1$; therefore, $f_{notch} = f_0$, and is given by

$$f_0 = 1/(2\pi)(R_F C)$$

where $R_F = R_{F1} = R_{F2}$, and $C = C_1 = C_2$. The −3-dB bandwidth is determined by the following relation: $BW_{-3dB} = f_{notch}/Q$, where $BW_{-3dB} = f_H - f_L$. The Q of the filter affects the passband gain (which should be adjusted to unity) and is related to the ratio of the resistances R_{Z3} to R_{Z1} and R_{Z3} to R_{Z2}. In other words, $Q = (R_{Z3}/R_{Z1}) = (R_{Z3}/R_{Z2})$. Q also is related to R_Q by the following relation: $R_Q = (25\ k\Omega/Q - 1)$.

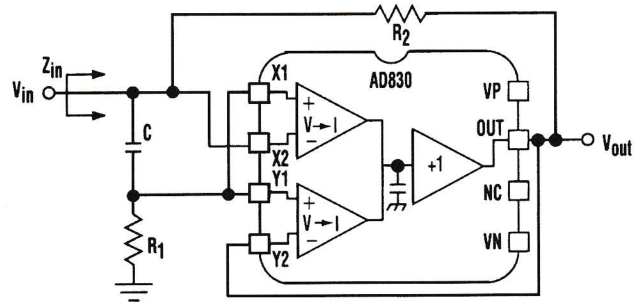
IF BANDPASS FILTER SWITCH

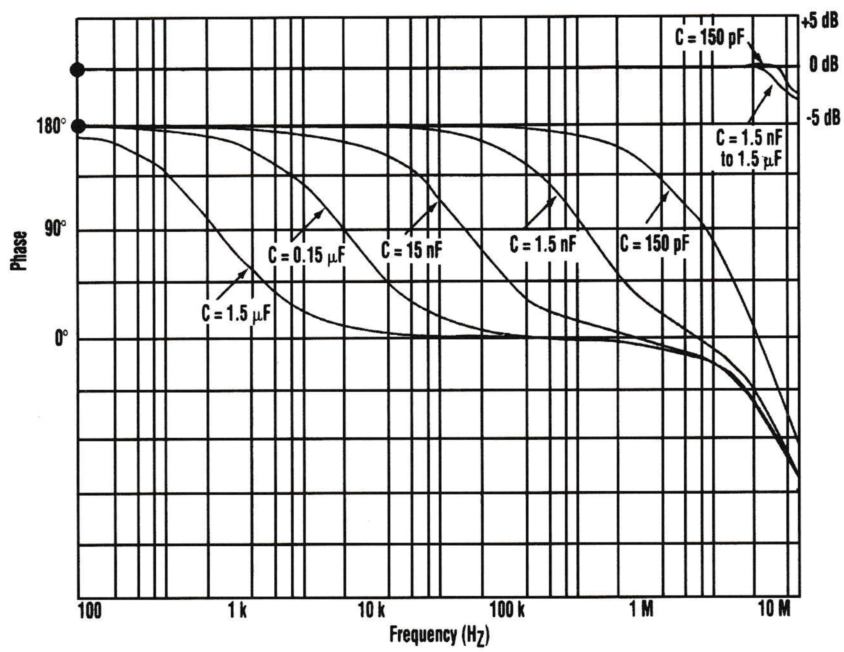


POPULAR ELECTRONICS *Fig. 35-9*

Selecting IF bandpass filters via series/shunt PIN-diode switching can be accomplished with this circuit. Diodes can be MV3404 or similar types.

SIMPLE ALL-PASS FILTER





ELECTRONIC DESIGN *Fig. 35-10*

A very simple all-pass implementation can be realized with an active-feedback amplifier like the AD830 or the LTC1193. R1 and C set the filter's actual transfer function, while R2 is needed to provide a purely real input impedance over the frequency range of the AD830 (necessary for measurement reasons). The filter's two basic equations are

$$Z_{\text{in}}(s) = \frac{R_2(1+sCR_1)}{2\left(1+\frac{sCR_2}{2}\right)}$$

and

SIMPLE ALL-PASS FILTER *(Cont.)*

$$\frac{V_{out}(s)}{V_{in}} = \frac{-1 - sCR_1}{1 + sCR_1}$$

from which we can see that the magnitude is constant:

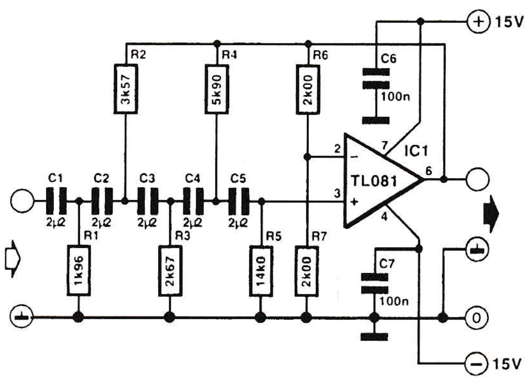
$$\frac{V_{out}(\omega)}{V_{in}=1}$$

and the phase of V_{out}/V_{in} as a function of ω is

$$180° - 2\tan^{-1}(\omega CR_1)$$

From the first equation, it's clear that for Z_{in} to be purely real, R_1 has to be equal to $R_2/2$, which implies $Z_{in}(\omega)=R_1$. Once C is chosen, R_1 and R_2 can be picked according to the termination and required phase shift. The graph shows the circuit's performance for $V_s=\pm15$ V, $R_1=100$, $R_2=200$, and values of C from 1.5 μF to 150 pF with 90° phase shifts at one-decade increments up to 10 MHz.

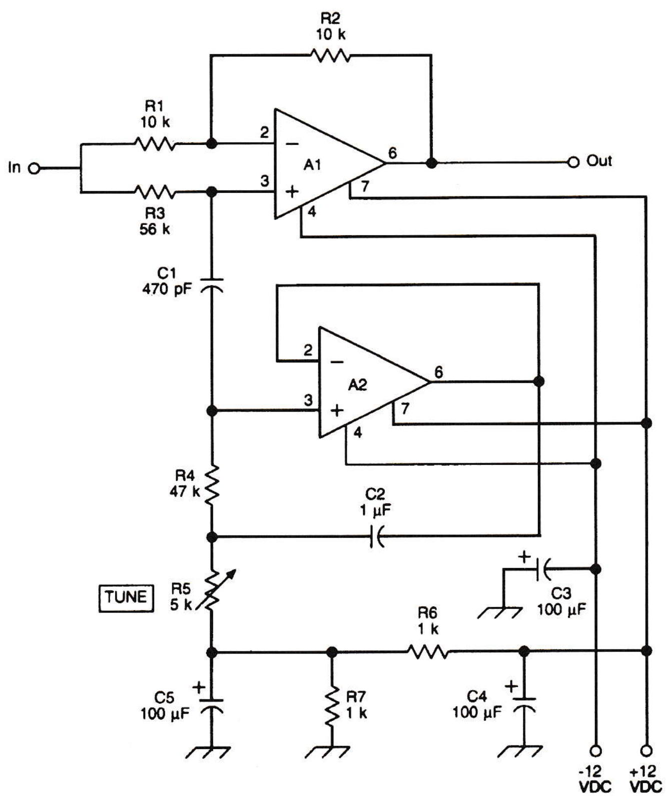
SUBSONIC FILTER



ELEKTOR ELECTRONICS *Fig. 35-11*

This filter is a fifth-order high-pass section that provides 1 dB attenuation at 20 Hz. Below that, however, the response drops off very steeply; the −3-dB point is at 17.3 Hz, and at 13.6 Hz, the attenuation is 10 dB. Note that it is important that C1 to C5 are within 1 percent of one another. Their individual tolerance is not so important because that merely affects the cutoff point. However, mutual deviation adversely affects the shape of the response, which should be a Butterworth characteristic as specified. All resistors are 1-percent types.

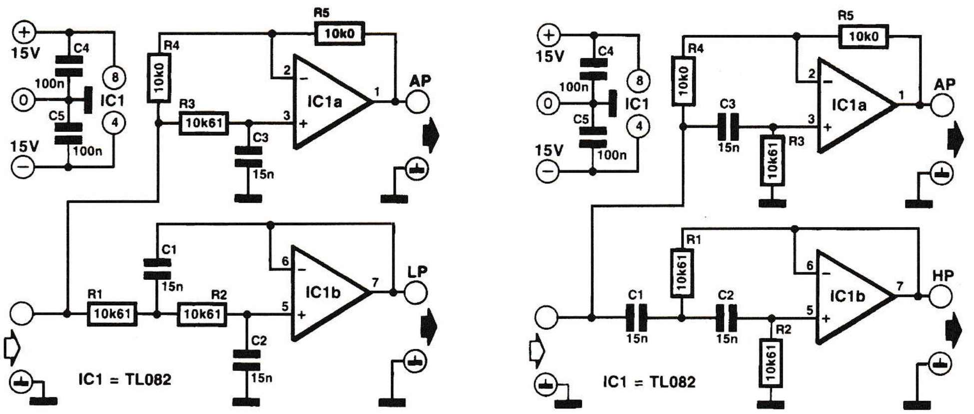
600- TO 3000-Hz TUNABLE NOTCH-FILTER CIRCUIT



73 AMATEUR RADIO TODAY ***Fig. 35-12***

This figure shows a notch filter that will tune roughly from 600 Hz to 3 kHz; it has been used by ham and SWL builders for a number of years. It is used for notching out unwanted CW stations or for notching out heterodynes in receiver outputs. Insert this filter between the headphone output of the receiver and a power amplifier stage.

LOW-PASS FILTER PHASE CORRECTOR

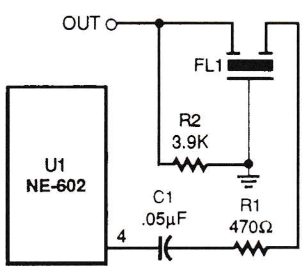


ELEKTOR ELECTRONICS

Fig. 35-13

In some applications, it might be desired, or even be essential, that the bandwidth of the audio signal be limited, but that the phase relationship with the original signal be retained. A surround-sound encoder is a good example of this. The requirement can be met by combining the low-pass filter with an all-pass section and having the filtered signal compared with the signal corrected by the all-pass network. As it happens, the phase transfer of a first-order all-pass filter is exactly the same as that of a second-order critically damped network. The design of such a combination is shown in Figure 1. In this, the all-pass network is based on IC1a and the low-pass section on IC1b. The −6-dB cutoff point is at exactly 1 kHz, and the −3-dB rolloff is at 642 Hz.

CERAMIC FILTER INTERFACING

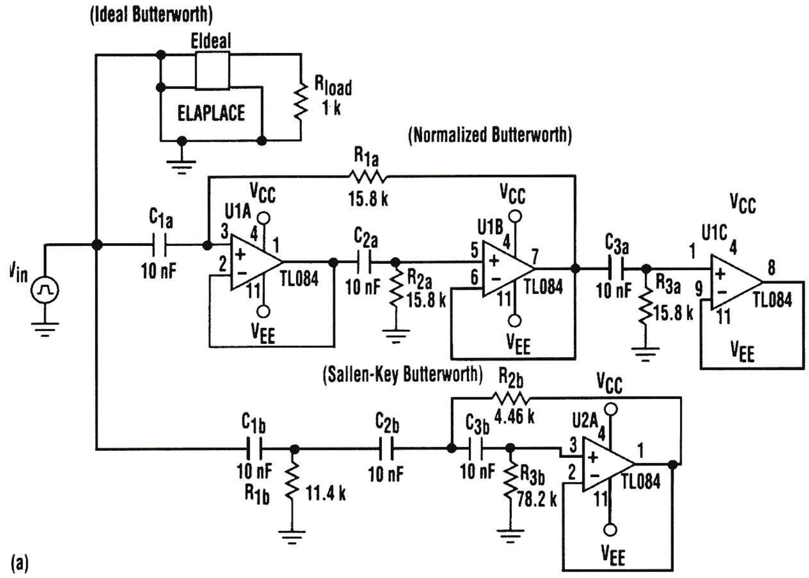


Ceramic or mechanical filters can be used to provide a frequency-selective output.

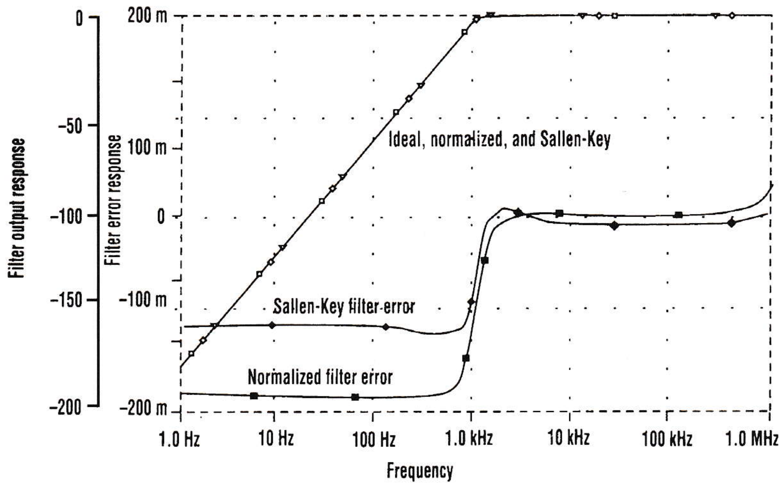
POPULAR ELECTRONICS

Fig. 35-14

ACTIVE HP FILTER



(a)



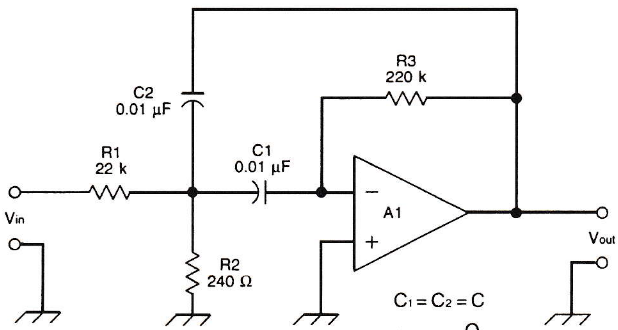
(b)

ELECTRONIC DESIGN

Fig. 35-15

Interchanging the resistors and capacitors transforms the normalized low-pass filter into a high-pass filter with the same corner frequency (*a*). Notice that the Sallen-Key filter must be modified according to impedance levels at each node. This yields a filter with equal-value capacitors and unequal-value resistors, an improvement over the traditional low-pass design of equal-value resistors and unequal-value capacitors. The graphs in (*b*) indicate that the normalized high-pass filter compares favorably with the Sallen-Key filter in high-pass applications.

MULTIPLE-FEEDBACK BANDPASS ACTIVE FILTER



$C_1 = C_2 = C$

$R_1 = \frac{Q}{2\pi f_c A_v C}$

$R_2 = \frac{Q}{2\pi f_c C(2Q^2 - A_v)}$

$R_3 = \frac{2Q}{2\pi f_c C}$

$\text{Gain} = A_v = \frac{R_3}{2R_1}$

73 AMATEUR RADIO TODAY ***Fig. 35-16***

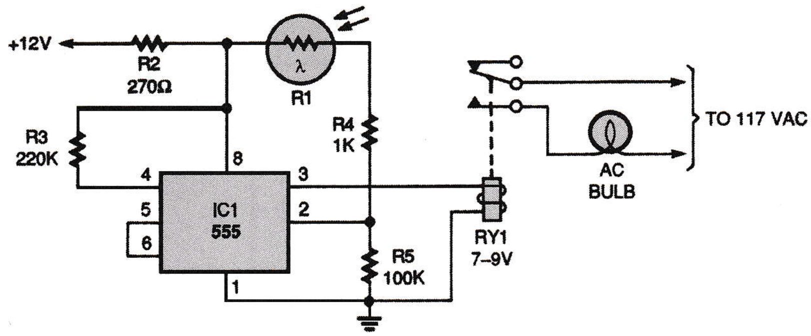
This is a multiple-feedback-path (MFP) bandpass filter.

36

Flasher Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

AC LIGHT FLASHER

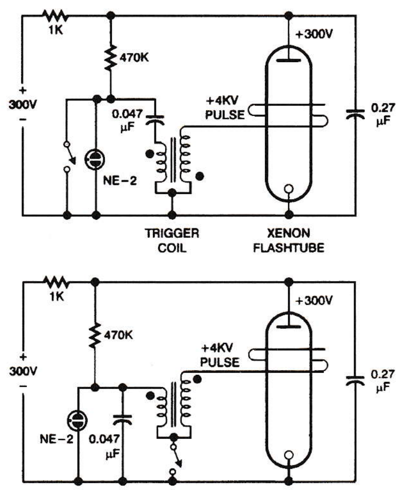


POPULAR ELECTRONICS

Fig. 36-1

This circuit uses a 555 timer that is configured in an unusual way in that trigger pin 2 is activated by the voltage from a photocell to sense when the lamp is lit. The relay time constants affect the flash rate.

XENON FLASHTUBE CIRCUIT

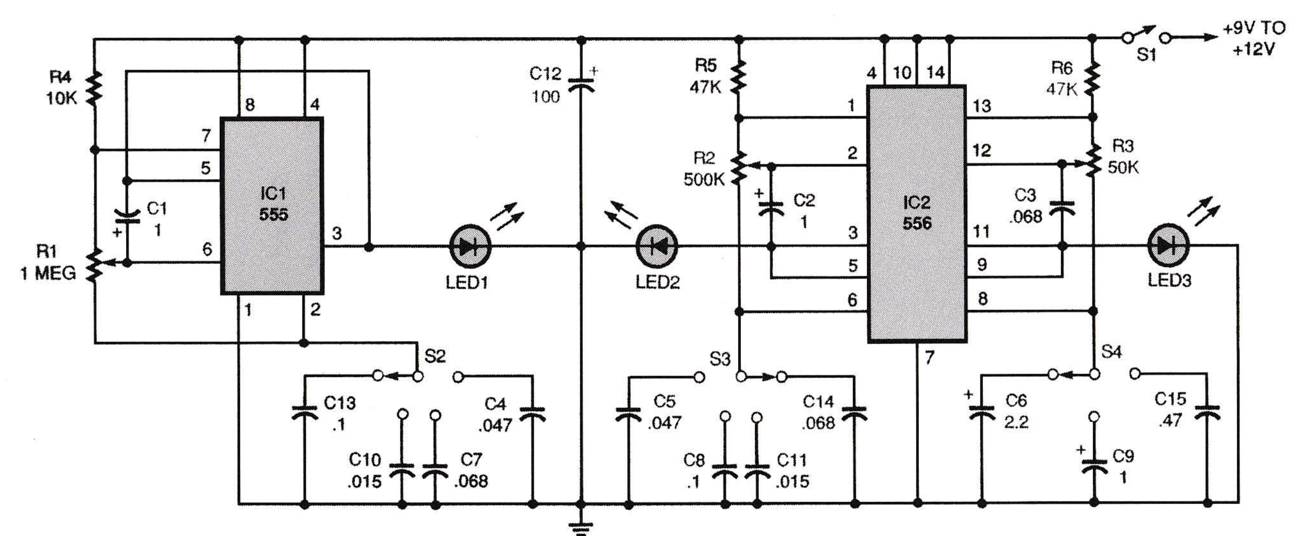


These are basic circuits for firing Xenon flashtubes. Which circuit is applicable depends on trigger coil polarity.

ELECTRONICS NOW

Fig. 36-2

NOVELTY LED FLASHER

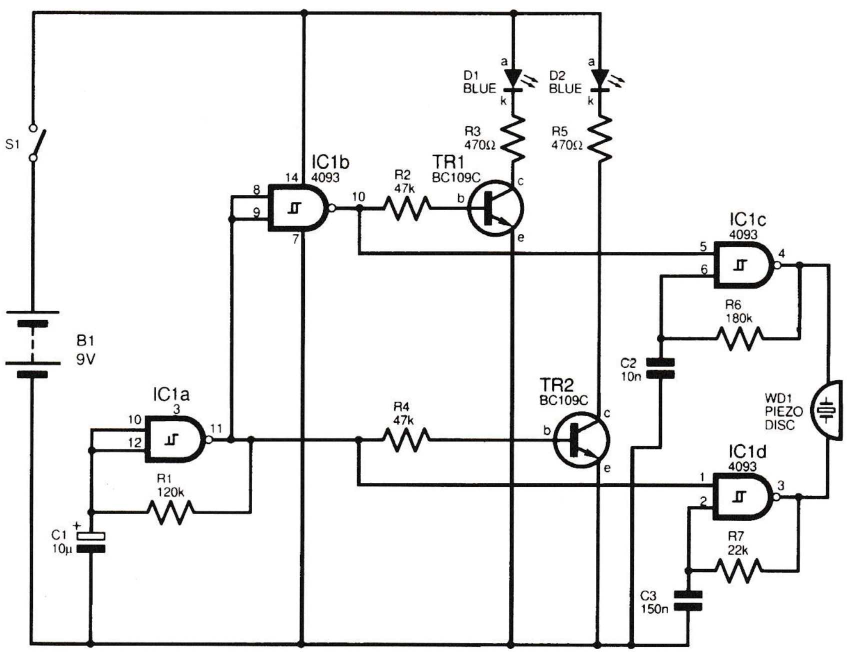


POPULAR ELECTRONICS

Fig. 36-3

The circuit contains three oscillators, one based on a 555 (IC1) and the others centered around a 556 (IC2). When S1 is on, each oscillator's LED (LED1, LED2, or LED3) will flash rhythmically, according to the setting of its potentiometer (R1, R2, or R3), and the capacitor selected by its switch (S2, S3, or S4). Each oscillator output is capable of driving up to 20 jumbo LEDs connected in parallel. Arrange the LEDs to suit your taste. The circuit will operate on a battery or any 9- to 12-Vdc source.

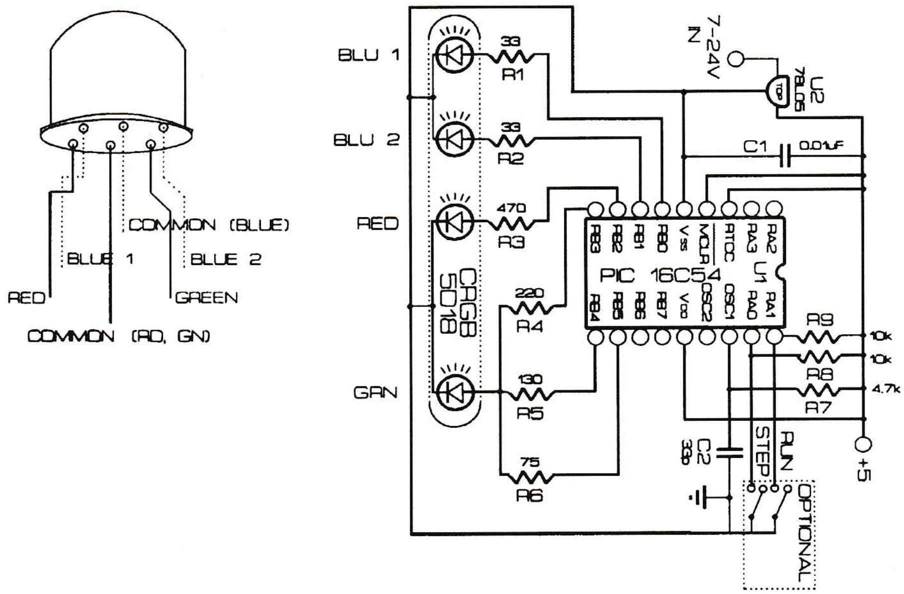
MODEL FIRE ENGINE FLASHER CIRCUIT



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 36-4***

This circuit was designed to add both a blue flashing light sequence and a two-tone siren effect to a model fire engine. IC1 is a quad NAND Schmitt trigger, and IC1a provides a low-frequency waveform, which is inverted by IC1b. Complementary outputs are obtained, which drive transistors TR1 and TR2. These driver transistors operate D1 and D2, which are two blue LEDs. The model had a blue translucent molding on its roof and yielded a surprisingly realistic effect. The outputs from IC1a and IC1b are also used to control gated astables IC1c and IC1d, which are set to oscillate at different frequencies to simulate a typical British two-tone siren, sounded by WD1, a piezo disk. Battery B1 should be an alkaline or rechargeable 9-V type because the current consumption is relatively high. S1 was a slide switch fitted on the rear of the vehicle. The sounder is optional.

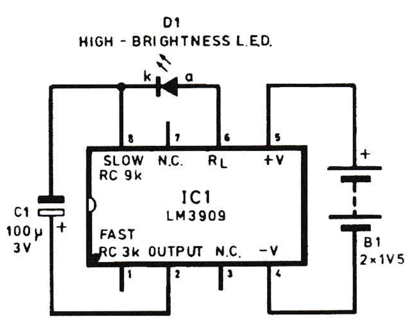
RAINBOW FLASHER



NUTS AND VOLTS ***Fig. 36-5***

This device flashes a multicolor RGB LED in various colors, in a sequence and speed determined by the PIC 16C54 software. The LED is by Cree Electronics.

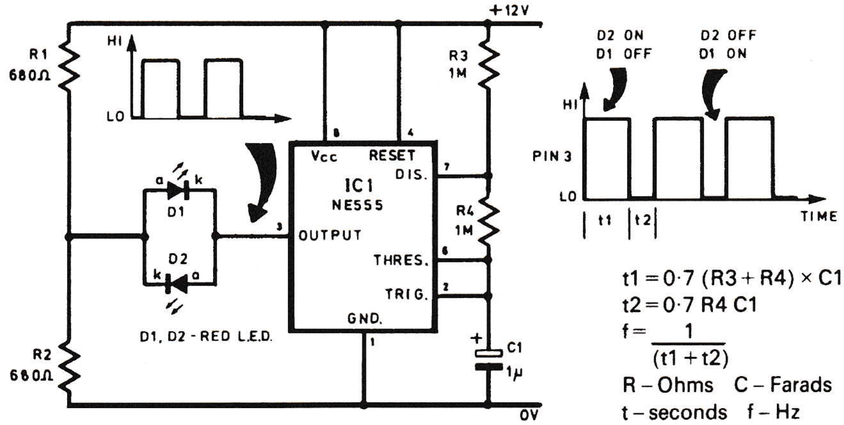
SINGLE-LED FLASHER



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 36-6***

This is a circuit diagram of an LED flasher with few components.

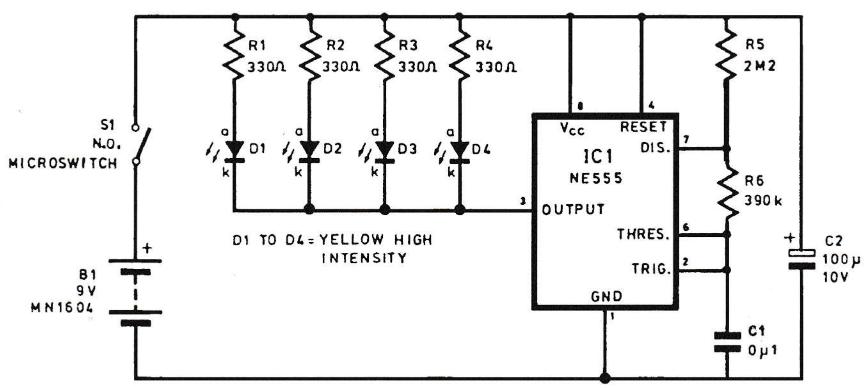
12-V LED FLASHER



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 36-7***

This is the circuit diagram of an LED flasher suitable for 12-V operation. The timing formulas are also shown.

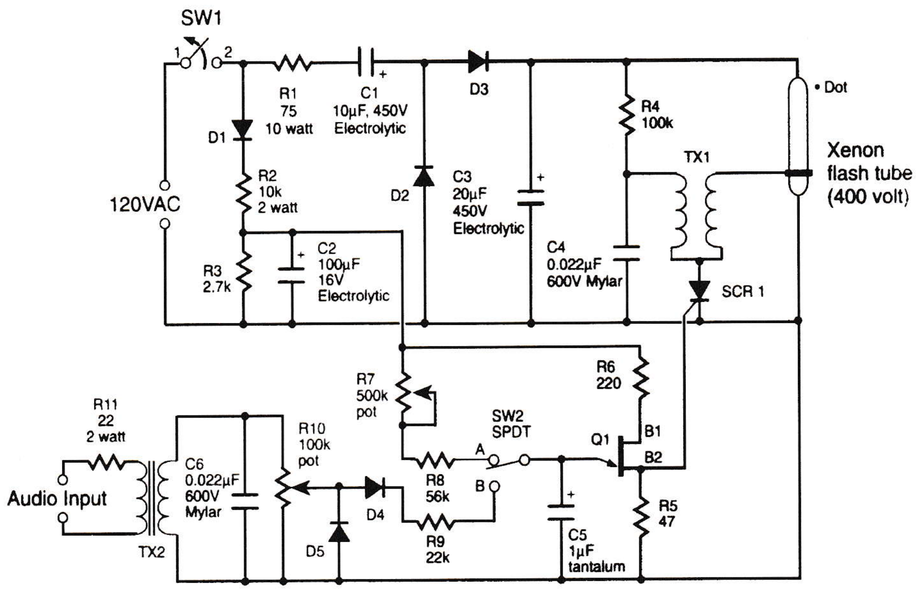
MODEL AIRCRAFT LED FLASHER



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 36-8***

This circuit was used to simulate machine gun firing on a model aircraft, but it can also be used for other applications.

AUDIO-DRIVEN XENON TUBE FLASH CIRCUIT



R1 75 ohm (10 watts)*
R2 10K ohm (2 watts)
R3 2.7K ohm
R4 100K ohm
R5 47 ohm
R6 220 ohm
R7 500K ohm POT
R8 56K ohm
R9 22K ohm
R10 100K ohm POT
R11 22 ohm (2 watts)
C1 10uF, 450 volt electrolytic
C2 100uF, 16 volt electolytic
C3 20uF, 450 volt electrolytic
C4, C6 0.022uF, 600 volt mylar
C5 1uF, 35 volt tantalum

Q1 2N4891 unijunction transistor or equiv.
SCR1 400 volt, 3 amp (or better) SCR
TX1 4000 volt trigger transformer
TX2 12.6 volt C.T. 100ma power transformer
SW1 SPST switch (power switch)
SW2 SPDT switch
D1, D2, D3 400 volt, 1 amp or better power rectifier diodes
D4, D5 1N934 or other small signal diode
FT Xenon flashtube (straight or horseshoe) in the 200 to 400 volt range

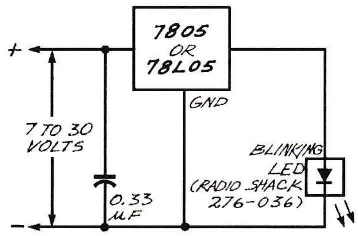
NUTS AND VOLTS

Fig. 36-9

This circuit uses a Xenon flashtube to create a very real lightning effect. It can be driven in one of two ways: by sound input or by straight repeated flashes, depending on where switch 2 is positioned. When producing straight flashes, the rate is determined by pot R7. If the switch is set to be triggered by sound, the sensitivity is determined by the level of the sound driving the transformer

TX2 and the setting of pot R10. This circuit uses high voltages (over 4000 V for the trigger), and extreme care should be taken by anyone that builds it. The high-voltage capacitors can hold dangerous voltages for some time (even after the circuit has been unplugged). Strong caution concerning part polarity is highly urged.

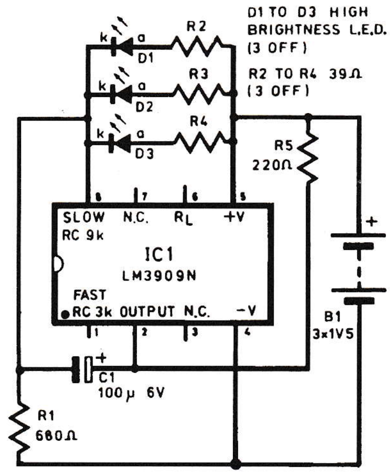
12-V LED FLASHER



ELECTRONICS NOW *Fig. 36-10*

The figure shows a 12-V LED flasher configuration.

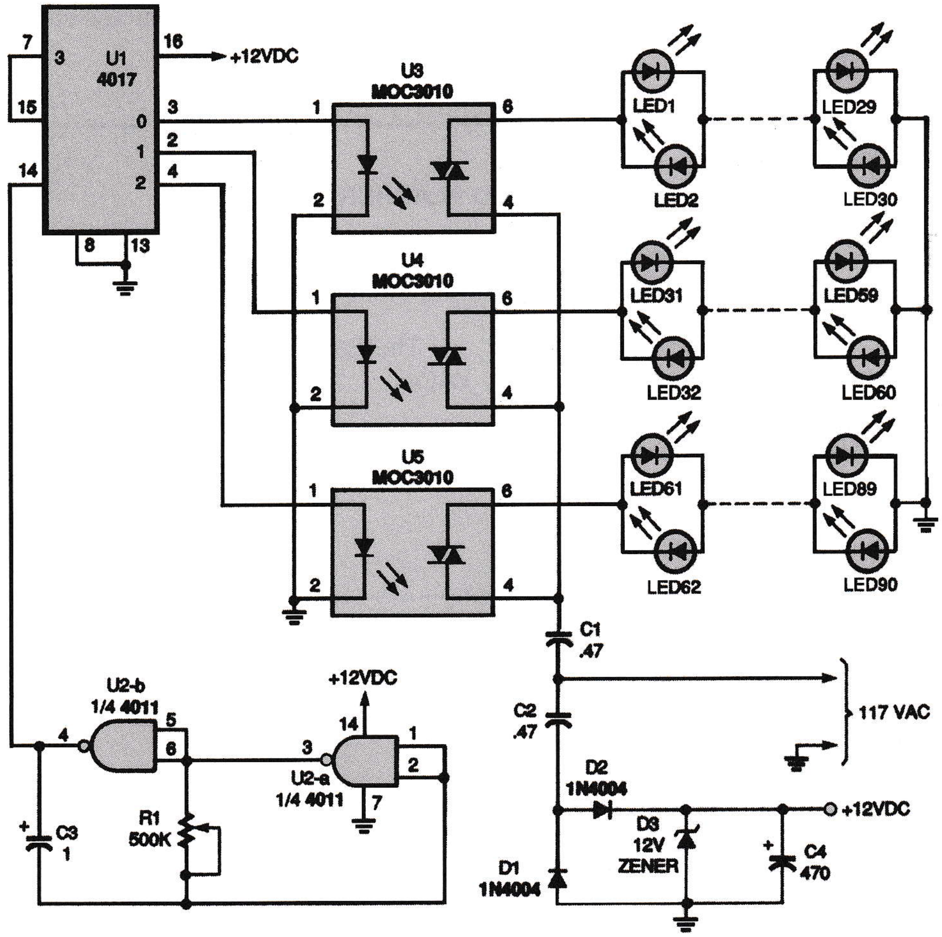
MULTIPLE-LED FLASHER



EVERYDAY PRACTICAL ELECTRONICS *Fig. 36-11*

This is a circuit diagram for driving multiple LEDs from a single LM3909 flasher chip.

MULTIPLE FLASHING LED LIGHT STRING

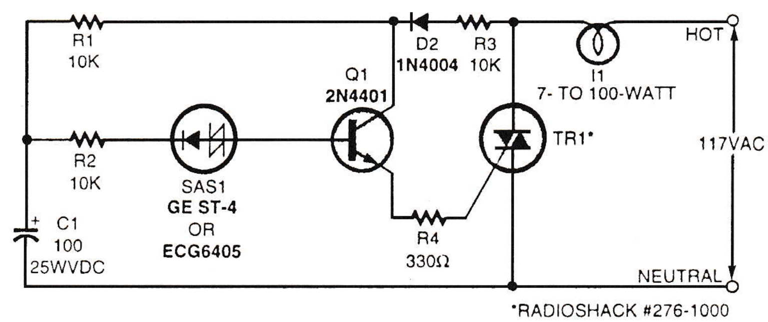


POPULAR ELECTRONICS

Fig. 36-12

This circuit offers up to 10 strings of LEDs that turn on one row at a time in a sequential manner; for the sake of space and simplicity, only three strings are shown. The LED pairs could be all one color, mixed red and green, or any combination of colors. C2, C4, and D1 to D3 make up a 12-Vdc power supply for the five ICs. Two gates of a 4011 quad two-input NAND gate, U2-a and U2-b, are connected in a low-frequency oscillator circuit. Resistor R1 controls the oscillator's frequency. The oscillator supplies a clock output to U1, a 4017 decade counter/divider, which operates as a counter that can be programmed to count from 0 to 9. With each clock pulse, the 4017 makes a single step up in count from 0. The first output, at pin 3, supplies voltage to the LED in U3, a 3010 optocoupler, turning on the IC's triac and lighting the first string of LEDs. The next clock pulse steps the 4017 to the next output, at pin 2, and repeats the sequence for that string. As each new light string turns on, the preceding string turns off, giving the effect of a climbing string of lights. Changing the setting of R1 will change the rate at which the strings turn on and off. If you want to add additional light strings beyond what is shown in the figure, simply duplicate the circuitry for the light strings as shown and connect to the appropriate output of the 4017, U1. You will also have to connect the first unused output to pin 15. If you use the maximum of 10 strings, pin 15 should be connected to ground.

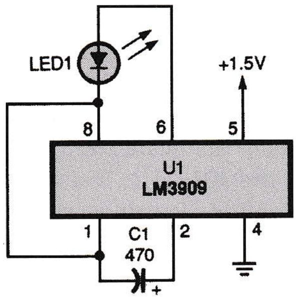
LIGHT FLASHER



POPULAR ELECTRONICS ***Fig. 36-13***

Shown here is a simple flasher circuit. With the component values shown, the flash rate is approximately once per second. The incandescent-lamp load glows at half brightness for about one-third of the total flasher period and is off for the remaining two-thirds. Electrolytic capacitor C1 charges during the positive half-cycle of the ac waveform through R1, R3, and D2. When the voltage across the capacitor reaches the break-over voltage of the silicon asymmetrical switch (SAS1), the capacitor starts to discharge through R2, SAS1, Q1, R4, and the triac. Emitter follower Q1 is driven by the discharge current from C1, and it, in turn, provides gate drive for the triac. Thus, the triac conducts and the light glows while C1 discharges. The lamp goes dark when C1 is depleted of charge and remains dark until the ac power waveform goes positive again and charges the capacitor sufficiently. The triac should be triggered into conduction by a gate current of no more than 5 mA. The flash rate can be varied by changing the value of capacitor C1. Using more capacitance results in a slower flash rate, and less capacitance results in a faster flash rate.

1.5-V LED FLASHER

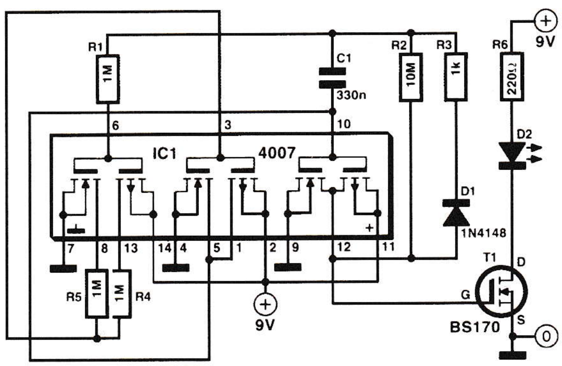


The LM3909 is one of the easiest-to-use LED flashers around. It can run for years using a standard D cell.

POPULAR ELECTRONICS

Fig. 36-14

THRIFTY LED FLASHER

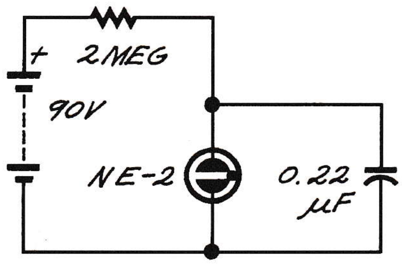


ELEKTOR ELECTRONICS

Fig. 36-15

The dual complementary pair of switching FETs and inverter contained in a CD4007 CMOS IC enables an LED flasher to be made that uses very little energy. The IC is arranged as a three-inverter oscillator. Resistors R4 and R5 in series with the drains of one pair of FETs ensure that the drive current for the following pair of FETs is tiny. The high time of the oscillator is determined by network R3-C1, and its low time by R2-C1 (D1 is then cut off so that R3 is inactive). The LED is provided with current during the high time of the oscillator by T1. The level of this current is determined by R6. The values of R_2, R_3, and C_1 cause an LED OFF time of 1 s and an ON time of 1 ms. Because the high-efficiency diode draws a current of 30 mA, its lighting will be clearly visible. A standard 9-V battery will give continuous operation for about three years.

NEON LAMP FLASHER

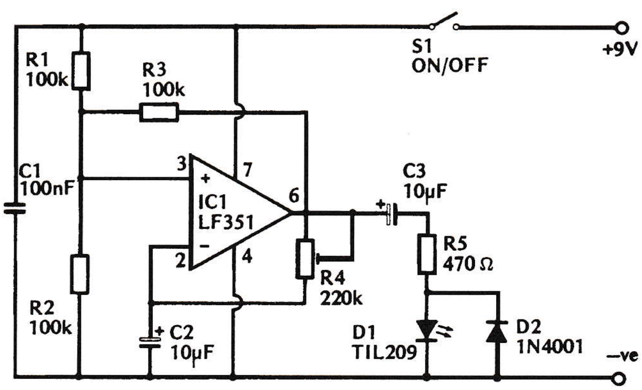


The neon lamp has negative dynamic resistance—the voltage across it falls while conduction is increasing. As a result, it flashes on and off.

ELECTRONICS NOW

Fig. 36-16

1-S FLASHER

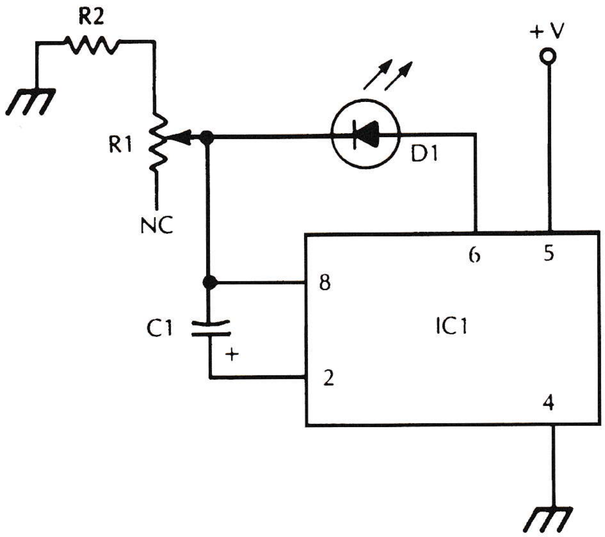


ELECTRONIC EXPERIMENTERS HANDBOOK

Fig. 36-17

The circuit diagram for the 1-s flasher is based on an operational amplifier that is biased by R1, R2, and R3 to act as a form of Schmitt trigger. The output goes to the low state if the inverting input is taken above $\frac{2}{3}$V+, and high if it is taken below $\frac{1}{3}$V+. The output, therefore, goes high initially, but C2 soon charges to $\frac{2}{3}$V+ via R4, and then the output, goes low. Capacitor C2 then discharges to $\frac{1}{3}$V+ via R4, sending the output high again, and producing continuous oscillation. Resistor R4 is adjusted to give an operating frequency of 1 Hz. The 1-s flasher can be calibrated against a watch or clock with a second hand by empirical means. The output of IC1 is coupled to the LED indicator, D1, by way of dc blocking capacitor C3 and current-limiting resistor R5, and the LED is briefly pulsed on as the output voltage swings positive. Diode D2 ensures that there is both a charge and a discharge path for C3 so that the output signal is properly coupled to D1. The current consumption of the unit is about 2 mA.

VARIABLE-FREQUENCY LED FLASHER



IC1	LM3909 LED flasher/oscillator IC
D1	LED
C1	330 μF 5 V electrolytic capacitor
R1	10 kΩ potentiometer
R2	330 Ω ¼ W 5% resistor

TAB BOOKS ***Fig. 36-18***

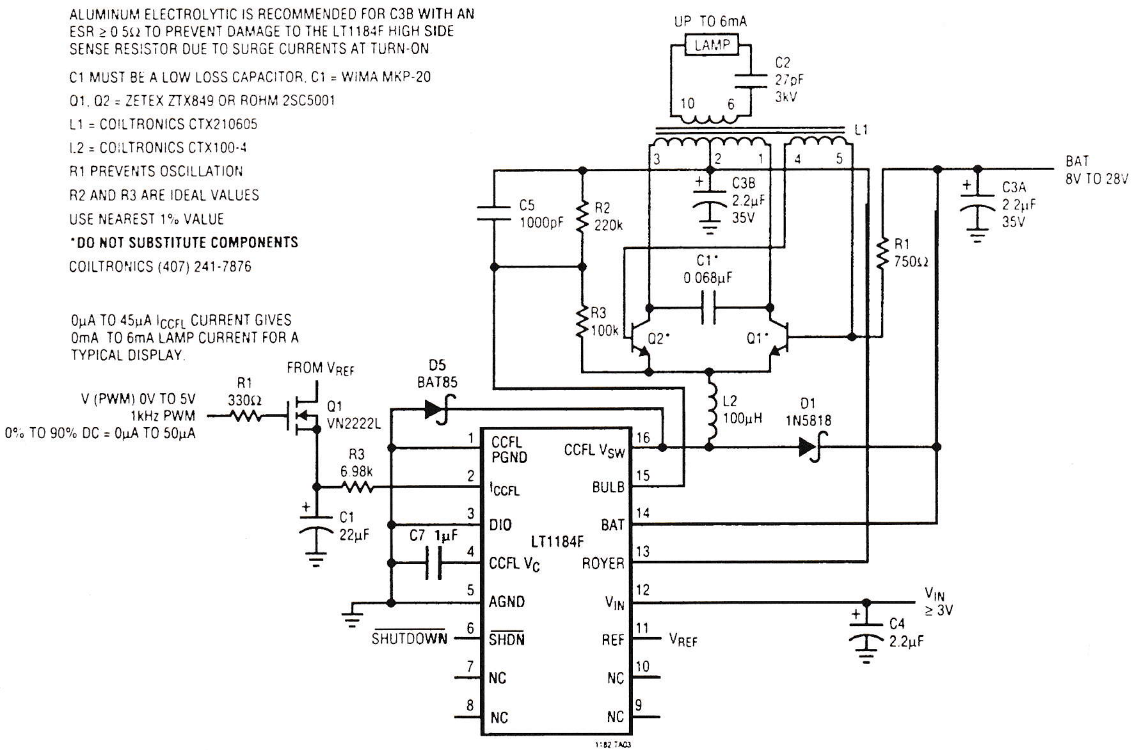
R1 varies the flash rate, and R2 limits the minimum resistance in the potentiometer circuit to prevent damage to D1 or IC1.

37

Fluorescent Lamp Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

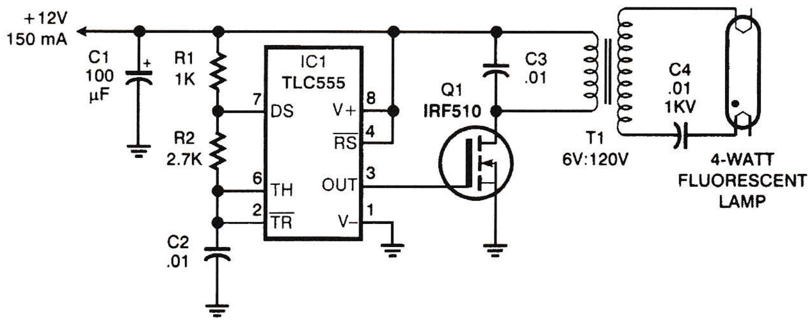
COLD-CATHODE FLUORESCENT LAMP DRIVER



LINEAR TECHNOLOGY POWER SOLUTIONS *Fig. 37-1*

In this circuit, the lamp is driven sinusoidally, minimizing RF emissions in sensitive portable applications. Lamp intensity is controlled smoothly from zero to full brightness with no hysteresis or "pop on." This floating bulb circuit configuration extends the illumination range for the bulb because parasitic bulb-to-display-frame capacitive losses are minimized. The feedback signal is generated by monitoring the primary-side Royer converter current between the BAT and Royer pins. The LT1184F current-mode switching regulator and L2 provide an average current to Q1 and Q2, which form a Royer-class converter along with L1 and C1. The lamp is driven by L1's secondary. Feedback to the LT1184F is provided on the primary side of L1 for floating bulb configurations, whereas feedback in the grounded configuration is provided by sensing one-half of the average bulb current. The oscillator frequency is 200 kHz, which minimizes the size of the required magnetics.

BATTERY-OPERATED FLUORESCENT LIGHT



ELECTRONICS NOW ***Fig. 37-2***

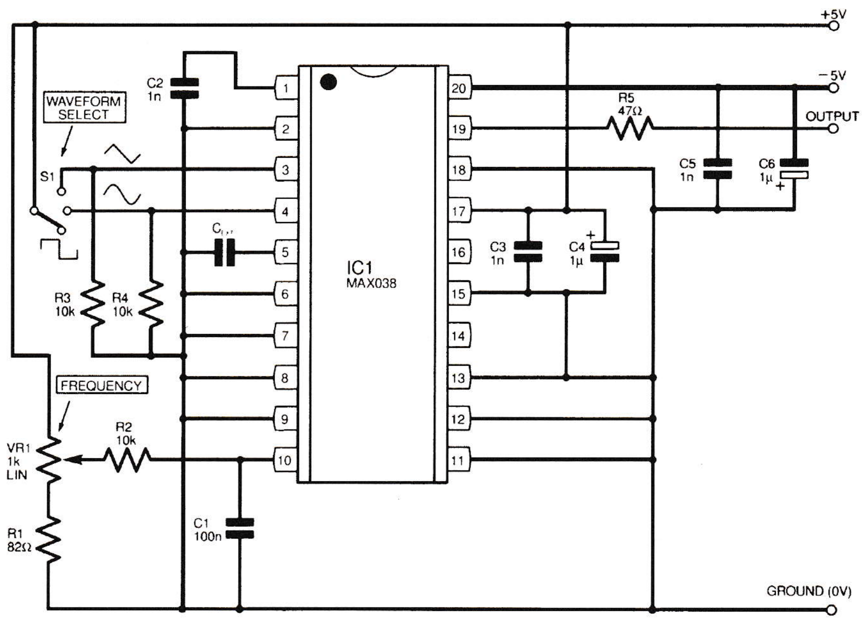
The circuit consists of a 20-kHz oscillator, a switching transistor to amplify its output, and a step-up transformer. A 120- to 6-V power transformer is connected backward, using half of the 12-V side for 6 V. In the circuit, the transformer is working at considerably more than its rated voltage, but the high frequency keeps it from saturating. Although the lamp does not glow at full brightness, the circuit is energy-efficient, requiring only 150 mA.

38

Function-Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

MAX038 HIGH-SPEED FUNCTION-GENERATOR CIRCUIT

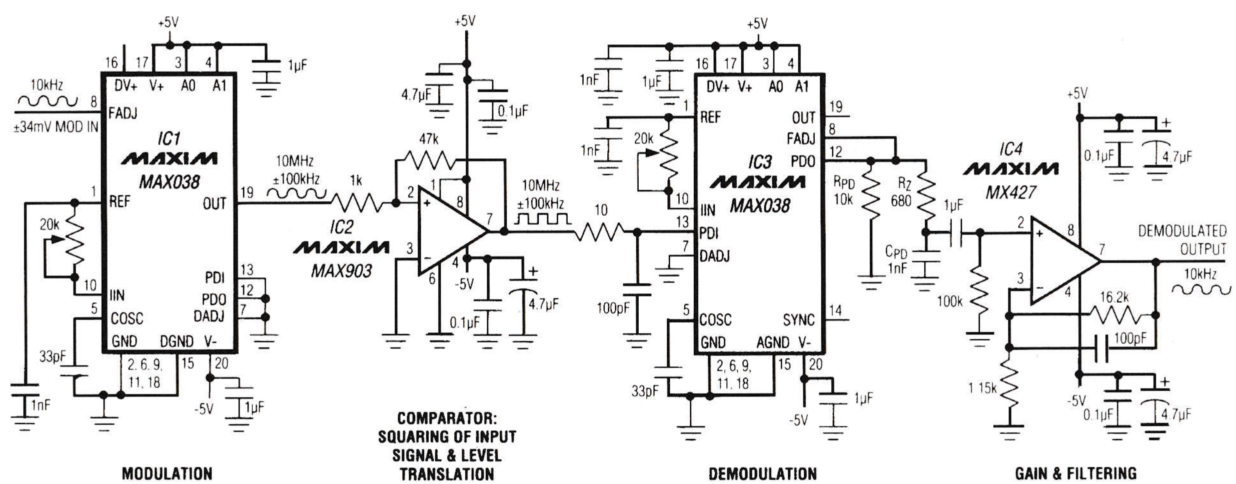


EVERYDAY PRACTICAL ELECTRONICS ***Fig. 38-1***

For enthusiasts who would like to experiment with this device for themselves, a simple circuit to get it up and running is shown. The only real problem with the MAX038 is inherent in its sheer speed. Maxim suggests careful layout on a double-sided ground-plane PC board for best results. In practice, a single-sided PC board seems to work well, provided it has plenty of copper areas at ground potential, good decoupling, and guard tracks around signal paths carrying rectangular waveforms, especially that from the sync output. The following component recommendations are offered:

Resistors—0.6 W, 1-percent metal film
C1, C2, C3, and C5—ceramic disk
C4 and C6—tantalum bead, 35 V
C_{ext}—as required for frequency, between 47pF and 47μF, polyester or polystyrene

FM DEMODULATOR

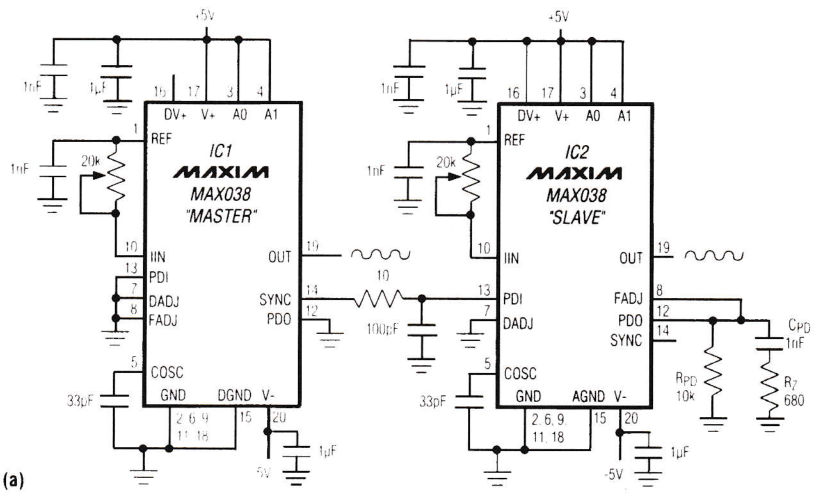


MAXIM ENGINEERING JOURNAL

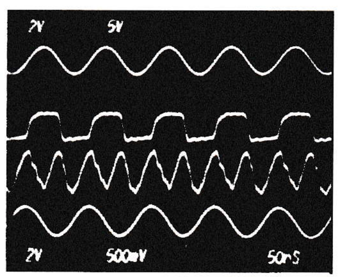
Fig. 38-2

The frequencies at IC3's phase-detector output are the sum and difference of the frequencies at PDI and OUT. Thus, with appropriate cutoff frequency and gain, the low-pass filter (IC4) passes only the original 10-kHz signal to the demodulated output. The pole for this filter is set by the 16.2-kΩ and 100-pF components. The frequency response for IC3's PLL is set by R_{PD}, C_{PD}, and R_Z. When the loop is in lock, PDI is in approximate phase quadrature with the output signal. Also, when in lock, the duty cycle at PDO is 50 percent, and PDO's average output current is 250 μA. The current sink at FADJ demands a constant 250 μA, so PDO outputs above and below that level develop a bipolar error voltage across R_{PD} that drives the FADJ voltage input. *Note:* The MAX038's internal phase detector is a phase-only detector, producing a PLL whose frequency-capture range is limited by the bandwidth of its loop filter. For wider-range applications, consider an external phase-frequency detector.

SYNCHRONIZED MAX038 FUNCTION GENERATORS



(a)



A - MASTER OUT, 2V/div
B - PDI, 5V/div
C - PDO, 500mV/div
D - SLAVE OUT, 2V/div
TIMEBASE - 50ns/div

(b)

MAXIM ENGINEERING JOURNAL ***Fig. 38-3***

The MAX038's internal phase detector is intended primarily for use in phase-locked-loop (PLL) configurations. In (*a*), for example, the phase detector in IC2 enables that device to synchronize its operation with that of IC1. You connect the applied reference signal to IC2's TTL/CMOS-compatible phase-detector input (PDI) and connect the phase-detector output (PDO) to the input (FADJ) of the internal voltage-controlled oscillator. PDO is the output of an exclusive-OR gate—a mixer—which produces rectangular current waveforms at frequencies equal to the sum and difference of the PDI frequency and the MAX038 output frequency. These waveforms are integrated by C_{PD} to form a triangle-wave voltage output at PDO (*b*). The 10-Ω/100-pF pair at PDI limits that pin's rate of rise to 10 ns.

BENCHTOP FUNCTION GENERATOR WITH BUILT-IN COUNTER

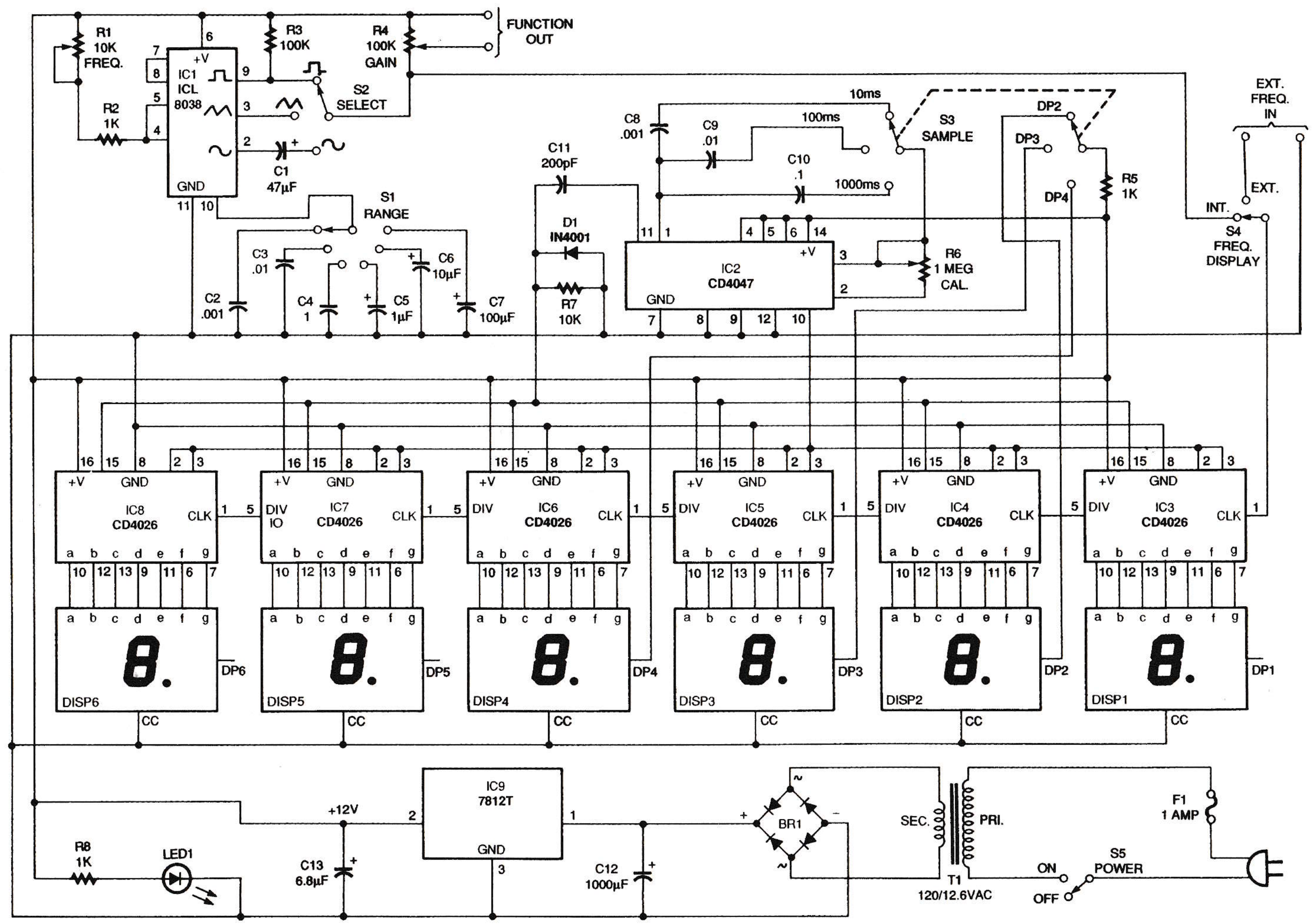
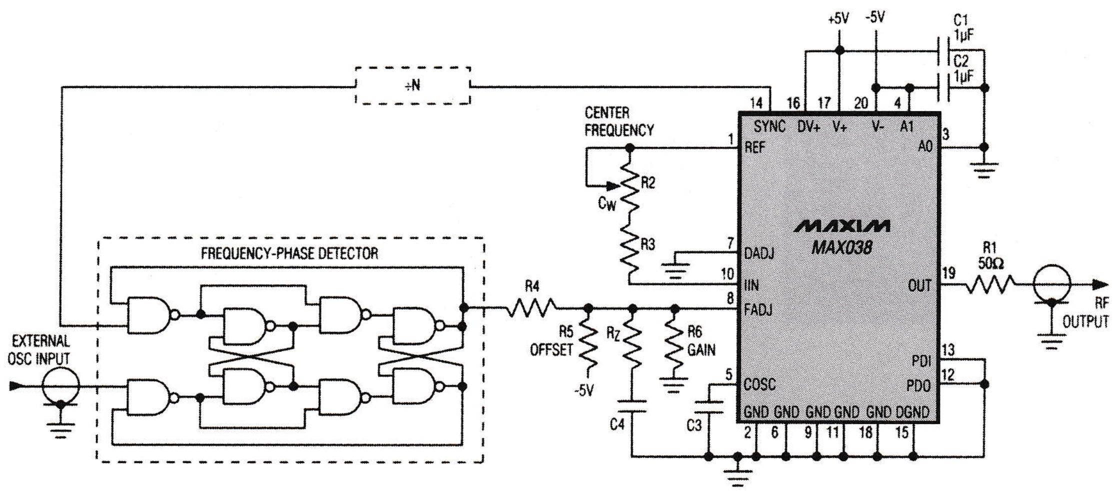


Fig. 38-4

BENCHTOP FUNCTION GENERATOR WITH BUILT-IN COUNTER *(Cont.)*

This circuit will produce sine, square, and triangle waves from 0.1 Hz to 1 MHz and has a counter which will read the frequency of the function generator or an external signal of a few volts peak-to-peak that will drive the CMOS counter.

PLL WITH DIVIDE-BY-*N* CIRCUIT

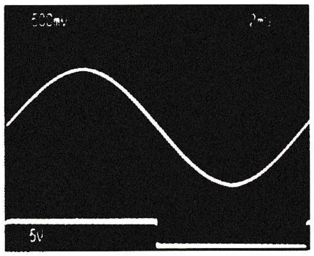


MAXIM ENGINEERING JOURNAL ***Fig. 38-5***

The MAX038 function generator IC can be used in a PLL system. To gain the advantages of a wider capture range and an optional divide-by-*N* circuit (which allows the PLL to lock onto arbitrary multiples of the applied frequency), you can introduce an external frequency-phase detector, such as the 74HC4046 or the discrete-gate version shown. Unlike phase detectors, which can lock to harmonics of the applied signal, the frequency-phase detector locks only to the fundamental. In the absence of an applied frequency, its output assumes a positive dc voltage (logic 1) that drives the RF output to the lower end of its range as determined by resistors R4 to R6. These resistors also determine the frequency range over which the PLL can achieve lock. Again, R4 to R6, C4, and R_Z determine the PLL's dynamic performance.

MAX038 FUNCTION GENERATOR

SINE-WAVE OUTPUT (50Hz)

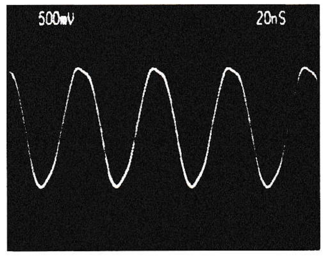


(a)

I_{IN} = 50μA
C_F = 1μF

TOP: OUTPUT 50Hz = F_0
BOTTOM: SYNC

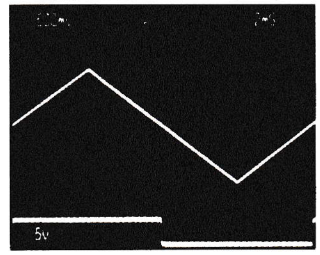
SINE-WAVE OUTPUT (20MHz)



(b)

I_{IN} = 400μA
C_F = 20pF

TRIANGLE-WAVE OUTPUT (50Hz)

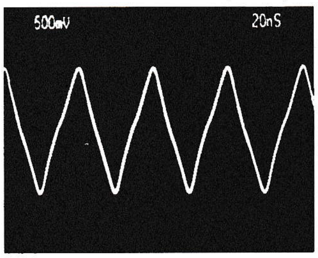


(c)

I_{IN} = 50μA
C_F = 1μF

TOP: OUTPUT 50Hz = F_0
BOTTOM: SYNC

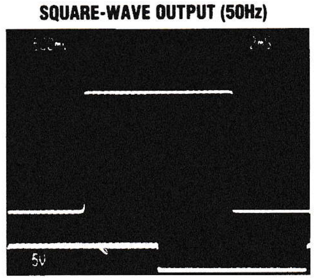
TRIANGLE-WAVE OUTPUT (20MHz)



(d)

I_{IN} = 400μA
C_F = 20pF

SQUARE-WAVE OUTPUT (50Hz)

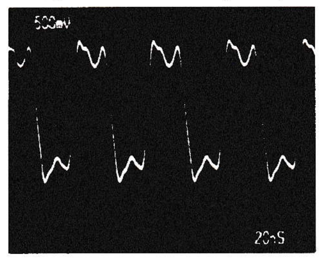


(e)

I_{IN} = 50μA
C_F = 1μF

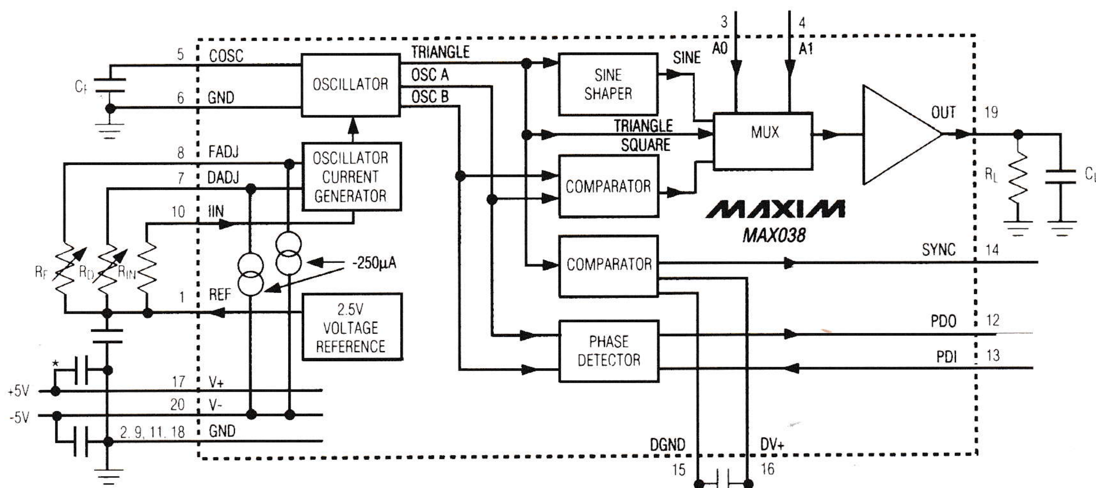
TOP: OUTPUT 50Hz = F_0
BOTTOM: SYNC

SQUARE-WAVE OUTPUT (20MHz)



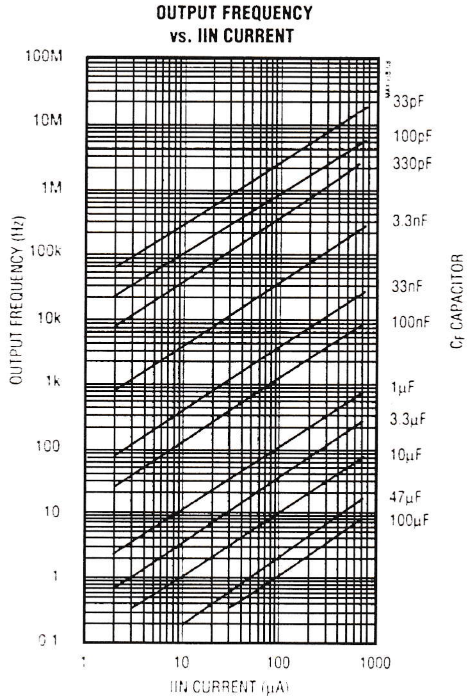
(f)

I_{IN} = 400μA
C_F = 20pF



→ = SIGNAL DIRECTION, NOT POLARITY

★ = BYPASS CAPACITORS ARE 1μF CERAMIC OR 1μF ELECTROLYTIC IN PARALLEL WITH 1nF CERAMIC.



MAXIM ENGINEERING JOURNAL ***Fig. 38-6***

The MAX038 is a precision, high-frequency function generator that produces accurate sine, square, triangle, sawtooth, and pulse waveforms with a minimum of external components. The internal 2.5-V reference (plus an external capacitor and potentiometer) lets you vary the signal frequency from 0.1 Hz to 20 MHz. An applied ±2.3-V control signal varies the duty cycle between 10 and 90 percent, enabling the generation of sawtooth waveforms and pulse-width modulation. A second frequency-control input—used primarily as a VCO input in phase-locked-loop applications—provides ±70 percent of fine control. This capability also enables the generation of frequency sweeps and frequency modulation. The frequency and duty-cycle controls have minimal interaction with each other. All output amplitudes are 2 $V_{p\text{-}p}$, symmetrical about ground. The low-impedance output terminal delivers as much as ±20 mA, and a two-bit code applied to the TTL-compatible A0 and A1 inputs selects the sine, square, or triangle output waveform:

A0	A1	WAVEFORM
X	1	Sine wave
0	0	Square wave
1	0	Triangle wave

X=Don't Care.

FREQUENCY SYNTHESIZER

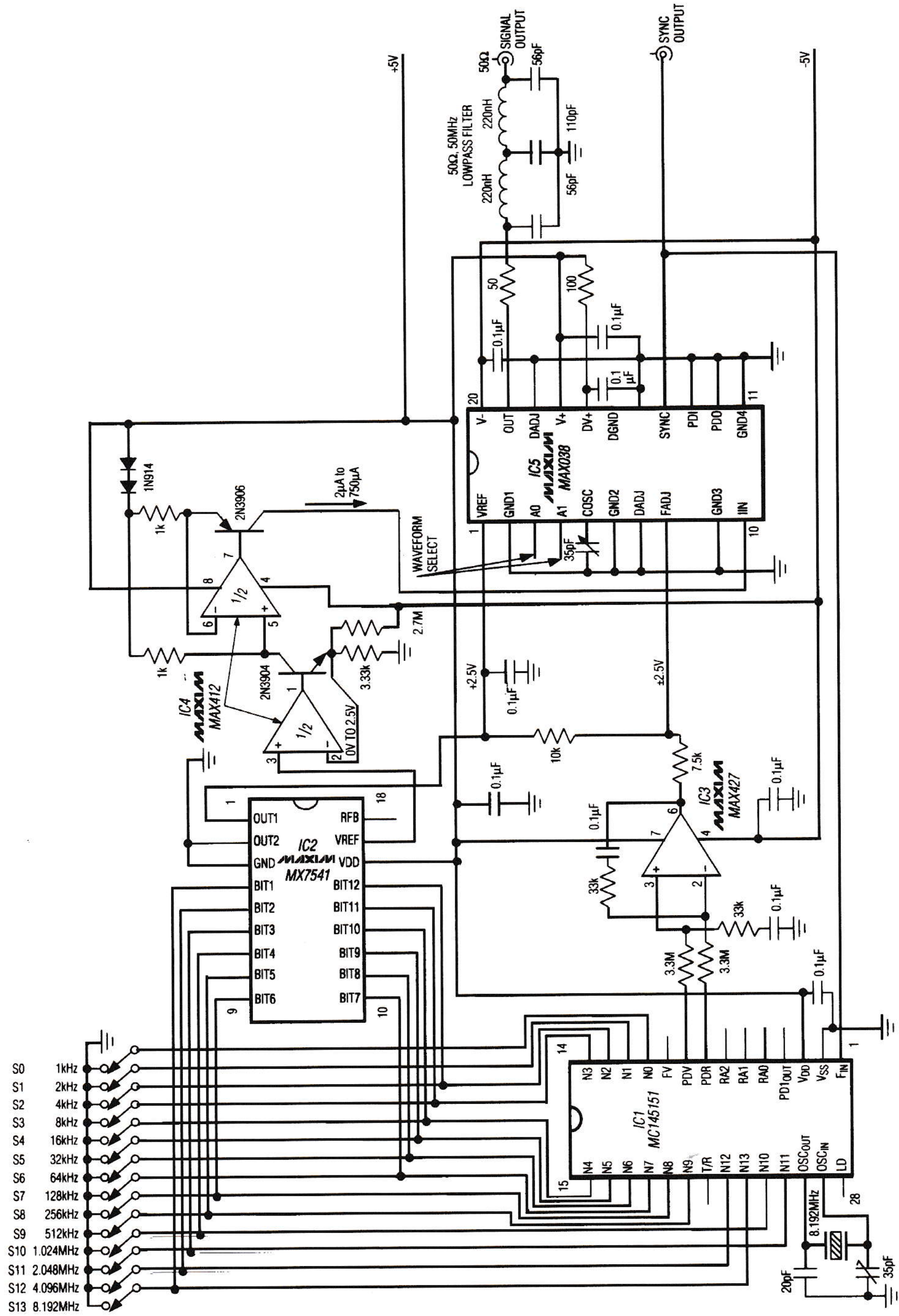
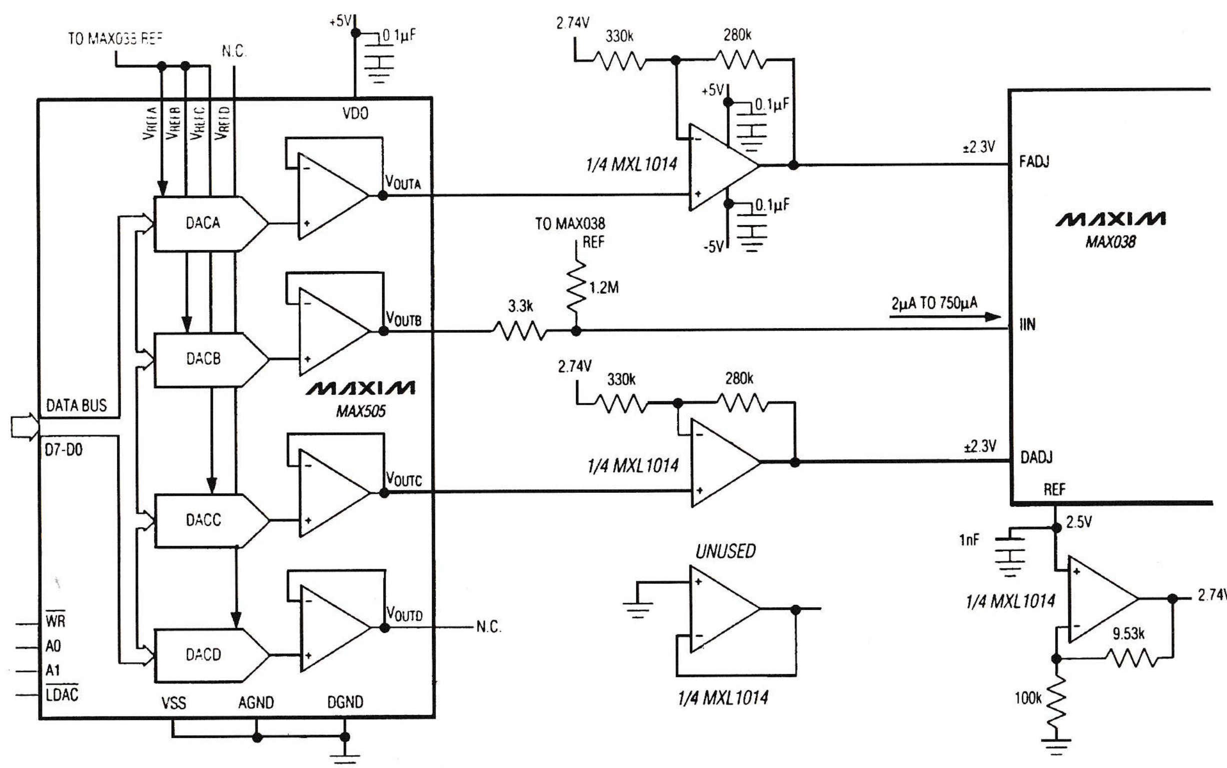


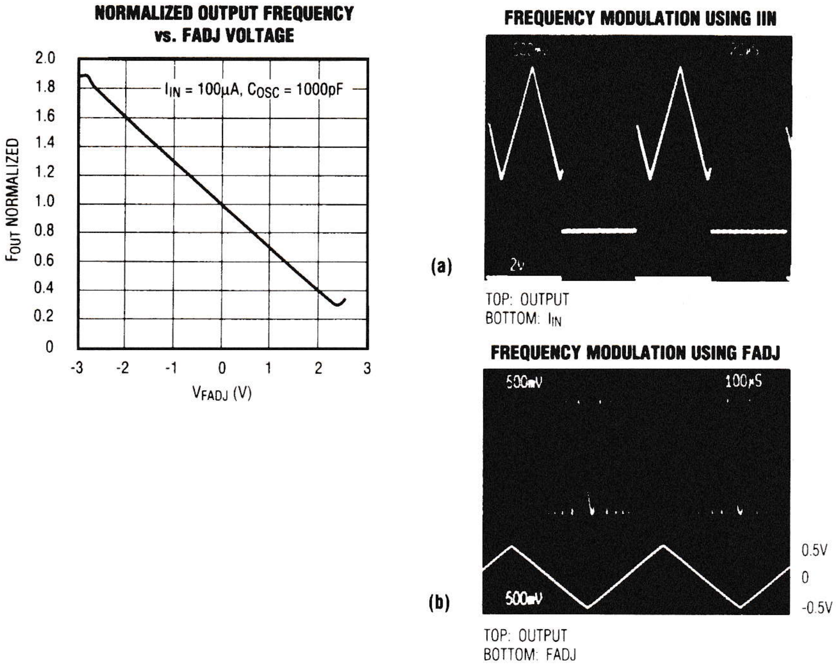
Fig. 38-7

FREQUENCY SYNTHESIZER *(Cont.)*

The MAX038 and four other ICs can form a crystal-controlled, digitally programmed frequency synthesizer that produces accurate sine, square, or triangle waves in 1-kHz increments over the range 8 kHz to 16.383 MHz. Each of the 14 manual switches (when open) makes the listed contribution to output frequency: Opening only S0, S1, and S8, for example, produces an output of 259 kHz. The switches generate a 14-bit digital word that is applied in parallel to the D/A converter (IC2) and a $\div N$ circuit in IC1. IC1 also includes a crystal-controlled oscillator and high-speed phase detector, which form a phase-locked loop with the voltage-controlled oscillator in IC5. The DAC and dual op amp (IC4) produce a 2- to 750-μA current that forces a coarse setting of the IC5 output frequency—sufficient to bring it within capture range of the PLL. This loop, in which the phase detector in IC1 compares IC5's SYNC output with the crystal-oscillator frequency divided by N, produces differential-phase information at PDV and PDR. IC3 then filters and converts this information to a ±2.5-V single-ended signal, which, when summed with an offset and applied to FADJ, forces the signal output frequency to the exact value set by the switches.

MAX038 FUNCTION GENERATOR DIGITAL CONTROL

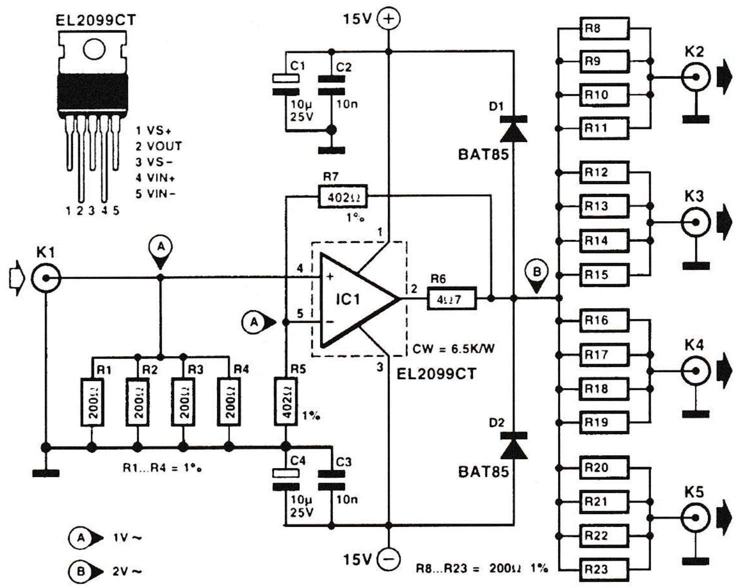




MAXIM ENGINEERING JOURNAL ***Fig. 38-8***

To adjust the frequency digitally, connect a voltage-output DAC to IIN via a series resistor, as shown. The converter output ranges from 0 V at zero to 2.5(255/256) V at full scale. Current injected by the converter into IIN, therefore, ranges from 0 to 748 µA. The 2.5-V reference and 1.2-Ω resistor inject a constant 2 µA, so (by superposition) the net current into IIN ranges from 2 µA (at a code of 0000 0000) to 750 µA (at 1111 1111). The quad DAC IC operates from 5 V or ±5 V. As described below, it can also provide digital control of FADJ and DADJ. For fine adjustments (±70 percent), apply a control voltage in the range ±2.3 V to the frequency adjust (FADJ) terminal. Both FADJ and IIN have wide bandwidths that allow the output frequency to be modulated at a maximum rate of about 2 MHz. As the more linear input, IIN is preferred for open-loop frequency use in a phase-locked loop. For digital control of FADJ, configure a DAC and external op amp (as shown in the figure) to produce an output ranging from −2.3 V (0000 0000) to 2.3 V (1111 1111).

FUNCTION GENERATOR POWER BUFFER



ELEKTOR ELECTRONICS *Fig. 38-9*

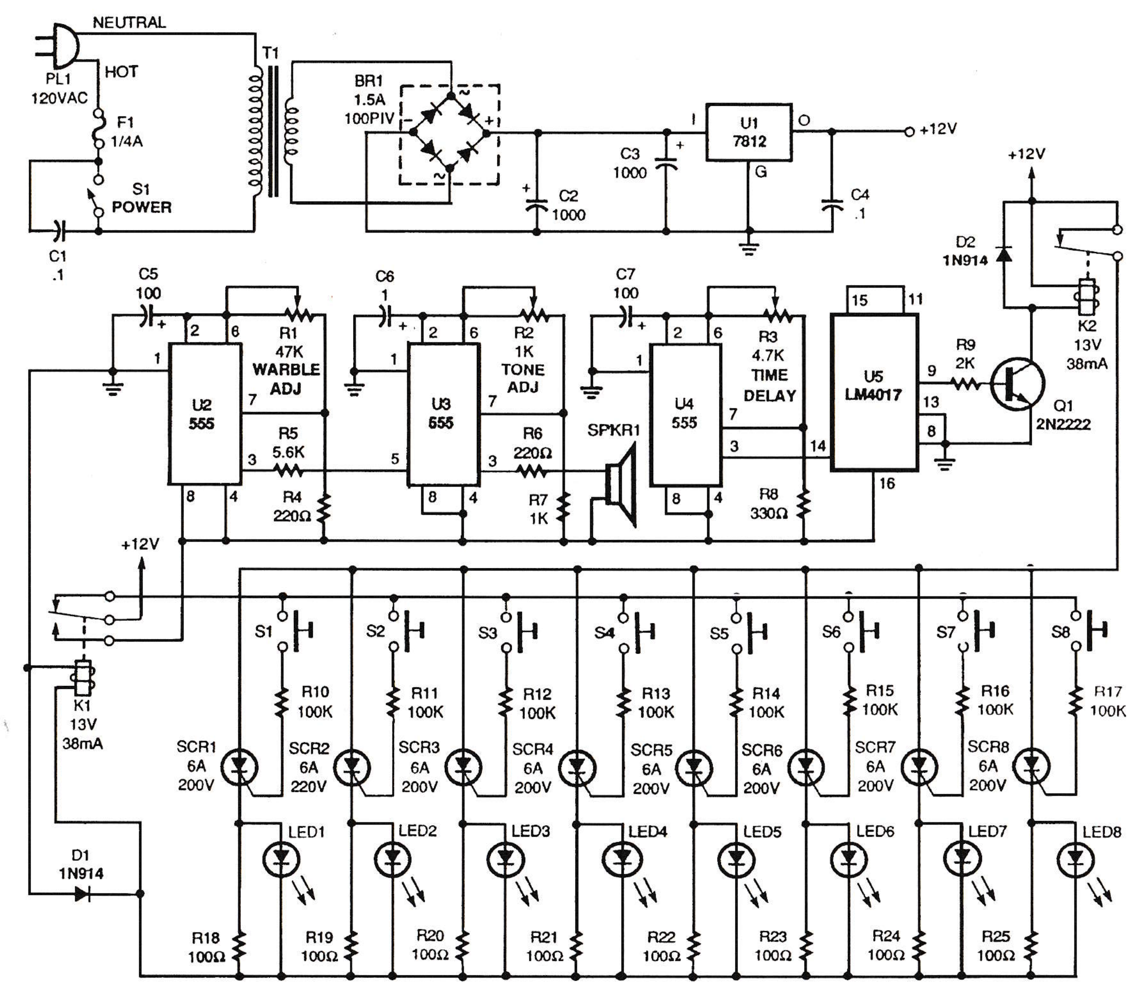
This buffer circuit can be used as an output booster for any function generator that has to be extended in order to drive several loads. The heart of the circuit is a video distribution amplifier IC from Elantec, the EL2099CT (listed by RS Components). This interesting device has a 3-dB power bandwidth of no less than 65 MHz at a gain of ×2. Here, it is used to drive up to four 50-Ω loads at a maximum signal level in excess of 10 V_{peak}. When used for video applications, the EL2099CT can drive up to six 75-Ω loads. The gain of the amplifier is ×2; unity gain is not possible because of instability problems. The bandwidth of the circuit shown here is >10 MHz, while the output achieves a drive margin of >10 V_{peak}. Current consumption will be of the order of 200 mA.

39

Game Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

LOCKOUT CIRCUIT



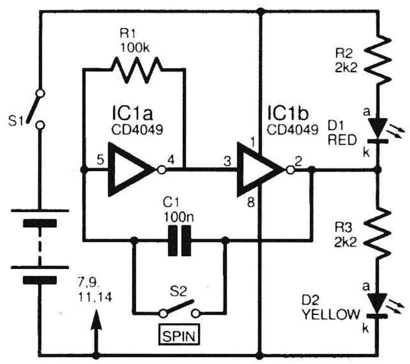
POPULAR ELECTRONICS

Fig. 39-1

LOCKOUT CIRCUIT *(Cont.)*

This circuit is used in contests, games, etc. When a contestant presses one of the buttons, an SCR triggers, lighting the associated LED and preventing anyone else from actuating the system. U2 and U3 form a warble tone generator, while U4 and U5 provide time delay.

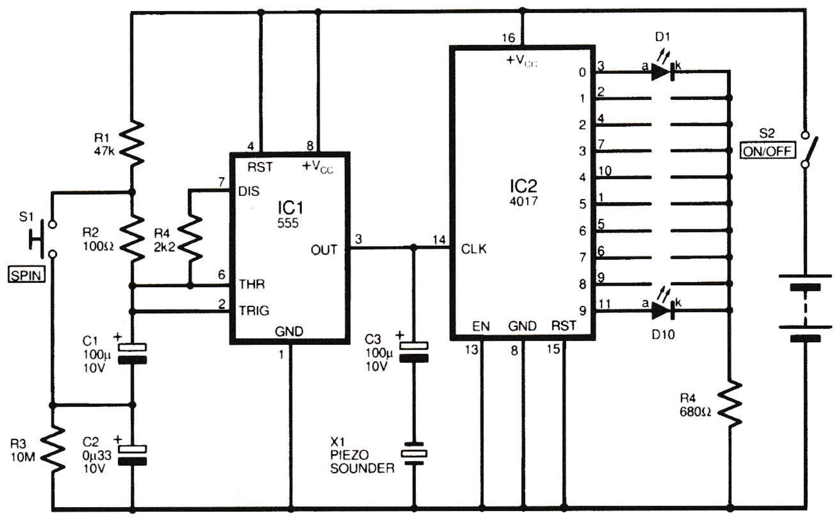
HEADS OR TAILS GAME CIRCUIT



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 39-2***

Designed to simulate by electronic means the tossing of a coin, the circuit is based upon a 4049 hex inverter IC, two of which are used. IC1a and IC1b are wired as an astable oscillator, which causes two LEDs (D1 and D2) to alternate rapidly, at a frequency too high to be distinguished by the naked eye. Both LEDs, therefore, appear to be constantly illuminated. When the SPIN switch S2 is closed, this has the effect of freezing the display, and the LED, which was illuminated at the instant that the switch was closed, will now be continuously alight. Opening the switch enables the oscillator once more. There is an equal chance of either LED lighting, and the circuit can be used in board games, for example, to choose which player will move first.

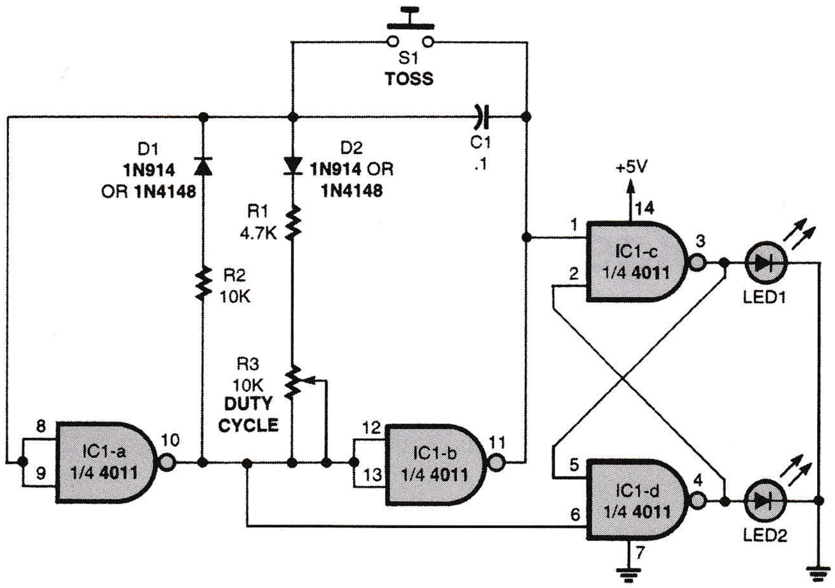
MINI ROULETTE



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 39-3***

A circuit diagram for a mini battery-powered version of roulette is shown. This circuit uses a 4017 decade counter (IC2) driving 10 LEDs. Because only one LED is ever illuminated at any one time, a common limiting resistor R4 is used. They can also be placed in alternating red/green order for added effect. The counter IC2 is clocked by IC1, a classic 555 timer connected as an astable. When the SPIN button is pressed and then released, full speed is achieved, and then the display gradually slows down until it stops on a single number. Capacitor C2 governs the oscillator speed, and resistor R3 prevents instability when the LED rotation stops. The piezo disk transducer, X1, is placed on the output of the oscillator to provide a sound effect.

COIN-TOSS CIRCUIT



POPULAR ELECTRONICS

Fig. 39-4

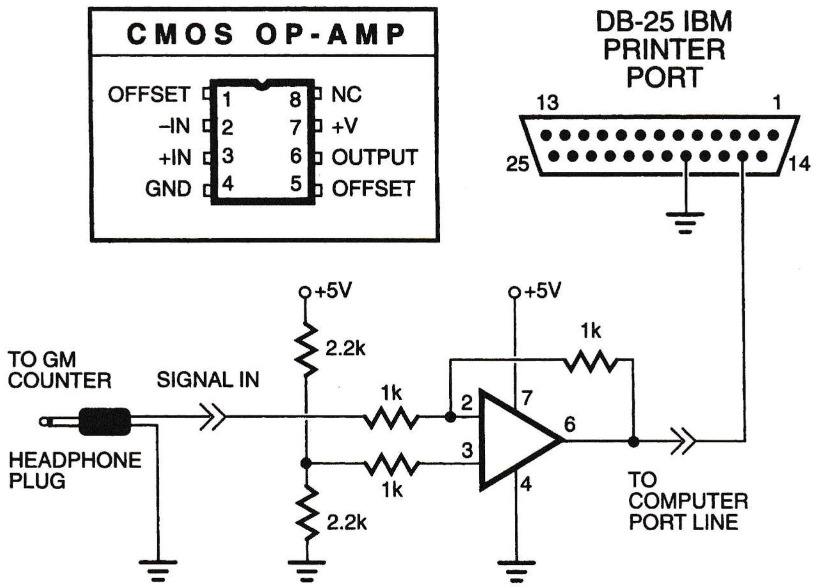
Shown is a CMOS coin-toss circuit that works well and can be built with a single 4011 or 4001 CMOS IC. (Note that the IC pin numbers in the diagram apply for both the 4001 and the 4011.) Two gates form a clock, and the others make up a bistable multivibrator. With this circuit, it is necessary to adjust the 10,000-Ω potentiometer (R3) for a 50-percent duty cycle. If you don't have a scope to do this, simply measure the direct current flowing through each LED while adjusting R3. When the same current level flows through each LED, the clock will be adjusted for a 50-percent duty cycle.

40

Geiger Counter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

GEIGER COUNTER TO IBM INTERFACE

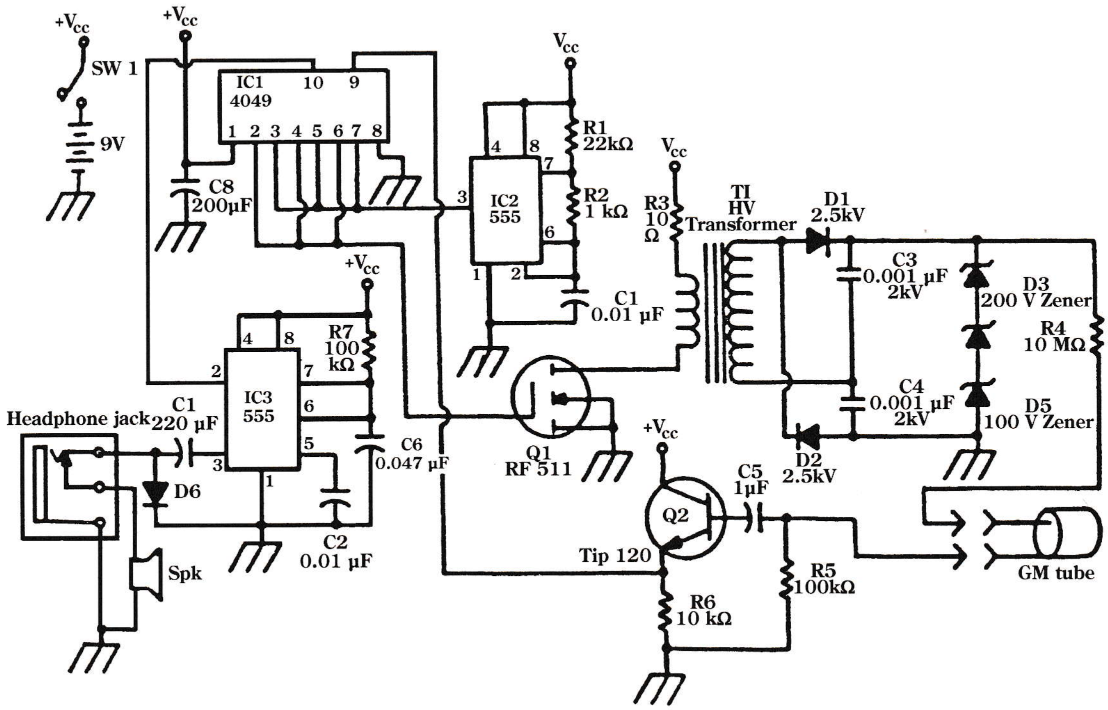


NUTS AND VOLTS

Fig. 40-1

The simplest way to connect a Geiger counter to the IBM computer is to use a headphone jack to the parallel port. The headphone jack provides a +5- to 0-V pulse for each radioactive particle detected. The signal from the Geiger counter is +5 V. When radiation is detected, the signal pulses down to 0 V for a few milliseconds before returning to +5 V. Even though the signal is in the 0- to +5-V range, it is not safe to connect wires from an external circuit or instrument directly to the parallel-port lines without buffering. The op amp operates from a single-pole +5-V power supply. In this particular case, invert the signal from the Geiger counter. Because the pulse signals are +5 to 0 V, we do not require any amplification from our op amp. Therefore, the op amp is configured as an inverting unity-gain follower. Remember that the Geiger counter outputs a steady +5 V through the headphone jack. When it detects radiation, the output pulses down to 0 V for a few milliseconds. By inverting the Geiger counter signal with the op amp, the computer reads a steady 0 V on its line and a +5-V pulse when radiation is detected.

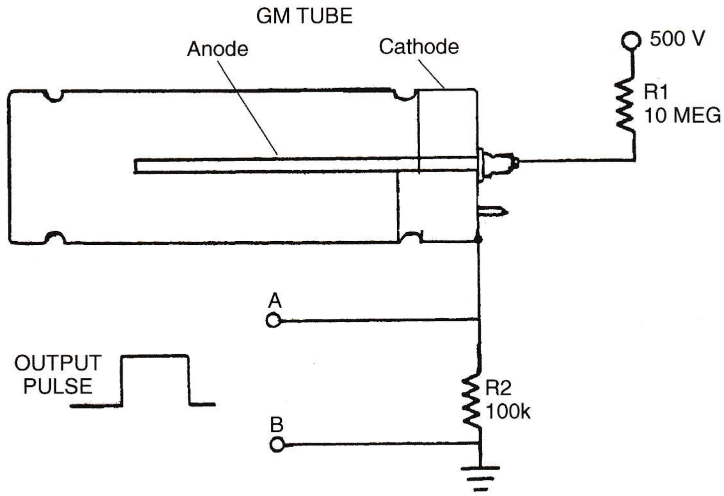
GEIGER COUNTER CIRCUIT



NUTS AND VOLTS ***Fig. 40-2***

IC2 is a 555 timer set in astable mode. The signal from IC2 is presented to three gates on the 4049 IC1. The 4049 inverts the signal to give an optimum pulse width that switches Q1 on and off. The MOSFET (Q1), in turn, switches the current to a step-up transformer (T1). The stepped-up voltage from T1 first passes through a voltage doubler; the output voltage from this section is approximately 600 to 700 V. Three zener diodes (D3, D4, and D5) are placed across the output of the voltage doubler to regulate the voltage to 500 V. This voltage is connected to the anode on the GM tube through a 10-MΩ resistor. When the GM tube detects a particle, a voltage pulse from the 100-kΩ resistor is amplified and clamped to V_{CC} via Q2, an NPN Darlington transistor. The signal from Q2 is inverted by IC1, where it acts as a trigger signal to IC3. IC3 is another 555 timer. The output of IC3 via pin 3 flashes on LED and provides a click into either the speaker or the headphones. The circuit is powered by a 9-V alkaline battery and draws about 28 mA when not detecting.

GEIGER-MUELLER TUBE CIRCUIT



NUTS AND VOLTS ***Fig. 40-3***

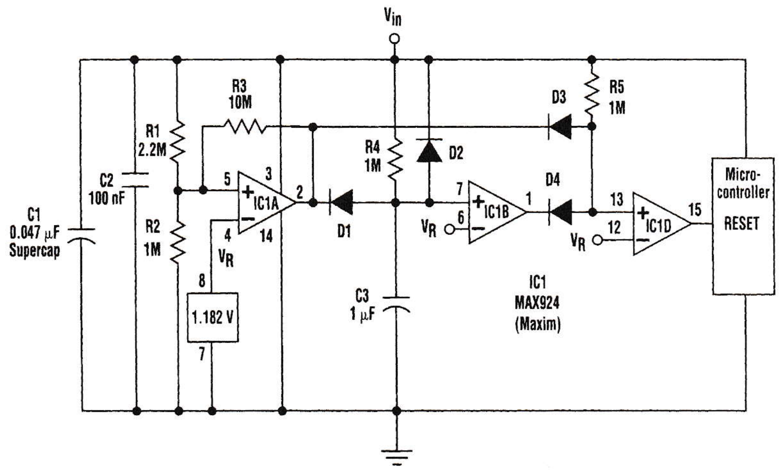
A GM tube is useful for detecting radioactivity. The tube is constructed with a cylindrical electrode (cathode) surrounding a center electrode (anode). The tube is evacuated and filled with a neon and halogen gas mixture. A voltage potential of 500 V is applied across the tube, through a 10-MΩ current-limiting resistor (R1). The detection of radiation relies upon its ability to ionize the gas in the GM tube. The tube has an extremely high resistance when it is not in the process of detecting radioactivity. When an atom of the gas is ionized by the passage of radiation, the free electron and positive ionized atom that are created move rapidly toward the two electrodes in the GM tube. In doing so, they collide with and ionize other gas atoms, which creates a small avalanche effect. This ionization drops the resistance of the tube, allowing a sudden surge of electric current that creates a voltage across the resistor R2, which can be seen as a pulse.

41

Generator Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

RESET GENERATOR

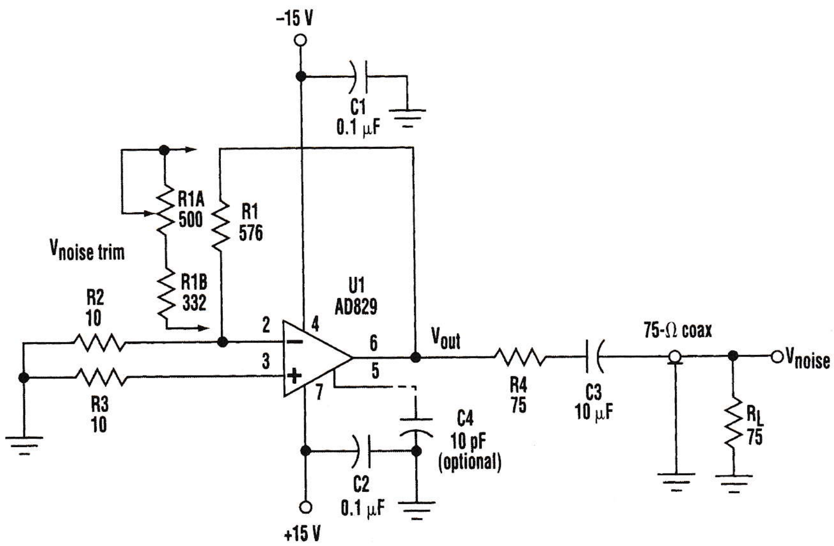


ELECTRONIC DESIGN *Fig. 41-1*

This V_{CC} monitor circuit draws less than 10 μA, yet it can generate reliable RESET signals.

Very low quiescent current makes the reset generator shown suitable for use with microcontrollers that spend most of their time in the "sleep" mode. The system's power source, a charged 0.047-μF "Supercap" capacitor (C1), can go without a recharge for intervals exceeding 24 hours. The allowable supply-voltage range for the reset circuit is 2 to 11 V. R1 and R2 make it possible to set the switching threshold as low as 1.2 V (the values shown provide a 4.10-V threshold). Hysteresis introduced by $R3$ minimizes the effect of reference noise and external EMI. Using the values given for R_2 and R_3 (1 MΩ and 10 MΩ), the switching threshold (V_t) is obtained by adjusting the value of R_1 alone: R_1 (in megohms)$=0.770V_t-0.910$. External connections tie the internal reference voltage to the inverting input of each comparator. Comparator A monitors the supply voltage. When this voltage dips below the threshold established by R1 and R2, the comparator output initiates a RESET pulse via D3 and discharges the timing capacitor (C3) via diode D1. Comparator B determines the pulse duration with help from the time-delay components R4 and C3. Because the duration (approximately 250 ms) is well above the 50-ms minimum required in this case, the effect of supply voltage on pulse width can be ignored. D2 allows the circuit to respond to short outages by providing another quick-discharge path for the timing capacitor. Comparator C is a spare that supplies a complementary output or low-line warning. Comparator D acts as an active AND gate for logic-high signals from comparators A and B. The pull-up resistor in many such circuits is 100 kΩ or less (to avoid errors caused by the comparator's input leakage). In this particular case, it would provide a path for excessive battery discharge during reset.

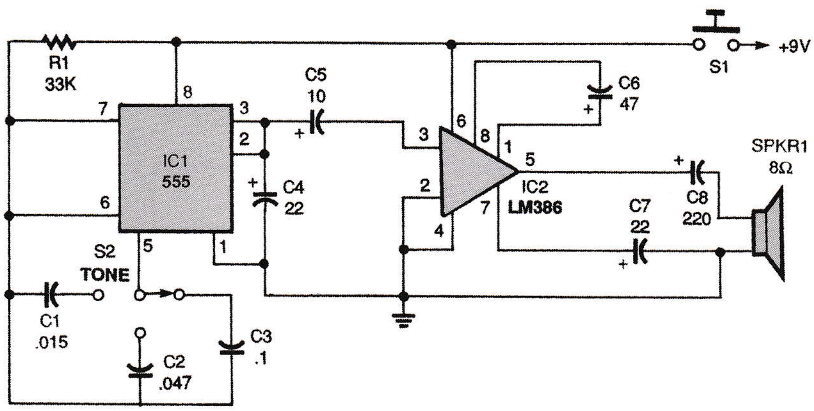
SIMPLE WIDEBAND NOISE GENERATOR



ELECTRONIC DESIGN ***Fig. 41-2***

This relatively simple op-amp noise generator amplifies the input voltage noise of a wideband, decompensated op amp. If a device with a single-gain stage is selected, the output noise will be spectrally flat up to the closed-loop bandwidth. The op amp is manufactured by Analog Devices Inc.

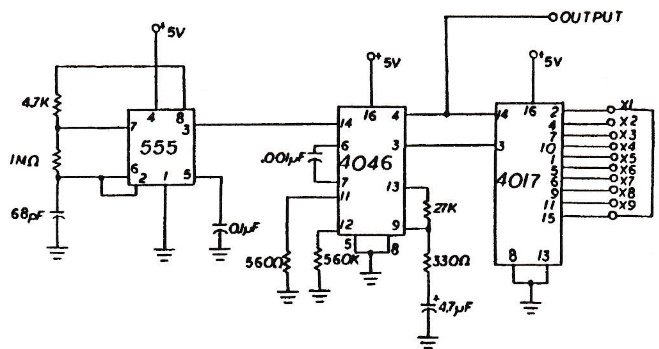
THREE-TONE GENERATOR



POPULAR ELECTRONICS ***Fig. 41-3***

The three-tone generator makes a great warning device. It has a lot of uses; for example, if the appropriate switch is used for S1, the circuit can be used as a burglar alarm. When S1 is pressed, it turns on the circuit. The tone frequencies depend on resistor R1 and capacitors C1 through C3. Switch S2 lets you select between those capacitors. Capacitor C1 produces the highest frequency because it has the lowest capacitance, while C3 generates the lowest frequency

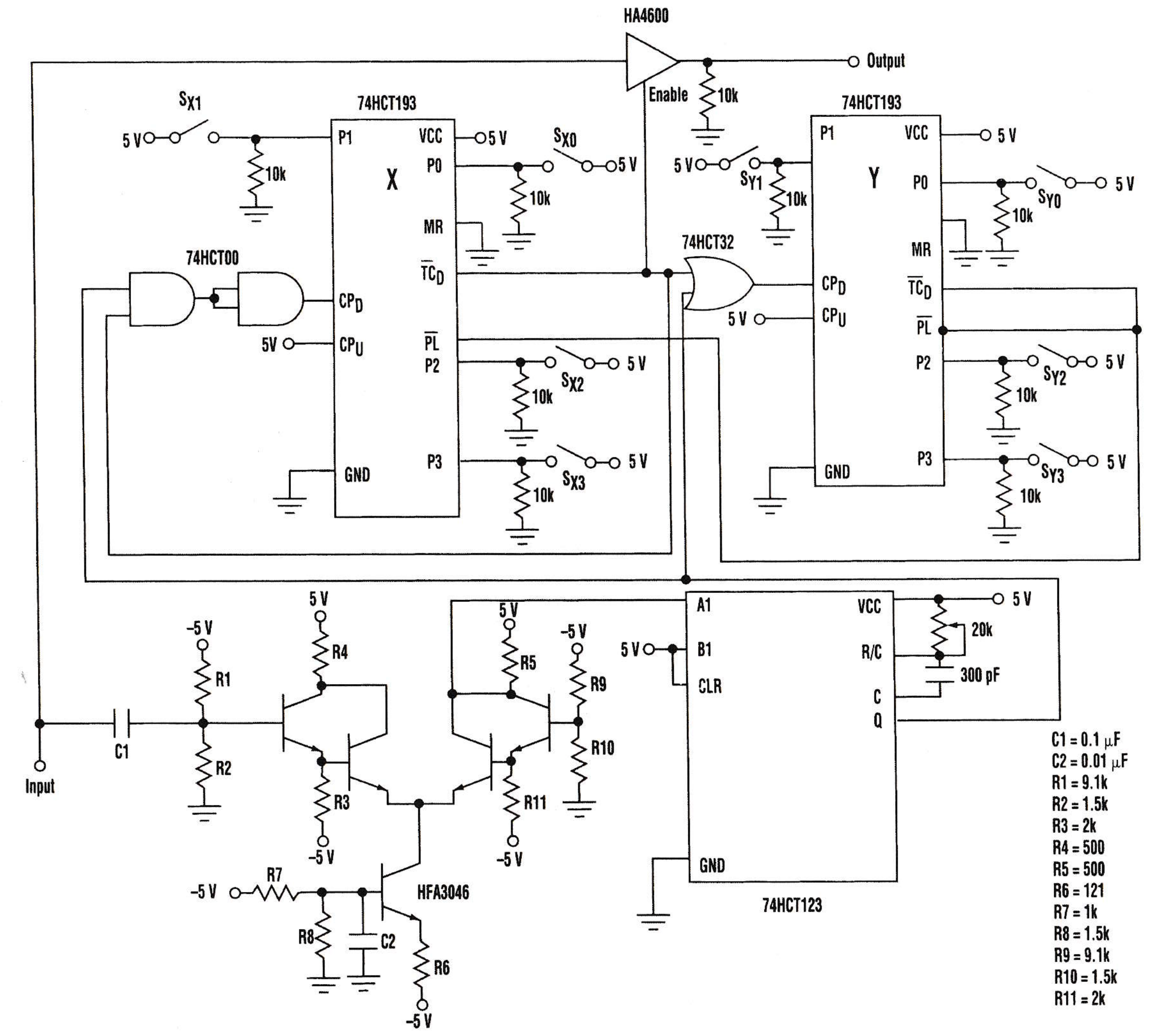
SIMPLE FREQUENCY SYNTHESIZER



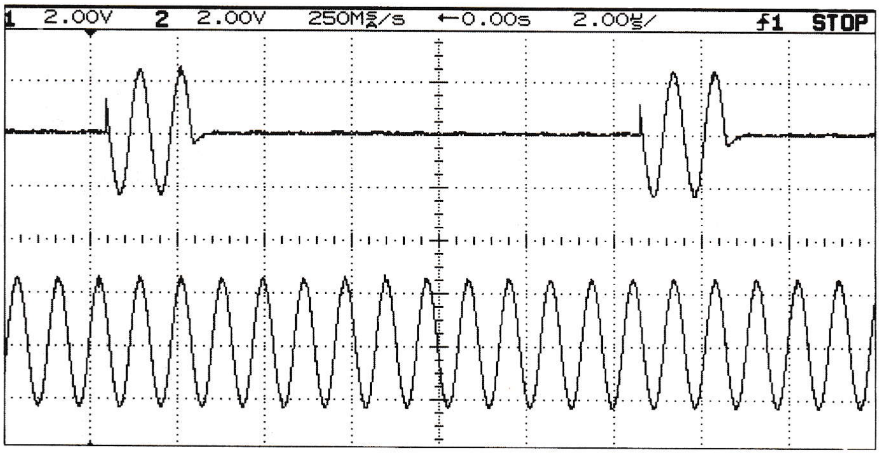
NUTS AND VOLTS ***Fig. 41-4***

The 555 timer circuit is configured as a 10-kHz astable multivibrator that feeds this signal to the 4046 phase-locked-loop IC. The signal that is fed from this chip is then coupled to a divide-by-*N* counter. This counter will take the 10-kHz signal locked on by the 4046 and will produce multiples of the fundamental frequency (10 kHz). Therefore, it is possible to generate frequencies as high as 100 kHz with this circuit. When building this circuit, remember to use safety precautions when handling CMOS chips.

DIGITAL BURST GENERATOR



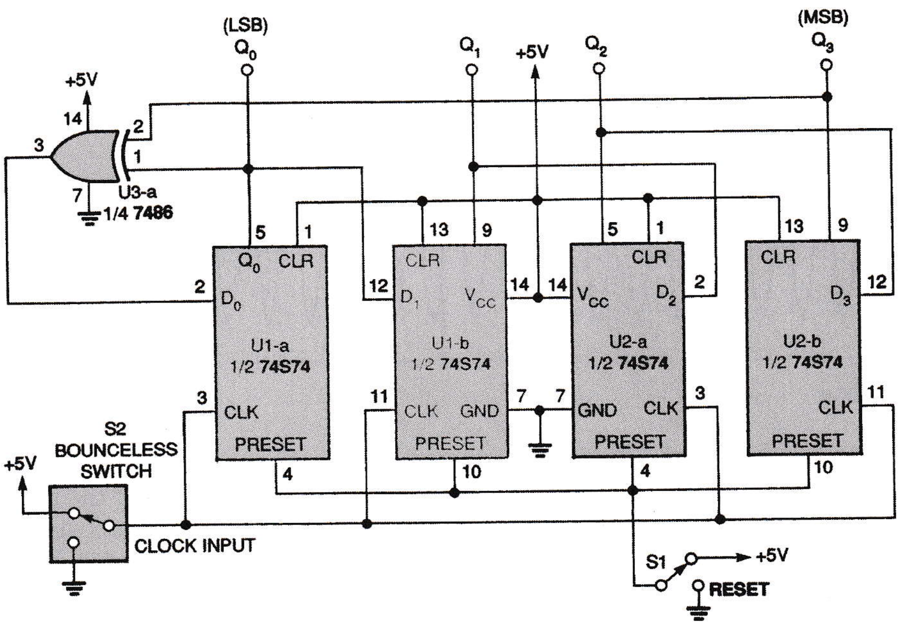
C1 = 0.1 μF
C2 = 0.01 μF
R1 = 9.1k
R2 = 1.5k
R3 = 2k
R4 = 500
R5 = 500
R6 = 121
R7 = 1k
R8 = 1.5k
R9 = 9.1k
R10 = 1.5k
R11 = 2k



ELECTRONIC DESIGN **Fig. 41-5**

Figure 1 illustrates the transmitted signal and the continuous sine wave from which it is derived. Because the HA4600 buffer has an enable/disable feature, it will pass or reject the input waveform, depending on the state of the enable pin (Fig. 2.). The input signal also is present at the input to the HFA3046 transistor array, which has been configured as a high-speed, high-gain comparator. The comparator squares up the input signal and applies it to the inputs of the two counters, X and Y. The X counter controls the buffer enable, and it determines how many cycles of the input waveform get passed to the output. The four switches S_{X0} through S_{X3} are binary-coded. Consequently, if two switches (S_{X0}=1 cycle and S_{X1}=2 cycles) are closed, three cycles of the input sine wave will be passed to the output. Furthermore, the input signal is connected to the Y counter, which controls the repetition rate by determining the OFF period between pulse bursts. The four switches S_{Y0} through S_{Y3} are binary-coded. When all of these switches are closed, the OFF period will be 16 times the period of the input waveform. With the X and Y counters set as described earlier, the repetition rate is the reciprocal of (16+3) times the period of the incoming waveform. If a longer repetition rate is desired, a flip-flop or another counter can be added in series with the output of the Y counter to extend the OFF time. R6, R7, and R8 bias the long-tailed transistor at 10 mA, which is the optimum point for speed. R5 and R6 are small enough to discharge quickly, thus preventing saturation. Configured as shown, the circuit will handle 10-MHz input signals with little degradation. The limit on frequency response is the speed of the logic and the comparator delay time. The comparator delay time can be eliminated by one-shotting out the delay.

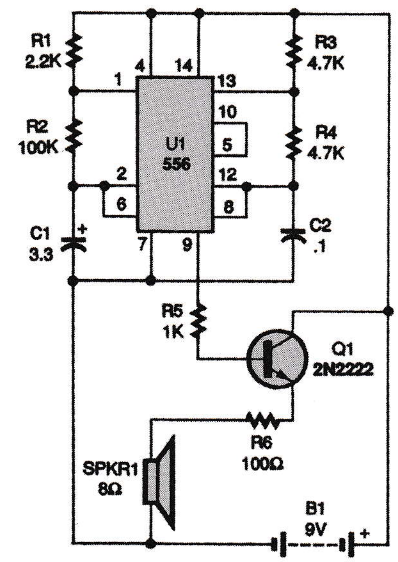
RANDOM-NUMBER GENERATOR



POPULAR ELECTRONICS *Fig. 41-6*

The 74S74 flip-flops shown in the circuit are arranged to form a 4-bit shift register. Binary data enter the D0 input on U1-a and are sequentially shifted to each output (Q0, Q1, Q2, Q3) with each clock pulse. The data input to the shift register come from the output of U3-a, one gate of a 7486. That exclusive-OR gate compares two of the output bits from the shift register. If the two bits are the same, then the output of U3-a is 0 V (or low). If the two bits are different, then the output of U3-a is +5 V (or high). Therefore, U3-a acts as a type of logical-feedback network that changes the data at D0, which, in turn, changes the outputs of the flip-flops. The effect of that network is to create a pseudo-random sequence of bits at the outputs of U1 and U2.

TONE-BURST CIRCUIT



POPULAR ELECTRONICS *Fig. 41-7*

The tone-burst circuit puts out a 500-Hz tone, at a rate of 1 Hz, through an 8-Ω speaker, SPKR1. Besides the 9-V battery and its connecting snap, SPKR1 is the only non-surface-mount component in the circuit. The circuit can also be built from standard-size components.

SIMPLE FREQUENCY GENERATOR

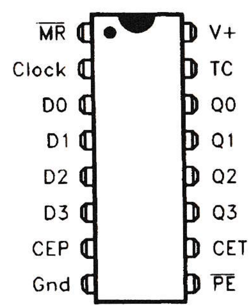


Fig. 1. Pinout details for the 74HC161 binary counter.

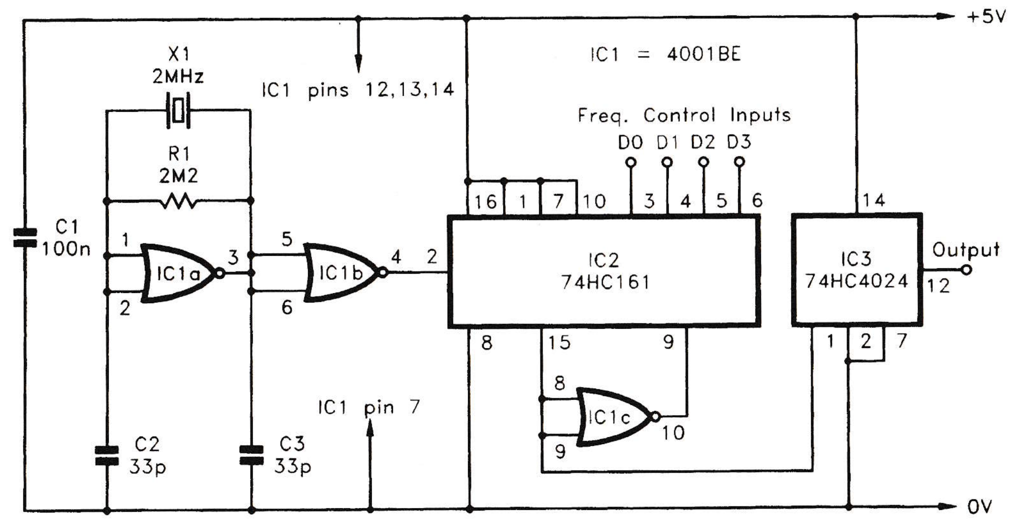


Fig . 2. Circuit diagram for a Simple Frequency Generator using the 74HC161.

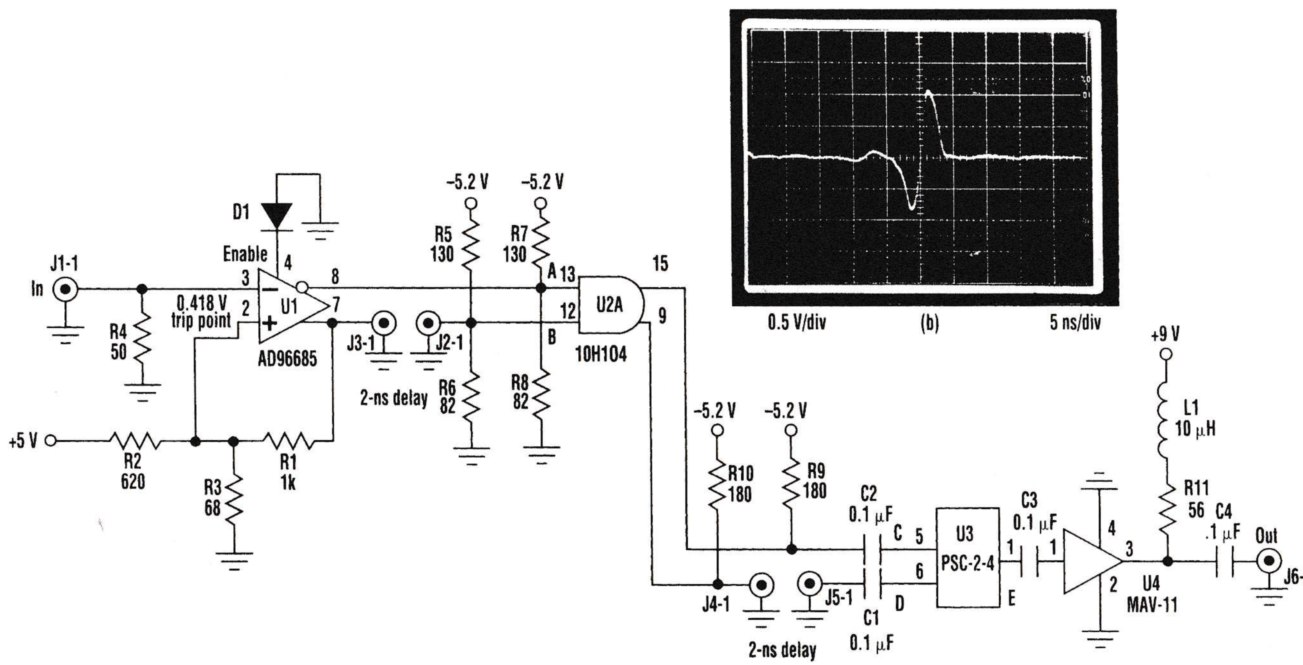
Table 1. Division Rates and Output Frequencies

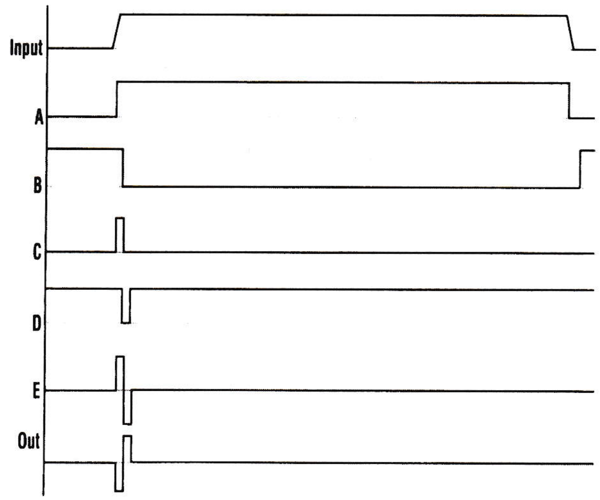
D0	*D1*	*D2*	*D3*	*Div.*	*Frequency (kHz)*
L	L	L	L	16	125
H	L	L	L	15	133·333
L	H	L	L	14	142·857
H	H	L	L	13	153·846
L	L	H	L	12	166·55
H	L	H	L	11	181·181
L	H	H	L	10	200
H	H	H	L	9	222·222
L	L	L	H	8	250
H	L	L	H	7	285·714
L	H	L	H	6	333·333
H	H	L	H	5	400
L	L	H	H	4	500
H	L	H	H	3	666·666
L	H	H	H	2	1000 (1MHz)
H	H	H	H	-	-

EVERYDAY PRACTICAL ELECTRONICS ***Fig. 41-8***

The circuit diagram for a simple frequency generator that uses a divide-by-*N* counter, based on a single 74HC161, is shown. Although IC1a and IC1b are NOR gates, in this circuit, they are used as inverters in a crystal clock generator. This provides an accurate 2-MHz output signal that is fed to the input of the divide-by-*N* circuit based on IC2 and IC1c. An inversion is needed between the comparator output and the negative active preset enable input, and this inversion is provided by IC1c. Unfortunately, setting the division rate is more convoluted than simply writing the required value for *N* to the data inputs (D0 to D3). Table 1 shows the division rates and output frequencies for the 16 input codes for IC2. If the data inputs of IC2 are controlled via inverters, the division rate is one more than the value written to the port. Without a hardware inversion, a software inversion is required. This is actually quite easy, and it is just a matter of deducting the required division rate from 16 (e.g., for a division rate of 4, a value of 12 is written to IC2). Notice that the minimum division rate is 2, and that writing a value of 15 to IC2 will not give a division by 1.

ULTRA-FAST MONOCYCLE GENERATOR



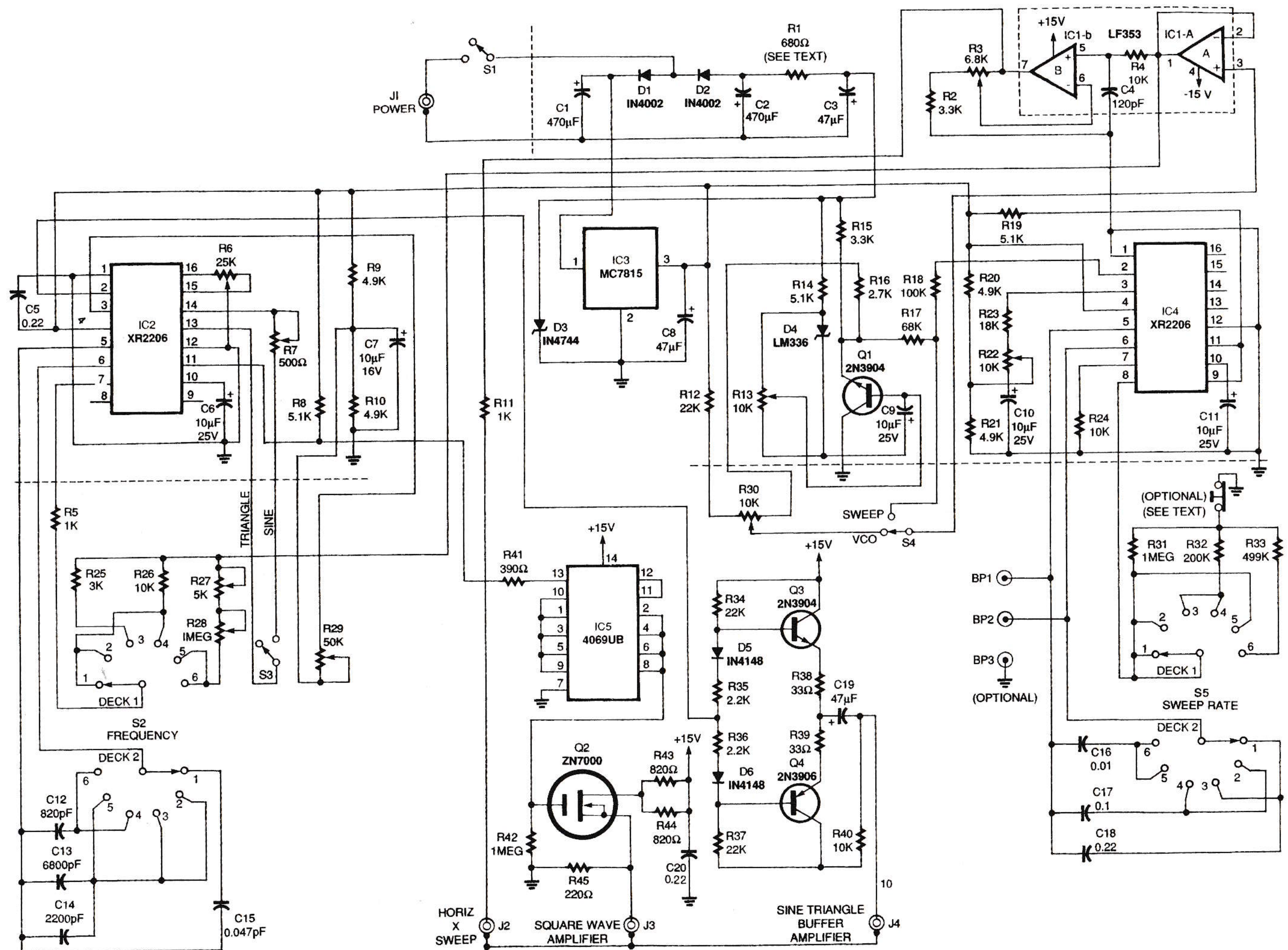


THE NONINVERTING OUTPUT feeds a 2-ns coaxial delay line, which causes the signal to be delayed with respect to the inverting output (A and B). The AND gate's inverting output drives a second 2-ns delay line. This delays the negative-going pulse with respect to the positive-going pulse (C and D).

ELECTRONIC DESIGN ***Fig. 41-9***

This circuit (Fig. 1*a*) can test an ultra-fast sample-and-hold amplifier, producing the monocycle shown (Fig. 1*b*) when triggered with a TTL input pulse. The comparator (U1) squares up the input pulse and drives the AND gate (U2) with complementary outputs. The hysteresis resistors are chosen to provide reliable triggering despite the 50-Ω loading of the TTL input. The noninverting comparator output feeds a 2-ns coaxial delay line, causing the signal to be delayed with respect to the inverting output (Fig. 2). This forces the outputs of the AND gate to change state for a period of time equal to the delay. The inverting output of the AND gate drives a second 2-ns delay line, delaying the negative-going pulse, with respect to the positive-going pulse (Fig. 2). The power combiner (U3) sums the signals, producing a monocycle output. U4 amplifies the signal. Shorter pulses could be produced by shortening the delay lines and replacing U2 with an AND gate from a faster ECL family.

SWEEP GENERATOR



SWEEP GENERATOR *(Cont.)*

TABLE 1
FUNCTION GENERATOR CHARACTERISTICS

Waveform output	Maximum P-to-P	Frequency	Conditions
Sine (1)	5V	10 Hz-100 kHz	1 V@800 kHz
Triangle (1)	8 V	10 Hz- 50 kHz	1 V>500 kHz
Square (2)	5 V		Positive output DC-coupled, ground ref: rise/fall >50 ns
Ramp (3)			Descending, 6 rates

(1) Output level variable frim min. to max.
(2) Output level not adjustable.
(3) X and Y amplitude internally adjustable.

TABLE 2
SWEEP RANGES OF THE FUNCTION GENERATOR

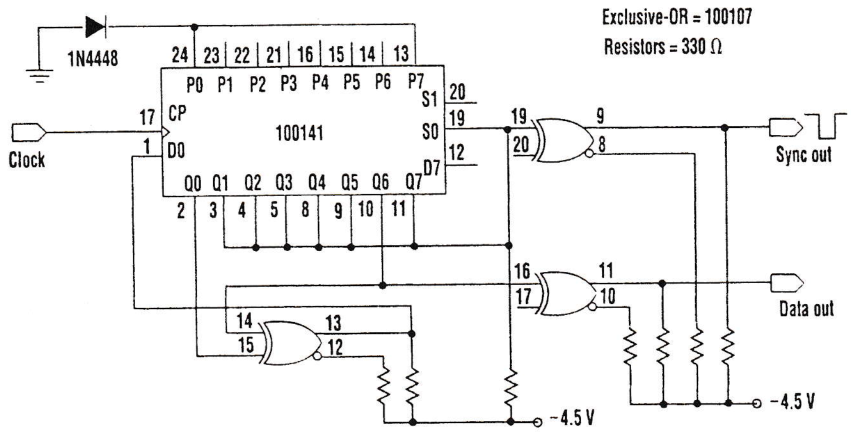
Switch	Condition	Frequency range
1	Preset	20Hz to >2kHz
2	Preset	<400Hz to >10kHz
3	Preset	<1kHz to >25kHz
4	Preset	5kHz to >100kHz
5*	Resistance tuned	2kHz to 100kHz
	Resistance & VCO tuned	<10Hz to >100kHz
6*	Resistance tuned	<40kHz to >800kHz
	Resistance & VCO tuned	<100Hz to >800kHz

* Ranges show for positions 5 and 6 represent the total tuning range of the function generator and do not imply one continuous sweep.

ELECTRONICS NOW ***Fig. 41-10***

Both IC2 and IC4 are Exar XR2206 monolithic function generators; IC4 functions as a ramp generator, and IC2 functions as a generator of sine, triangular, and square waveforms. Dual operational-amplifier IC1 produces a scaled, level-shifted ramp output that is capable of deflecting an oscilloscope's horizontal sweep. This ensures that the sweep generator and the oscilloscope's sweep circuit are always properly synchronized. Any frequency of interest along the horizontal axis of an oscilloscope that is coupled to this function generator can be measured with an external frequency counter by manually tuning the function generator's VCO instead of sweeping it. The performance characteristics of the sweep/function generator are summarized in Table 1. The generator's sweep rate and frequency can be set by front-panel rotary six-position switches, SWEEP RATE switch S5 and FREQUENCY switch S2. The VCO control R30 manually tunes the VCO. Table 2 lists the sweep ranges of the function generator. Sweep ranges other than those covered in ranges 1 to 4 can be set up as required on positions 5 and 6. Selecting the VCO setting on the front-panel toggle switch S4 permits tuning any fixed frequency within the total frequency range of the instrument with both the FREQUENCY switch S2 and VCO control R30.

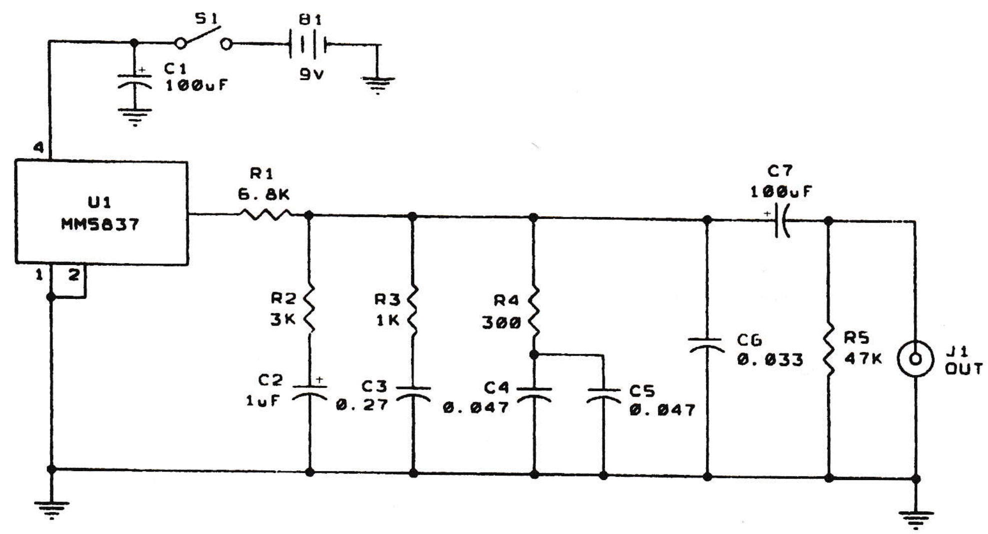
SELF-STARTING DATA GENERATOR



ELECTRONIC DESIGN *Fig. 41-11*

Pseudo-random sequence generators built from shift registers and exclusive-OR gates often are used to supply binary test data. If constructed from 100-kΩ ECL parts, such generators can run at up to 200 Mbits/s. Although an N-bit shift register can be connected to generate a sequence that repeats every 2^{N-1} bits, if it should start up in its all-zeros state, no output will be generated. What is needed is a counter to inject a 1 or to preload the shift register whenever N consecutive zeros are detected. The illustrated circuit generates a 127-bit sequence. It provides a synchronization pulse once per repetition without using additional parts, and it is guaranteed to start. It uses the "wired-OR" property of ECL to generate a 1-bit period negative sync pulse when six zeros are present (only five of these zeros are consecutive; a different set of $N-1$ bits might be needed if a longer shift register is used). The sync pulse parallel-loads the shift register with the next state in the sequence. As a result, no seven-zero detector is needed to start the generator.

PINK-NOISE GENERATOR



NUTS AND VOLTS ***Fig. 41-12***

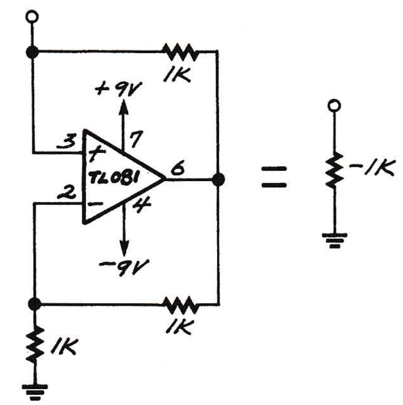
The MM5837 is a digital white-noise generator IC. It produces a clean white noise signal with only a power source. White noise appears at the output of this IC when power is applied. The white-noise signal is then fed through a −3-dB/octave filter to give pink noise. Because the minimum rolloff with a single-stage *RC* (resistor/capacitor) filter is −6 dB/octave (because of capacitive reactance), an unconventional filter design is needed. The technique involves cascading several stages of lag compensation so that the zeros of one stage partially cancel the poles of the next stage. The result is shown in the schematic of the figure. The response of this circuit is accurate to within ±1/2 dB.

42

Impedance Converter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

NEGATIVE IMPEDANCE CONVERTER

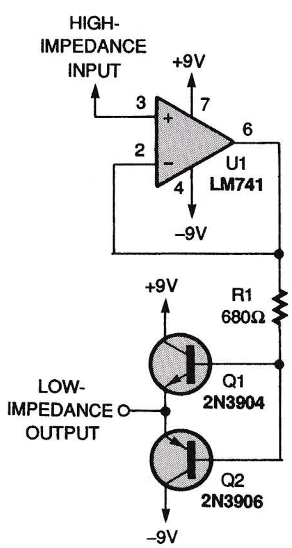


Recall that a resistor is a voltage-dependent current consumer; this circuit does the opposite. If you connect a battery to its input, the battery will get charged, with exactly the current that the battery would have delivered if you had connected it to a 1000-Ω resistor. If you put a 1000-Ω resistor in parallel with a 1000-Ω NIC, you'll get what looks like an infinite resistance. No matter what voltage you apply (within limits), the NIC will match it, and all the current flowing in the resistor will come from the NIC, not from the externally applied voltage.

ELECTRONICS NOW

Fig. 42-1

IMPEDANCE CONVERTER



This circuit is a high-input-impedance–to–low-output-impedance converter circuit with unity voltage gain. In the circuit, an LM741 op amp, U1, is connected in a voltage-follower circuit that drives a complementary transistor-emitter-follower circuit. The output of the circuit can be used to drive low-current lamps, relays, speakers, etc. The feedback can be taken from the output to pin 2 of the LM741 if it is desired to include Q1 and Q2 in the feedback loop.

POPULAR ELECTRONICS

Fig. 42-2

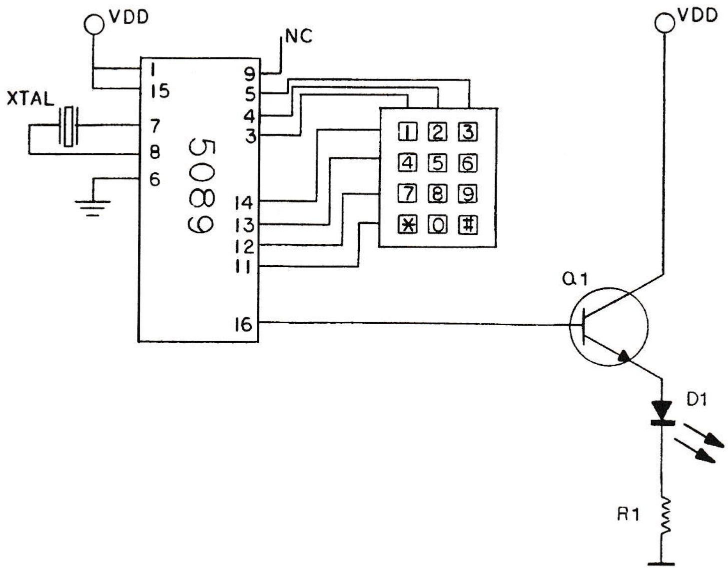
43

Infrared Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

- DTMF Infrared Transmitter
- Warble Tone Infrared Transmitter
- IR Remote Tester
- Low-Cost IR Filter
- Infrared Alarm Transmitter
- Mono Cordless Headphone IR Transmitter
- Infrared Alarm Receiver
- Mono Cordless Headphone IR Receiver
- IR Remote-Control Repeater
- Infrared Data Receiver
- IR Local Talk Link Receiver
- IR Remote-Control Receiver
- IR Remote-Control Dc Interface
- IR Remote-Control Test Set
- IR Remote-Control Relay Interface
- Phototransistor IR Receiver
- Infrared Data Transmitter
- IR Remote-Control Triac Interface
- See-Through Sensor
- IR Remote-Control Model Railroad Application
- IR Remote-Control Receiver
- IR Relay Circuit
- Cordless Headphones

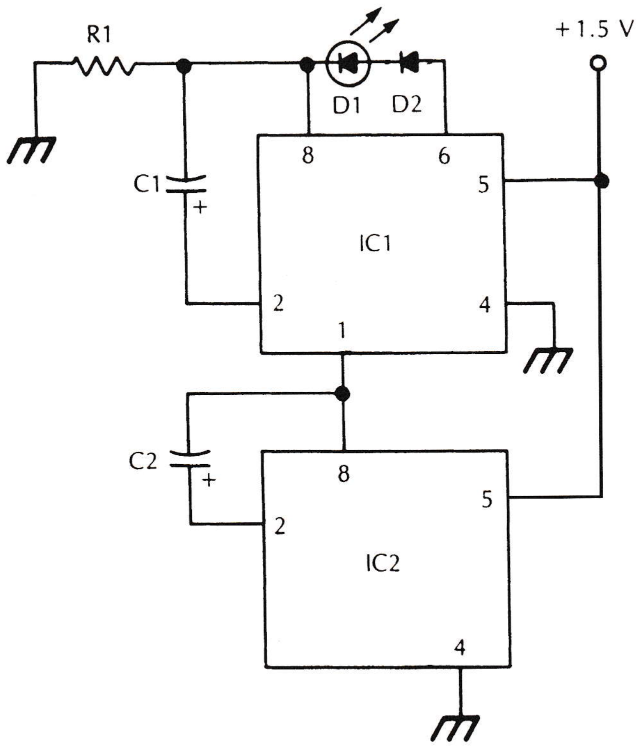
DTMF INFRARED TRANSMITTER



NUTS AND VOLTS *Fig. 43-1*

This transmitter uses a 5089 DTMF generator chip and a keypad to generate DTMF signals and modulate them on an IR light beam from an IR LED. Xtal is a 3.579-MHz TV burst crystal.

WARBLE TONE INFRARED TRANSMITTER

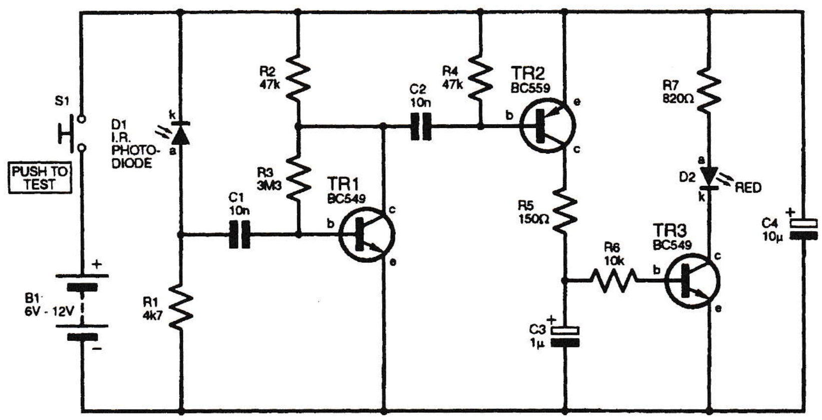


IC1, IC2	LM3909 LED flasher/oscillator IC
D1	infrared LED
D2	diode (1N4148, 1N914, or similar)
C1	1 μF 5 V electrolytic capacitor
C2	47 μF 5 V electrolytic capacitor
R1	470 Ω ¼ W 5% resistor

TAB BOOKS *Fig. 43-2*

IC1 and IC2 are LM3909 LED flasher-oscillator devices. IC2 generates a low-frequency square wave that modulates the frequency of IC1. This circuit is useful for IR testing, communications link signal source, etc.

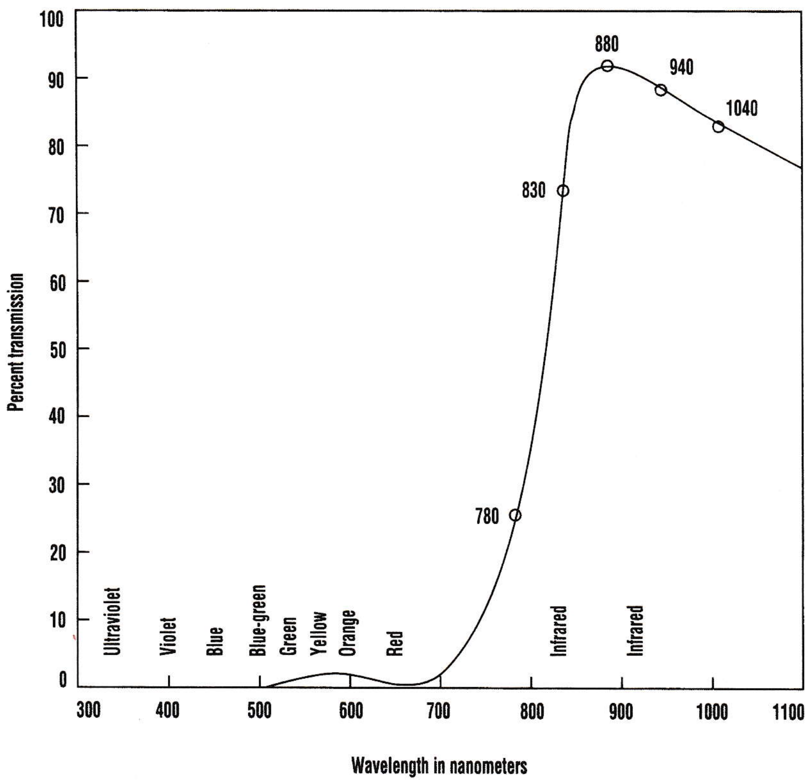
IR REMOTE TESTER



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 43-3***

This remote handset tester is very useful and convenient, and could quickly pay for itself by helping to recover faulty remote handsets. Pulsed infrared light generated by the handset falls upon D1, which creates a pulse waveform across R1. This is capacitively coupled to the base of TR1, and an amplified signal is coupled to C2 to a pulse-stretching circuit based around TR2, C3, and associated components. Hence, driver transistor TR3 conducts and illuminates the LED D2 whenever infrared light is received by D1. A functional remote controller with fresh batteries will operate the tester from approximately 500 mm, while one with nearly exhausted batteries may work over only a few centimeters. Coincidentally, the design also self-tests its own battery by giving an initial flash of D1 when the switch S1 is first closed. The IR photodiode should be roughly a 940-nm type, while the types of the transistors themselves are not crucial. The circuit will operate from a 6- to 12-V rail (e.g., a 12-V battery), and C4 decouples the power supply rails.

LOW-COST IR FILTER

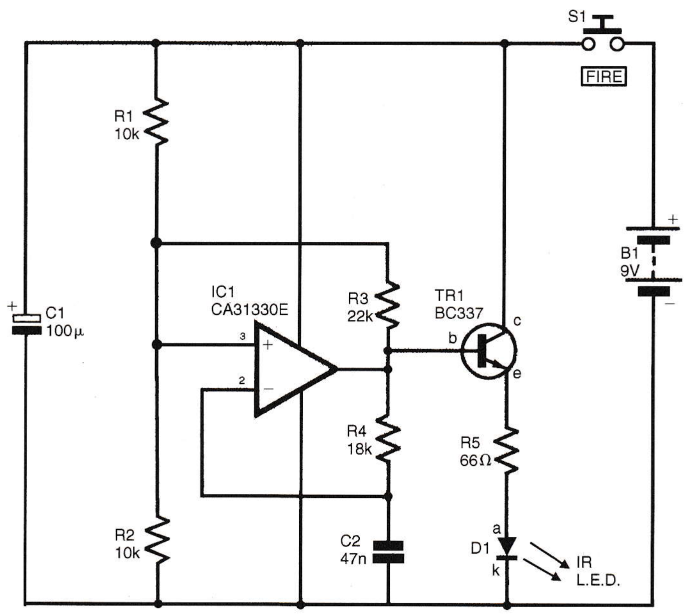


ELECTRONIC DESIGN

Fig. 43-4

When exposed to "cool white" fluorescent light for 5 s, the color negative (using Kodacolor 100 ASA film) produced after the developing process exhibits a sharp cutoff at about 830 mm. This is perfect for many IR LEDs and other IR devices.

INFRARED ALARM TRANSMITTER

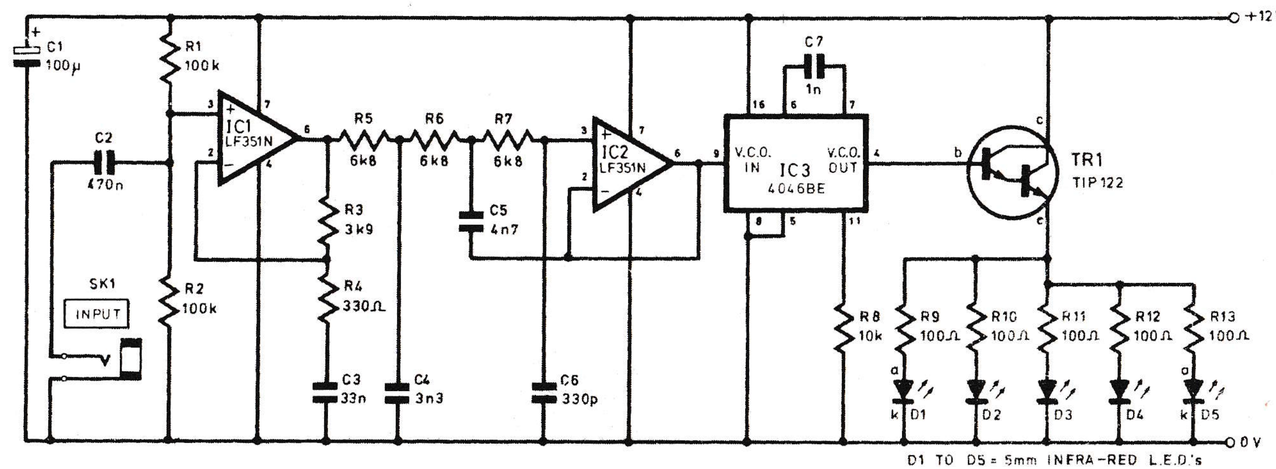


EVERYDAY PRACTICAL ELECTRONICS

Fig. 43-5

The circuit diagram for the infrared transmitter appears in this figure. The oscillator uses IC1 as a relaxation oscillator. Capacitor C2 and resistor R4 are the timing components, and they are connected between the output of IC1 and its inverting input (pin 2). A roughly square-wave signal is generated at pin 6, the output of IC1. This signal is used to drive infrared LED D1, via emitter-follower buffer-stage transistor TR1 and current-limiter resistor R5. The specified value for R_5 sets the LED current at nearly 100 mA, but as the LED is switched off for about 50 percent of the time, the average LED current is a little under 50 mA. This is the maximum acceptable drive current for most normal infrared LEDs.

MONO CORDLESS HEADPHONE IR TRANSMITTER

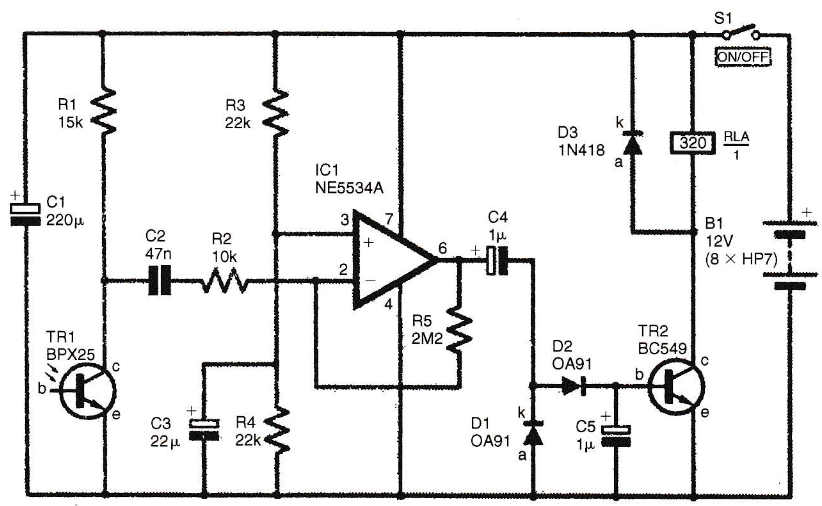


EVERYDAY PRACTICAL ELECTRONICS

Fig. 43-6

The preemphasis is applied by IC1, which is basically just an op amp used in the noninverting mode. At low and middle frequencies, the impedance of capacitor C3 is high in relation to the resistance of R3. Consequently, there is virtually 100 percent negative feedback through resistor R3, and IC1 has a closed-loop gain of little more than unity. At higher frequencies, the impedance of C3 is relatively low, and a significant proportion of the feedback through R3 is decoupled. This gives a closed-loop voltage gain that steadily rises with increased input frequency, with almost 20 dB of gain being provided at the highest audio frequencies. Resistor R4 limits the closed-loop voltage gain of IC1 at frequencies above the audio range. The low-pass filter is a conventional third-order (18 dB/octave) type based on IC2. This filter gives fractionally less than the full 20-kHz audio bandwidth, but does not significantly impair the quality of the input signal. A CMOS "micropower" PLL is used for IC3, but, in this circuit, only the VCO stage is utilized. No connections are made to any of the other sections of IC3. The output of IC2 is direct-coupled to the control input of the VCO (IC3 pin 9). Capacitor C7 and resistor R8 are the VCO timing components, and they set the center frequency at roughly 100 kHz. Transistor TR1 is a high-gain power Darlington device, which is used here as an emitter-follower buffer stage. This can easily source the 500-mA ON current of the LEDs. A bank of five LEDs is used, and each one has a separate current-limiting resistor (R9 to R13).

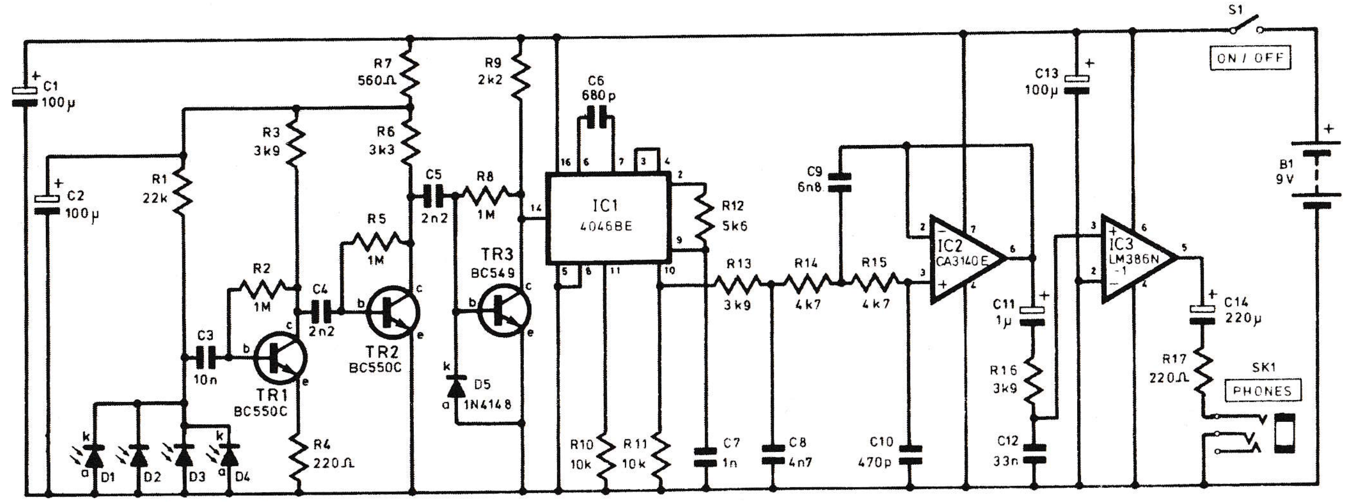
INFRARED ALARM RECEIVER



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 43-7***

The full circuit diagram for the infrared receiver is given here. TR1 is a phototransistor, and it is used in a common-emitter amplifier. However, no bias current is fed to the base (b) terminal of TR1. The collector (c) current is governed by the light level received by TR1. The higher the received light level, the higher the current flow. The pulses of infrared from the transmitter therefore produce pulses of leakage current through TR1, which give small negative pulses at TR1's collector. The output signal for TR1's collector is coupled by capacitor C2 to a high-gain inverting amplifier based on IC1. Resistors R2 and R5 are the negative-feedback network, and these set the closed-loop voltage gain of IC1 at 220 times. Capacitor C4 couples the amplified output signal from IC1 pin 6 to a conventional half-wave rectifier and smoothing circuit (D1, D2, and C5). Germanium, rather than silicon, diodes are used in the rectifier, because germanium types have a lower forward voltage drop. This gives slightly improved sensitivity. The positive dc signal developed across capacitor C5 drives the base of TR2, which is a simple common-emitter switch that controls the relay RLA coil.

MONO CORDLESS HEADPHONE IR RECEIVER

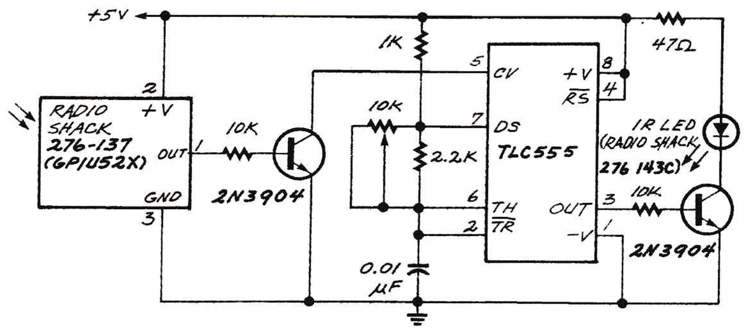


EVERYDAY PRACTICAL ELECTRONICS

Fig. 43-8

The full receiver circuit diagram appears in the figure. As many as four infrared detector diodes can be used (D1 to D4). They are wired in parallel and used in the reverse-biased mode, and under dark conditions only very small leakage currents flow. The pulses of infrared from the transmitter result in increased leakage currents, which generate small negative pulses at the cathodes (k) of the diodes. Transistor TR1 is used as a low-noise preamplifier and buffer stage. The majority of the voltage gain is provided by transistors TR2 and TR3, which are also common-emitter stages. These provide a combined voltage gain of over 80 dB. The signal from TR3 is connected to one of the phase comparator inputs of IC1 (pin 14). This IC provides demodulation and is a CMOS 4046BE p.l.d. Resistor R10 and capacitor C6 are the timing components for the VCO, and they set the center frequency at approximately 100 kHz. Resistor R12 and capacitor C7 form a simple low-pass filter between the phase comparator's output at pin 2 and the input of the VCO at pin 9. The audio output signal is obtained from the low-pass filter via an integral source-follower stage. The output from IC2 is fed to resistor R16 and capacitor C12, which form a simple low-pass filter that provides the deemphasis. The signal is then fed to the input of IC3, which is a small class-B audio power amplifier. The current consumption of the receiver is about 12 mA.

IR REMOTE-CONTROL REPEATER

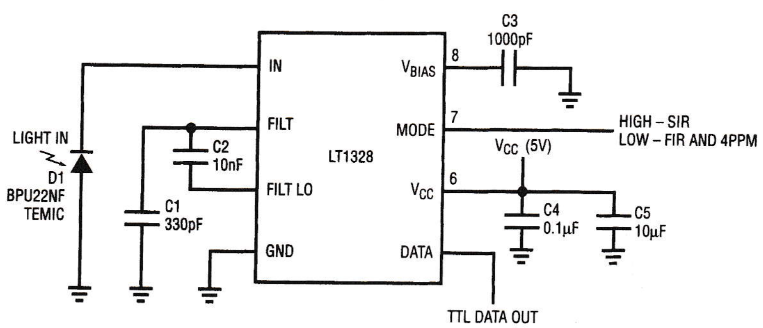


ELECTRONICS NOW

Fig. 43-9

This circuit receives TV remote-control signals and retransmits them. The 555 oscillator provides 40-kHz modulation. The IR LED can be placed in another room.

INFRARED DATA RECEIVER

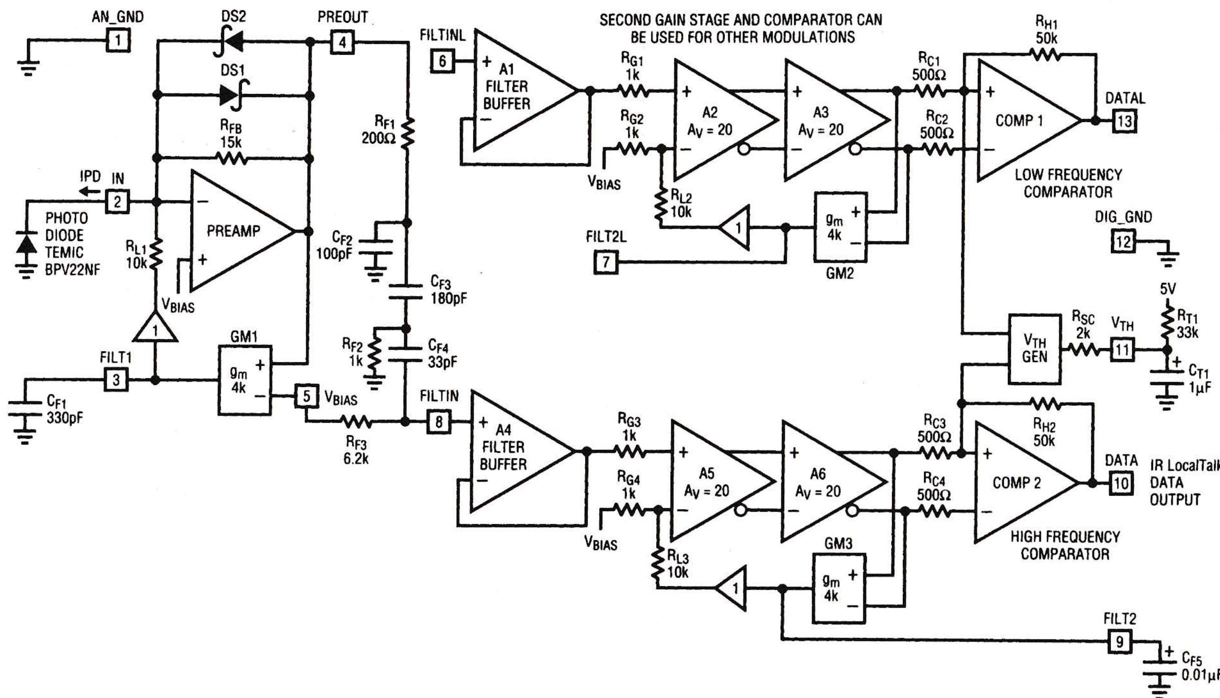


LINEAR TECHNOLOGY

Fig. 43-10

The LT1328 circuit operates over the full 1-cm to 1-m range of the IrDA standard at the stipulated light levels. For IrDA data rates of 115 kb/s and below, a 1.6-μs pulse width is used for a 0 and no pulse for a 1. Light levels are 40 to 500 mW/sr (milliwatts per steradian). The figure shows a scope photo for a transmitter input (top trace) and the LT1328 output (bottom trace). Note that the input to the transmitter is inverted; that is, transmitted light produces a high at the input, which results in a zero at the output of the transmitter. The MODE pin (pin 7) should be high for these data rates.

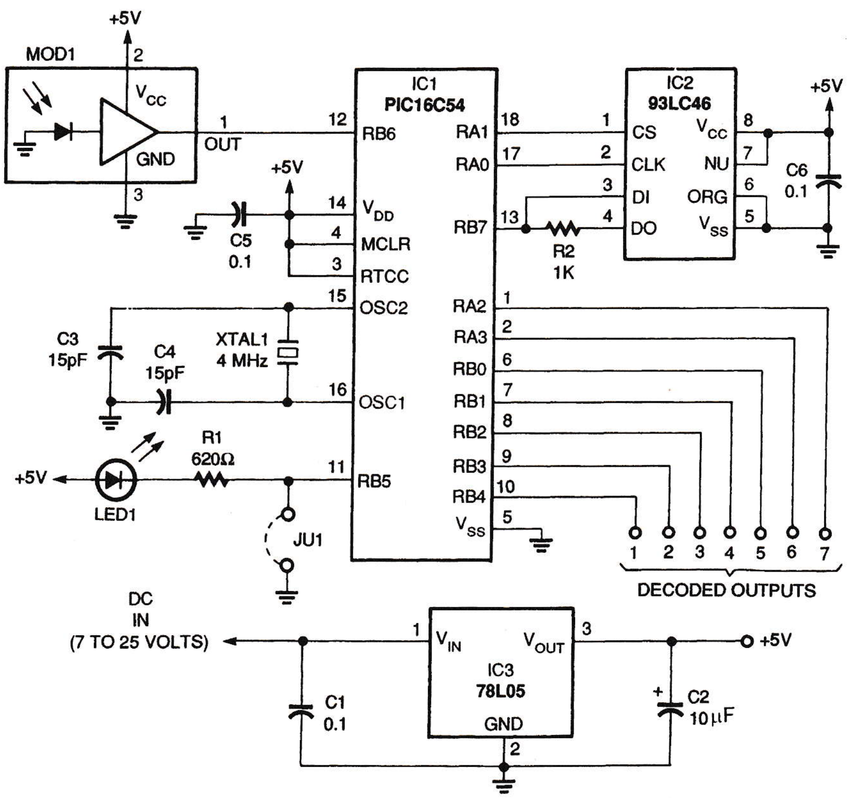
IR LOCAL TALK LINK RECEIVER



LINEAR TECHNOLOGY

Fig. 43-11

A low-noise (2 pA / $\sqrt{\text{Hz}}$), high-bandwidth (7 MHz) current-to-voltage converter formed by the preamplifier and its associated components transforms the reverse current from an external photodiode to a voltage. Although the 7-MHz bandwidth of the preamp supports 4-Mbit data rates, a low-pass filter on the preamp output is used to reduce the bandwidth to just the required amount in order to reduce noise. As shown, capacitance C_{F1} sets the break frequency of an ac high-pass loop around the preamp to 180 kHz. This loop rejects unwanted ambient light, including sunlight and light from incandescent and fluorescent lamps. The preamp stage is followed by two separate channels, each containing a high-impedance filter buffer, two gain stages, high-pass loops, and a comparator. The only difference between the two channels is the response time of the comparators: 25 and 60 ns. For the 125-ns pulses of IR LocalTalk, the 25-ns comparator with its active pull-up output is used. The low-frequency comparator with its open collector output (with 5 kΩ internal pull-up) is suitable for more modest speeds, such as the 1.6-μs pulses or the IrDA-SIR. Each gain path has an ac coupling loop similar to the one on the preamp. Capacitance C_{F5} sets the high-pass corner at 140 kHz for IR LocalTalk. The loops serve the additional purpose here of maintaining an accurate threshold for the comparators by forcing the dc level of the differential gain stages to zero. As the preamp output is brought out and the inputs to the two comparator channels are buffered, the user is free to construct the exact filter required for the application by the careful selection of external components. R_{F1}, C_{F2}, C_{F3}, R_{F2}, C_{F4}, and R_{F3} form a bandpass filter with a center frequency of 3.5 MHz and 3-dB points of 1 MHz and 12 MHz. Together with the high-pass ac loop in the preamp and the 7-MHz response of the preamp, this forms an optimal filter response for IR LocalTalk.

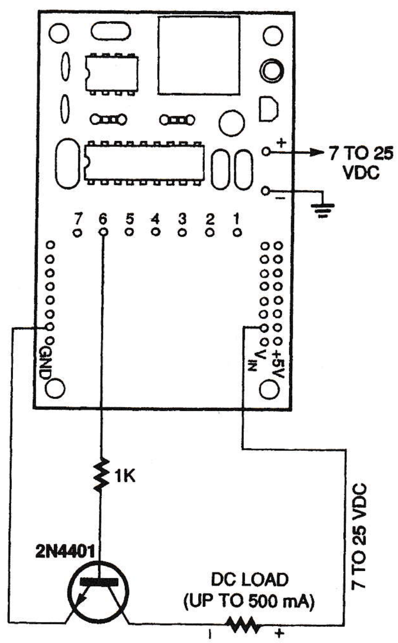


ELECTRONICS NOW

Fig. 43-12

A schematic diagram of the remote-control receiver is shown. The heart of the circuit is IC1, a PIC16C54 8-bit CMOS manufactured by Microchip. The microcontroller stores its data in IC2, a 93LC46 1-kbit serial EEPROM (electrically erasable programmable read-only memory), also manufactured by Microchip. In this application, the 93LC46 has a three-line interface with the microcontroller. The three lines are CHIP SELECT, CLOCK, and DATA IN/OUT. Because DATA IN and DATA OUT share the same line, a resistor (R2) limits the current flow during transitions between writing and reading when there are conflicting logic levels. The microcontroller communicates with the 93LC46 by placing a logic high on the CHIP SELECT pin. Data are then transferred serially to and from the 93LC46 on the positive transition of the CLOCK line. Each read or write function is preceded by a start bit, an opcode identifying the function to be performed (read, write, etc.), then a 7-bit address; this is followed by the 8 bits of data which are being written to or read from that address. Immediately preceding and following all write operations, the microcontroller sends instructions to the 93LC46, which enables or disables the write function, thereby protecting the data that have been stored. In the programming mode, IC1 reads an IR data stream from MOD1 and converts it to data patterns that can be stored in IC2. These data patterns are held for comparison while the unit is in normal operation. Power for the circuit is conditioned by IC3, a 78L05 low-current, 5-V regulator, which will accept any dc input voltage between 7 and 25 V. Capacitors C1 and C2 stabilize the operation of the regulator. Crystal XTAL1 sets the internal oscillator of IC1 to 4 MHz. Jumper JU1 consists of two closely spaced pads on the PC board that, when momentarily jumpered with a screwdriver or other piece of metal, places IC1 in the programming mode and lights LED1. The source and object code are available on the Gernsback BBS (516-293-2283, v.32, v.42bis) as a file called IREC.ZIP for those who wish to program their own PICs and have the proper equipment to do so.

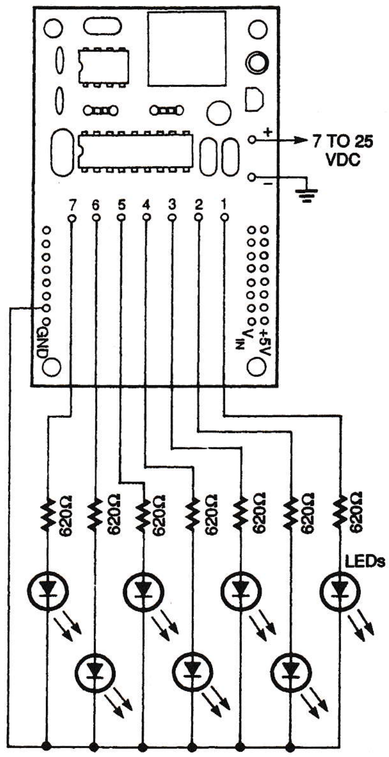
IR REMOTE-CONTROL DC INTERFACE



ELECTRONICS NOW ***Fig. 43-13***

This circuit can be used to interface dc loads up to 500 mA to the IR remote control.

IR REMOTE-CONTROL TEST SET

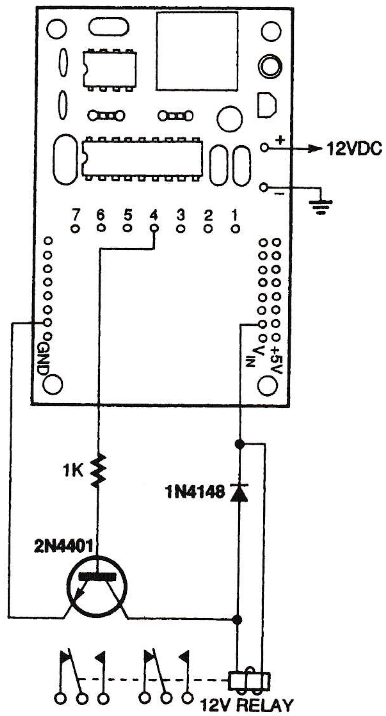


ELECTRONICS NOW

Fig. 43-14

This circuit can be used to check out operation of the IR remote control.

IR REMOTE-CONTROL RELAY INTERFACE

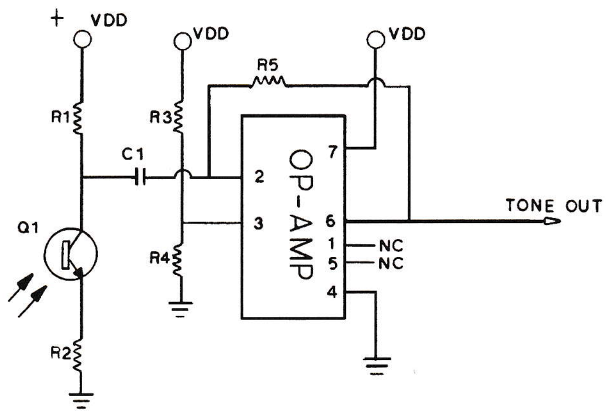


ELECTRONICS NOW

Fig. 43-15

This circuit can be used to interface a relay to the IR remote control in order to control large ac or dc loads.

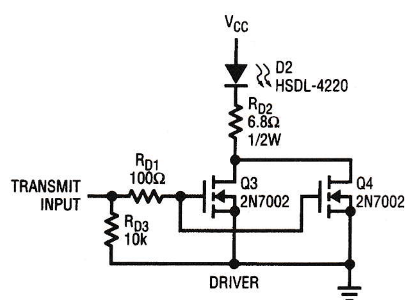
PHOTOTRANSISTOR IR RECEIVER



NUTS AND VOLTS *Fig. 43-16*

This receiver uses an op amp and a phototransistor as a receiver for IR signals. Q1 is a phototransistor. The op amp is a CMOS or FET, such as a TL081, etc. R1 and R2 depend on the phototransistor, but are typically 2.2 kΩ and 330 Ω, respectively. R3=R4=10 kΩ, R5=100 kΩ, and C1 is 0.1 μF (or larger).

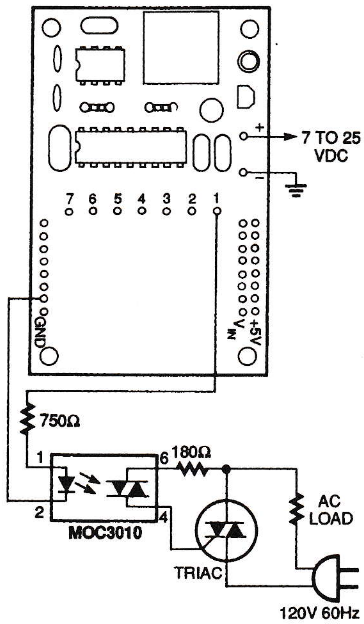
INFRARED DATA TRANSMITTER



LINEAR TECHNOLOGY *Fig. 43-17*

This figure shows an IrDA transmitter.

IR REMOTE-CONTROL TRIAC INTERFACE

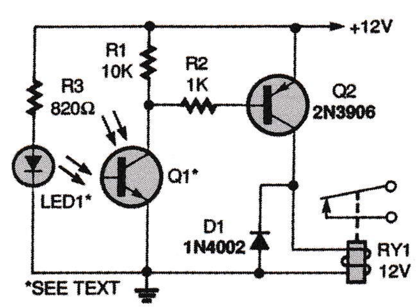


ELECTRONICS NOW

Fig. 43-18

This circuit can be used to interface a triac to the IR remote control in order to control ac loads.

SEE-THROUGH SENSOR

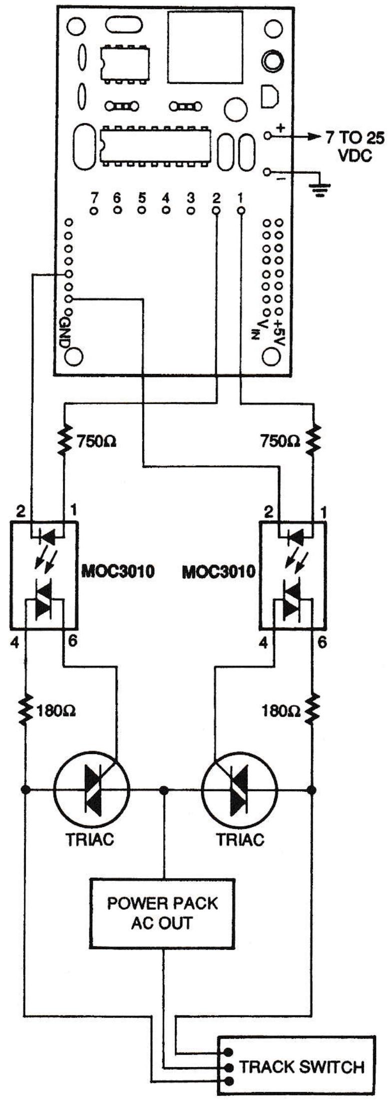


POPULAR ELECTRONICS

Fig. 43-19

In this circuit, an infrared emitter LED (LED1) is aimed at an infrared phototransistor (Q1). As long as the IR light path between the two remains uninterrupted, transistor Q2 will keep relay RY1 closed. Any opaque object blocking the light path will cause RY1 to open. This circuit is a see-through-type light sensor. Such units are often used as part-in-place detectors and parts-counter sensors.

IR REMOTE-CONTROL MODEL RAILROAD APPLICATION

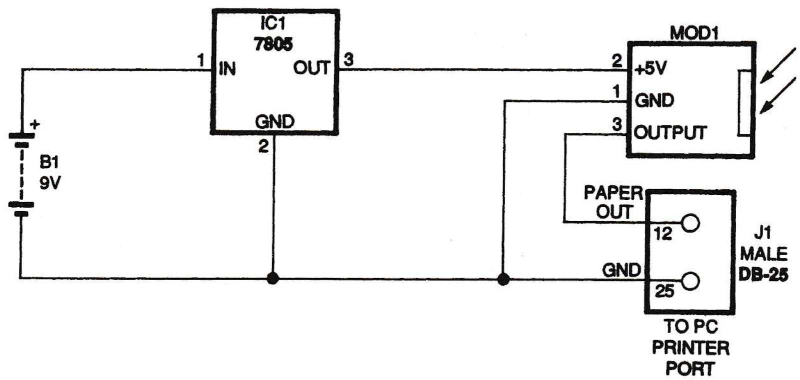


ELECTRONICS NOW

Fig. 43-20

This circuit can be used to interface two triacs to the IR remote control in order to control model railroad track switches.

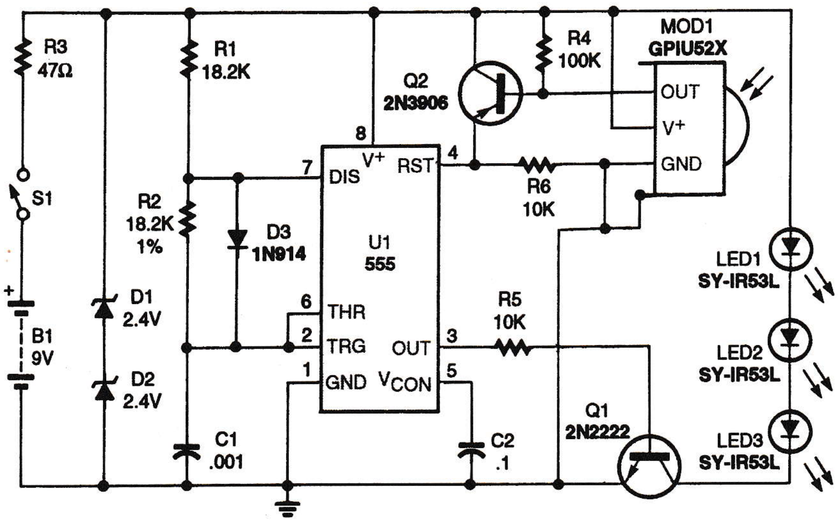
IR REMOTE-CONTROL RECEIVER



ELECTRONICS NOW ***Fig. 43-21***

The figure shows a schematic of the IR receiver circuit. The heart of the circuit is MOD1, an infrared detector module that removes the IR carrier frequency and transmits only the data that are encoded in the received IR signal. A suitable IR module is available at Radio Shack (No. 276-137). The IR module needs a clean 5-V power supply, which is provided by IC1, a 7805 regulator. Power is supplied to the regulator by 9-V battery B1. The output of the module is wired to a male DB-25 multipin connector. The infrared detector module receives a signal, filters it, and removes the 40-kHz carrier. The output of the module is a TTL-level signal consisting of long and short pulses. The PC records those voltage levels over time, while the signal is being sent, and stores the data in a file. The line normally used by the PC's printer port to indicate that the printer is out of paper (pin 12) is used in this project to accept data from the IR module. The I/O port is located at the address ox379. Bit 5 corresponds to input pin 12. Various software programs are required to let a PC store information input to its printer port. (All of the software is available on the Gernsback BBS—516-293-2283, v.32, v.42bis—contained in a file called IR-TEST.ZIP.) The source code of the first program, IRLOG.EXE, is written in C. The program stores the value it reads from the PC's printer port into an array. When the input line is logic high, the ASCII character 1 is stored in the array. When the input line is logic low, ASCII character 0 is stored.

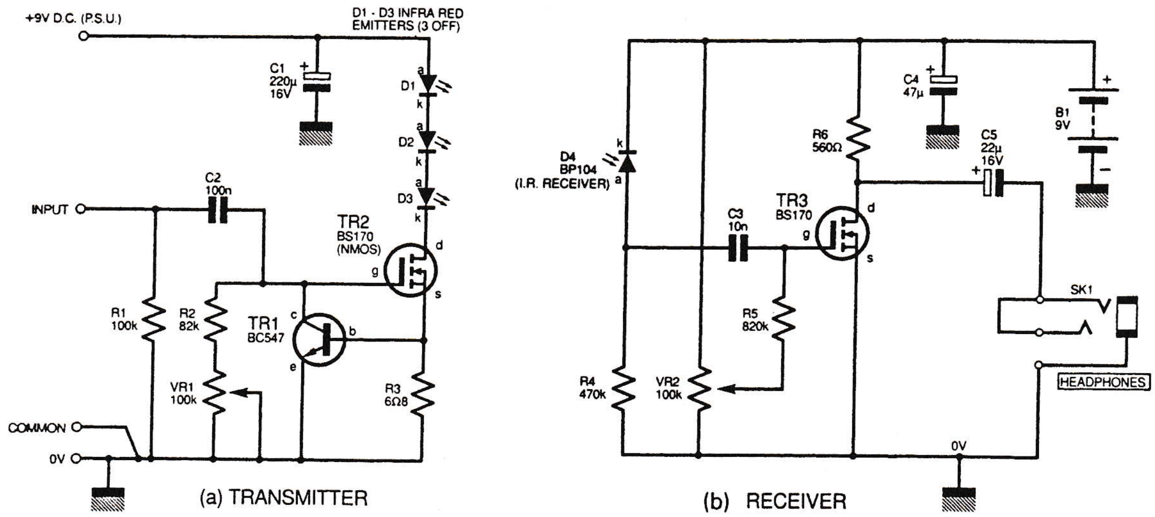
IR RELAY CIRCUIT



POPULAR ELECTRONICS ***Fig. 43-22***

All IR is received by the module via an IR photodiode. The diode is operated in its reverse-bias current it permits depends on the intensity of the received infrared light. The fluctuations in the reverse-bias current are then amplified by a high-gain stage. The output of the amplifier is then "limited" by the next stage. The limiter chops the extreme highs and lows off the amplified signal, and the result is a quasi-digital pulse train. The simplified wave then passes through a bandwidth filter that has its center frequency at 40 kHz. At that point, the circuit has effectively retrieved the 40-kHz remote carrier. The reproduced carrier is then integrated. The next stage is an "inverting Schmitt trigger." It will not go low unless the filter's output signal surpasses a certain amplitude, and it will not go high again until the signal drops below a certain minimum. Thus, the Schmitt trigger responds only to large changes in the filter's output (caused by bursts and pauses) and ignores small changes (caused by the 40-kHz carrier, to which the filter can't respond quickly). The Schmitt trigger's output is thus low when a 40-kHz burst is received, and high during pauses between bursts. The resulting waveform is an inverted version of the pulses that were modulated and transmitted by the remote. The oscillator circuit is based on a 555 timer. With D3, R2 is bypassed (or shorted out) while C1 is charging, but is in the current path during discharge. Therefore, with R1 and R2 made equal, the charge time equals the discharge time, yielding an output with a 50-percent duty cycle. Components R1, R2, and C1 (all precision, drift-free units) have been chosen to provide a 40-kHz output in this configuration. That output strobes the IR LEDs via Q1. The oscillator functions only when pin 4 of the 555 is high. Because that pin is connected to the inverter circuit, the oscillator functions when the inverter's input is low. The inverter's input is connected to MOD1's output, which goes low with each remote burst received. So overall, the circuit produces a 40-kHz IR burst when it receives one.

CORDLESS HEADPHONES



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 43-23***

It is possible to use headphones with no direct connection to the audio source. This figure illustrates a simple transmitter circuit suitable for connecting to an audio source. Diodes D1 to D3 are three infrared LEDs in series, driven by TR2, an n-channel MOSFET with a drain-current rating of 500 mA maximum. The illumination of the LEDs is modulated by the audio signal applied, with TR1 current-limiting and shunting the bias of TR2 to ground when TR2 source current exceeds roughly 100 mA. Potentiometer VR1 should be adjusted for best results. A 9-V power supply is best used in this circuit because of the sustained level of current consumption. The transmitter range is about 1 to 2 m, but this can be extended by using reflectors behind the diodes.

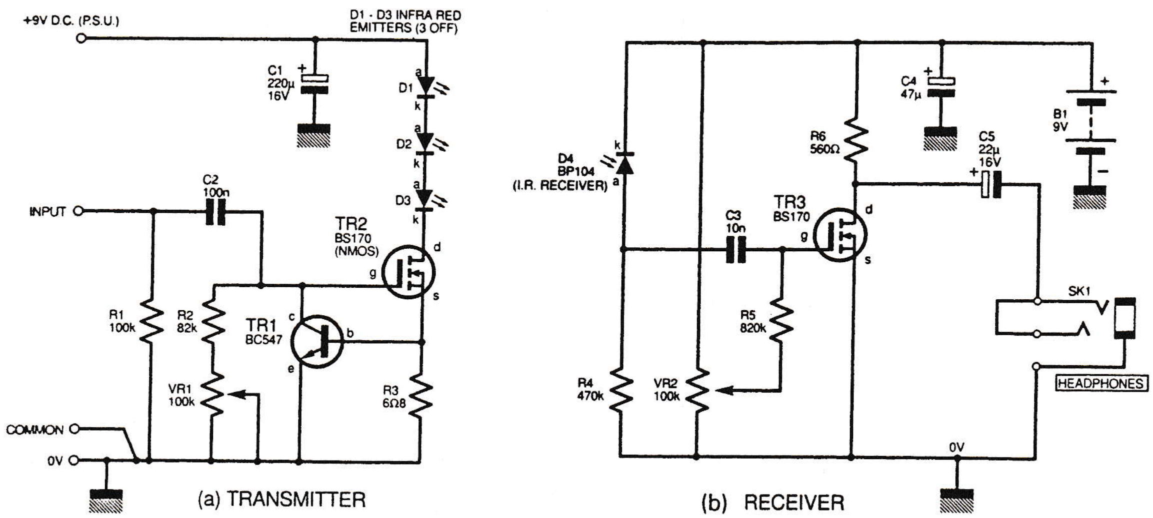
A suggested receiver circuit, using an infrared photodiode, D4, to detect the IR light, is shown in (b). The received signal is ac-coupled to TR3, another MOSFET. The output is taken from the drain terminal via capacitor C5. Bias can be adjusted with potentiometer VR2, and this circuit will run from a PP3-size battery. The circuit could be adapted for other applications.

44

Inverter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

CORDLESS HEADPHONES



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 43-23***

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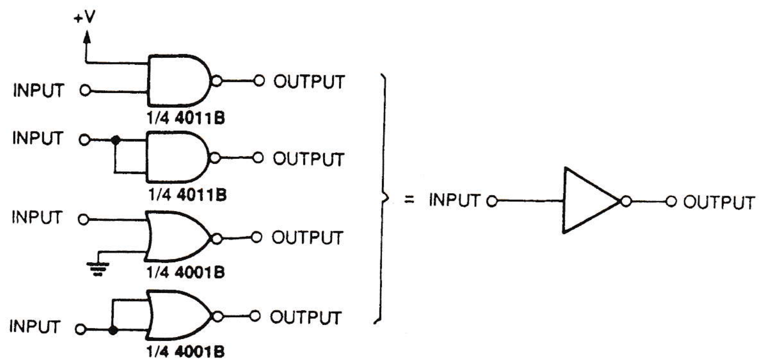
A suggested receiver circuit, using an infrared photodiode, D4, to detect the IR light, is shown in (b). The received signal is ac-coupled to TR3, another MOSFET. The output is taken from the drain terminal via capacitor C5. Bias can be adjusted with potentiometer VR2, and this circuit will run from a PP3-size battery. The circuit could be adapted for other applications.

44

Inverter Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

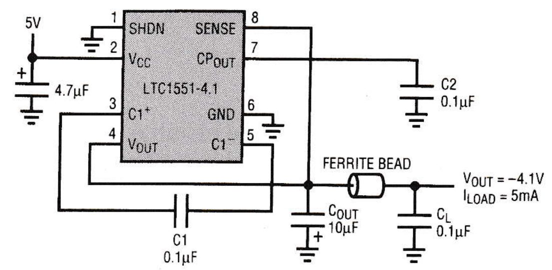
GATE INVERTERS



WILLIAM SHEETS ***Fig. 44-1***

Logic gates can be configured as buffers or inverters as shown.

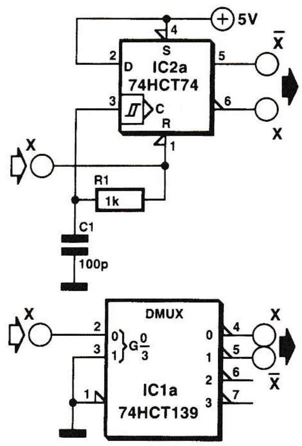
REGULATED LOW-NOISE VOLTAGE INVERTER



LINEAR TECHNOLOGY POWER SOLUTIONS ***Fig. 44-2***

Analog cell phones rely on a quiet GaAsFET bias supply to maximize the signal-to-noise ratio, providing a high-quality transmission. The LTC1551CS8-4.1 switched-capacitor voltage inverter circuit shown reduces output noise to less than 1 mV on the −4.1-V regulated output voltage. The 900-kHz operating frequency allows the charge pump to use small 0.1-μF capacitors. The 10-μF and 0.1-μF output capacitors effectively reduce the 900-kHz ripple noise to relatively insignificant levels. The LTZ1550CS8-4.1 has an active low shutdown input (SHDN), whereas its counterpart, the LTC1551CS8-4.1, has an active high shutdown input. This choice minimizes component count. The LTC1550 and 1551CS8-4.1 both operate in doubler mode, meaning that they only invert the input voltage and regulate it down to −4.1 V. The input voltage range is 4.5 to 7 V.

D-BISTABLE



ELEKTOR ELECTRONICS ***Fig. 44-3***

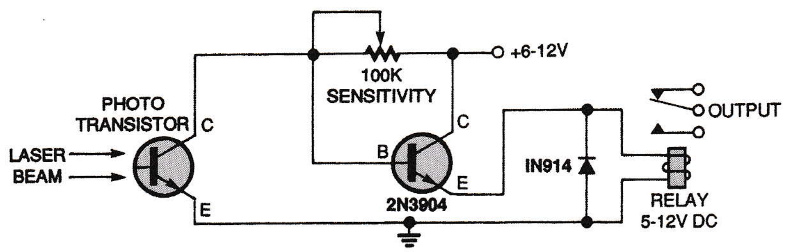
The interesting point of this circuit is that a D-type bistable is used as an inverter. When the level at the input changes from high to low, the bistable is reset and its *Q* output goes high. When the input becomes low, the reset is removed and the *Q* output goes low. The delay introduced by network R1-C1 between the RESET input and the CLOCK input makes it possible to trigger the bistable at the leading edge of the input signal. As an example, in the case of a dual D-bistable Type 74HCT74, the time needed for a clock pulse to be accepted after the reset has been removed is 5 ns. Therefore, an RC introducing a delay of 7.5 ns gives a reasonable safety margin. The reduced edge gradient of the clock pulse does not create any problems because the maximum allowed rise time of the clock input is 500 ns. To obviate asymmetrical output signals, it is advisable to limit the input frequency to about 1 MHz with component values as specified.

45

Laser Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

LASER-BEAM-ACTIVATED RELAY

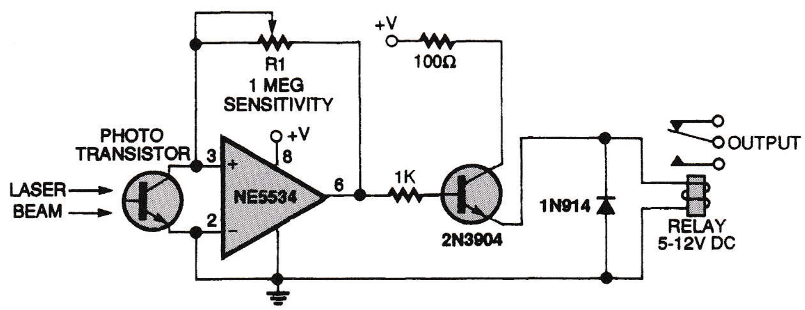


ELECTRONICS NOW

Fig. 45-1

This two-transistor relay circuit will activate whenever the phototransistor is illuminated by the laser beam.

LASER-ACTIVATED RELAY WITH AMPLIFIER



ELECTRONICS NOW

Fig. 45-2

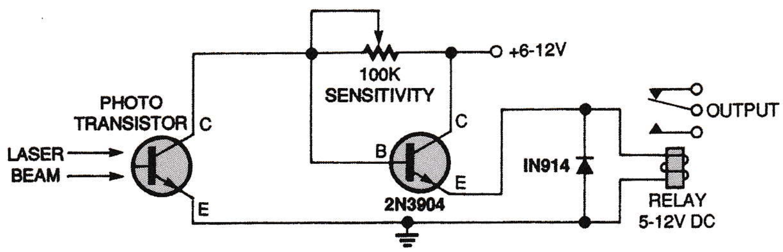
A low-noise amplifier is used in conjunction with a phototransistor to allow higher sensitivity than the detector alone would provide.

45

Laser Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

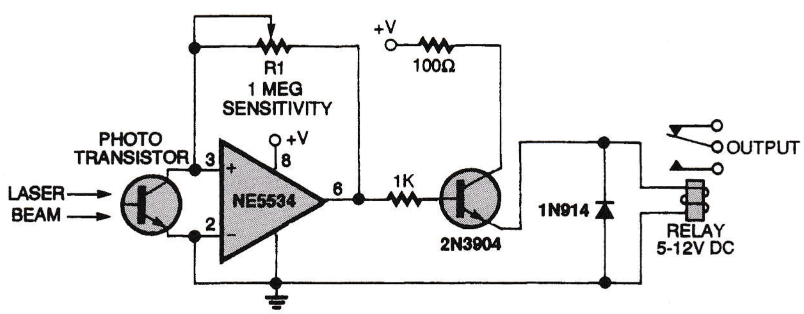
LASER-BEAM-ACTIVATED RELAY



ELECTRONICS NOW ***Fig. 45-1***

This two-transistor relay circuit will activate whenever the phototransistor is illuminated by the laser beam.

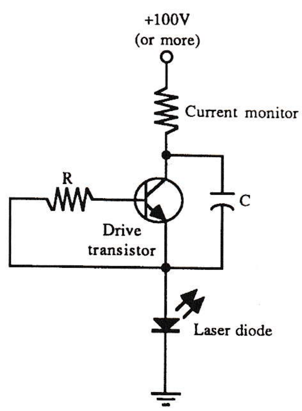
LASER-ACTIVATED RELAY WITH AMPLIFIER



ELECTRONICS NOW ***Fig. 45-2***

A low-noise amplifier is used in conjunction with a phototransistor to allow higher sensitivity than the detector alone would provide.

LASER-DIODE DRIVER

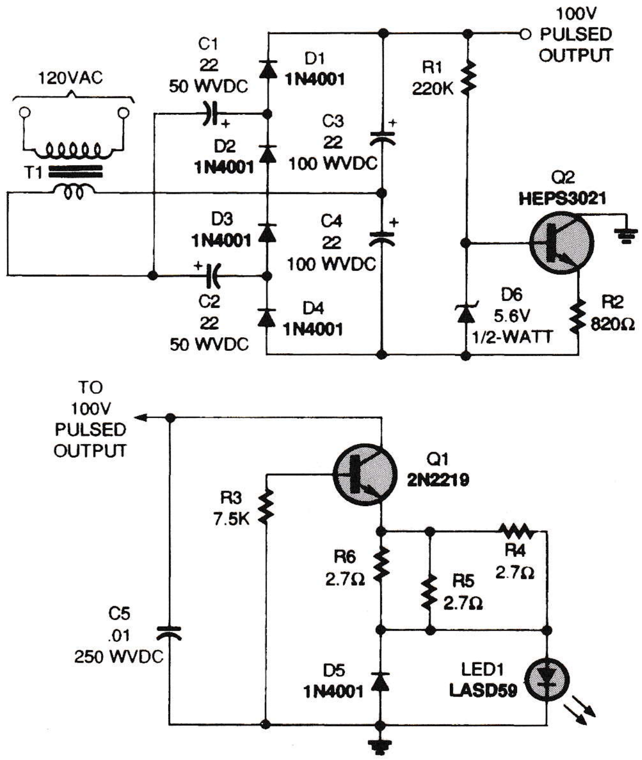


TAB BOOKS

Fig. 45-3

One way to drive a single heterostructure laser diode. The transistor is driven in avalanche mode, producing short-duration pulses of current.

LASER-DIODE DRIVER

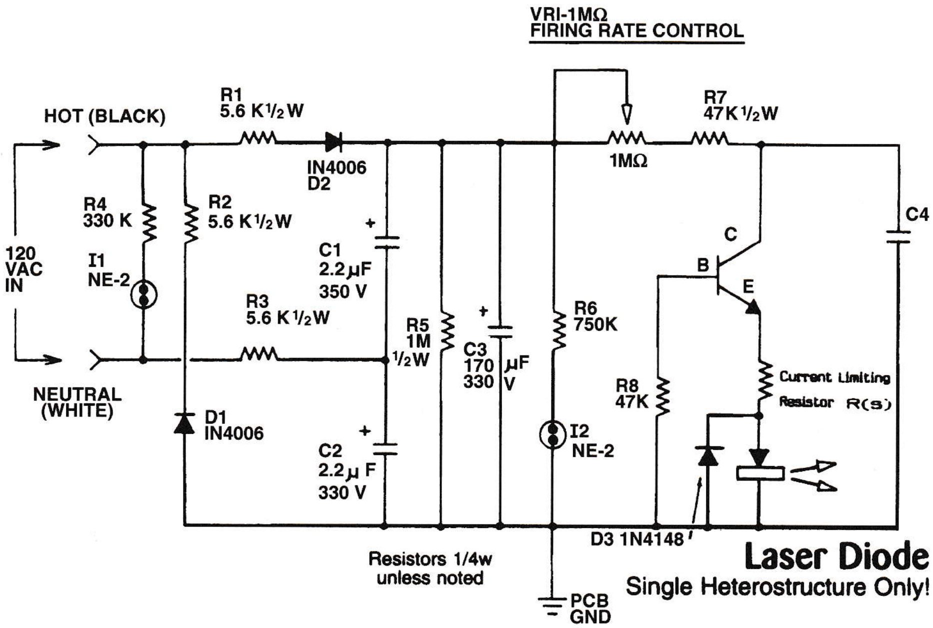


POPULAR ELECTRONICS

Fig. 45-4

T1, a 24-Vac wall transformer, drives a full-wave quadrupling rectifier, which produces an output of 120 Vdc. Components R1, R2, D6, and Q2 form a 5.7-mA constant-current sink. Use a heatsink on Q2, which can be any 1-W or higher transistor with a V_{CE} of 150 V or more. Components C5, R3 to R6, D5, LED1, and Q1 form an avalanche oscillator that runs at 4 to 5 kHz. Transistor Q1 must be selected for a breakdown voltage of between 90 and 110 Vdc. Build the power-supply section on a separate circuit board using standard methods. Resistors R4 to R6 must be carbon, ½-W units, or can all be replaced by a single 1-W, 1-Ω carbon unit. Use short leads throughout on this assembly to avoid current undershoot. Use a socket to mount Q1, and a female pin for the anode of LED1.

120-VAC SINGLE HETEROSTRUCTURE LASER DRIVER SUPPLY

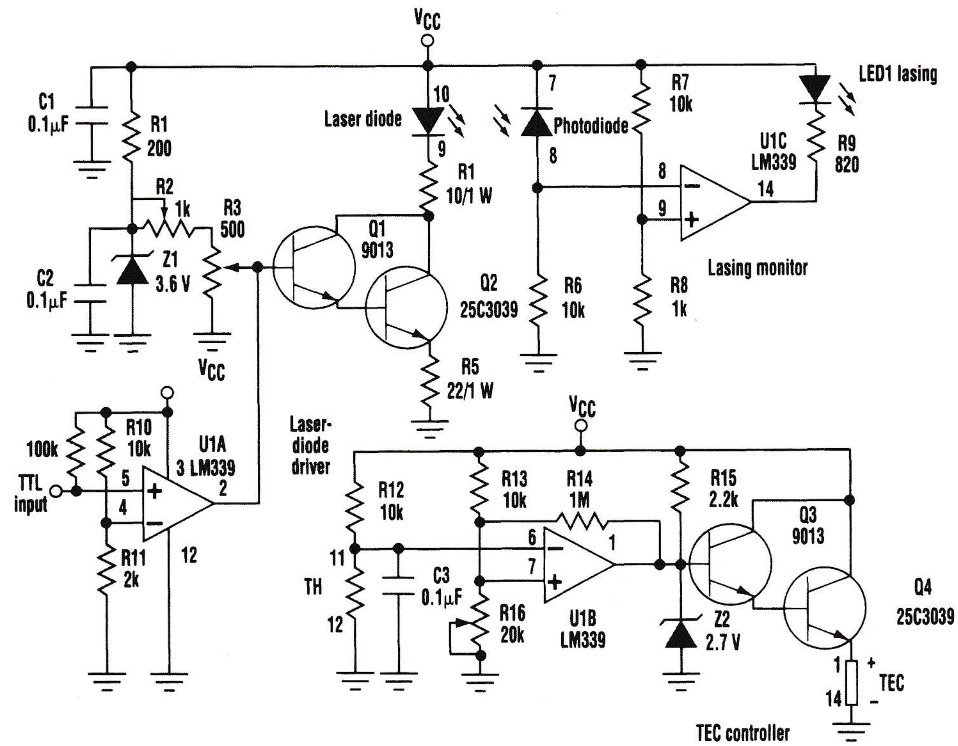


NUTS AND VOLTS

Fig. 45-5

This figure shows the complete schematic of the 120-Vac SH power supply. The circuit consists of a polarized plug, a voltage doubler, a large-value storage capacitor, two neon indicator lamps to show that ac and dc power is present, a 1-MΩ firing-rate control, and the "business end" of the supply to drive the laser. I2 is also a safety light. When it is on, the laser should be considered operating. The current to the voltage doubler is limited by resistors R1 to R3, and by the capacitive reactance of capacitors C1 and C2. R7—a 47-kΩ, ½-W resistor—limits the input power to the avalanche circuit to avoid overheating the diode or Q1. The capacitors and resistors serve to isolate the circuit from the ac line sufficiently to prevent a shock hazard; however, you can still get shocked. Be sure you put this device in a nonconductive case, and obey the polarity of the polarized plug. Don't touch any part of the circuit while it's plugged in, either. Be sure you use a plastic potentiometer for VR1, or a separate control-type potentiometer with a plastic knob.

LOW-COST LASER-DIODE DRIVER

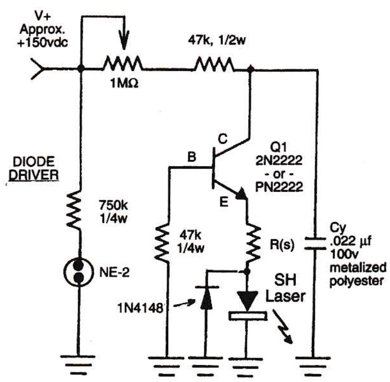


Note: The laser diode, photodiode, TH, and TEC come in one package (QLM5S876 laser-diode module).

ELECTRONIC DESIGN *Fig. 45-6*

The circuit presented is a low-cost laser-diode driver with current limiting and a lasing monitor for safe operation of the laser. The driver provides temperature stabilization using a built-in thermo-electric-cooler (TEC) controller. This circuit is designed around a QLM5S876 1.55-m laser-diode module with a built-in monitor photodiode, TEC, and thermistor. The laser is driven by the constant-current source Q2 and Q1. The drive current is stabilized against supply changes by Z1, limited by R2, and adjusted by R3 to the desired operating point. When the current is small, the laser doesn't lase and the monitor photodiode detects no optical power. As a result, comparator U1C goes below its trip point and LED1 turns off. When the current is adjusted and passes the laser threshold, sufficient optical power is generated by the laser, causing photo current to flow through the photodiode and the comparator to drive the lasing monitor indicator LED1 to on. The TEC controller circuit is a feedback system. If the temperature is higher than the set point, the comparator output will be high and will drive the TEC element through R15, Q3, and Q4. The TEC drive current is limited by Z2. When the TEC is driven on, it cools the laser. This increases the voltage of the inverting output until it passes the comparator's upper trip point and turns off the TEC's drive current. The temperature set point can be adjusted by R16.

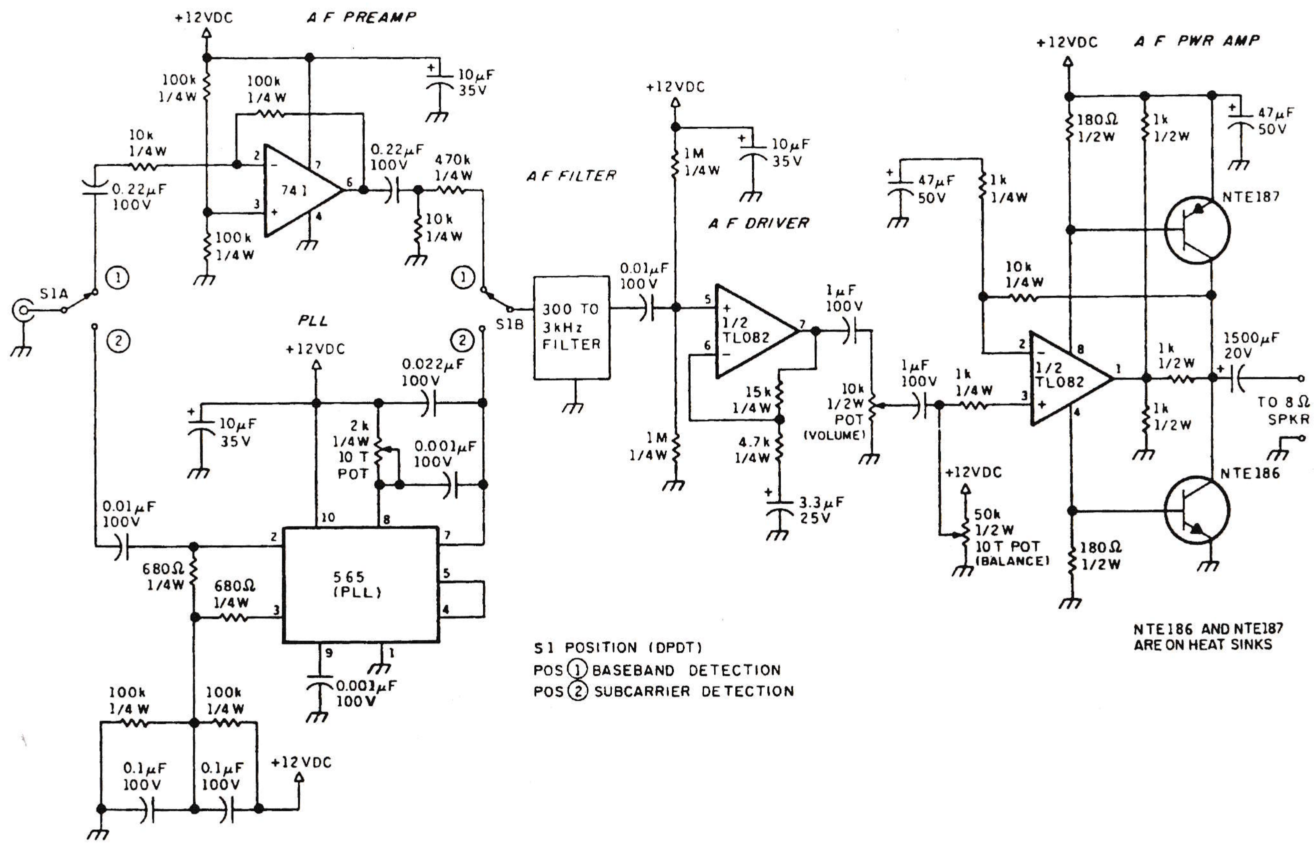
SINGLE HETEROSTRUCTURE LASER DRIVER



NUTS AND VOLTS ***Fig. 45-7***

One way to reliably achieve the fast rise-time pulse of short duration to power SH lasers is to use the phenomenon called *avalanching* of a transistor. Transistors have a breakdown voltage, abbreviated as V_{ceo}; if this voltage is exceeded, the transistor will spontaneously conduct with its collector-to-emitter circuit open and the voltage applied to its collector. This produces a pulse whose rise time is extremely fast—on the order of 1 or 2 ns—and whose duration is related to the capacitance at the collector. The operation is as follows: Current rises at the base of transistor Q1 when the voltage on Cy has risen enough to jump over and break down the collector-to-emitter junction. For the 2N2222, this is typically 75 to 80 V. When the transistor breaks down, it sends a high-current pulse of energy through a low-value carbon film resistor R(s). With the 0.022-µF capacitor shown, the duration of the pulse to the laser is only about 75 ns. This capacitor must be specifically designed for low inductance. Note the high-speed switching diode D1. Its purpose is to damp out any reverse voltage that might occur in the circuit due to inductance after the laser turns off because a reverse voltage of only 3 V can destroy the SH laser. The rise time of this diode is comparable to that of the avalanche circuit, being only 6 ns before it begins to switch. The values of the capacitor and the current-limiting resistor R(s) depend upon the type of SH laser diode used.

LASER RECEIVER DETECTOR AND AUDIO CIRCUIT



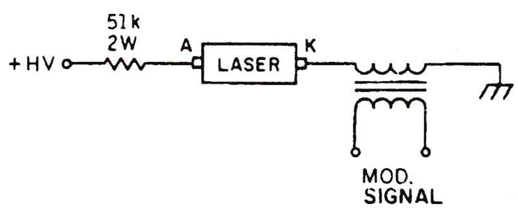
S1 POSITION (DPDT)
POS 1 BASEBAND DETECTION
POS 2 SUBCARRIER DETECTION

Fig. 45-8

LASER RECEIVER DETECTOR AND AUDIO CIRCUIT *(Cont.)*

Video preamplifier output can be switched between an audio amplifier that establishes the correct level of the demodulated baseband audio when used for short distance test links and a phase-locked loop (PLL) that demodulates an FM subcarrier when used for longer distances. The PLL VCO is set to a 100-kHz subcarrier frequency. The error signal generated as the VCO tracks the FM signal produces the demodulated audio. Either of these outputs is sent to a 300- to 3000-Hz bandpass filter and on to an audio power amplifier capable of driving a low-impedance speaker or headphones. The dc power supply for these circuits is provided by a filtered 12 Vdc input.

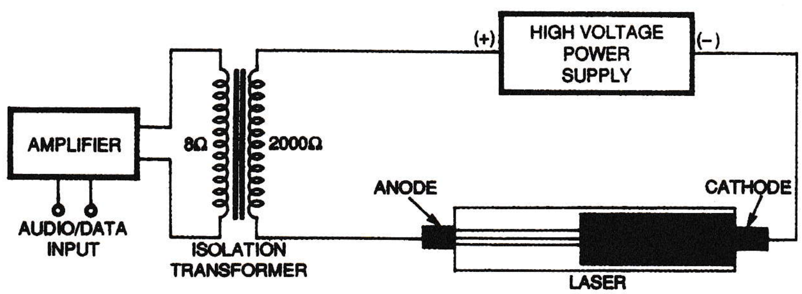
LASER CURRENT MODULATION



COMMUNICATIONS QUARTERLY *Fig. 45-9*

Depending on laser characteristics, only a small percentage of modulation (around 10 percent) is possible with this scheme.

LASER MODULATION



ELECTRONICS NOW *Fig. 45-10*

Electronic modulation of a laser beam can be handled in this fashion. Obviously, the transformer you use must be able to handle the high voltage from the laser power supply without breaking down.

LASER TRANSMITTER

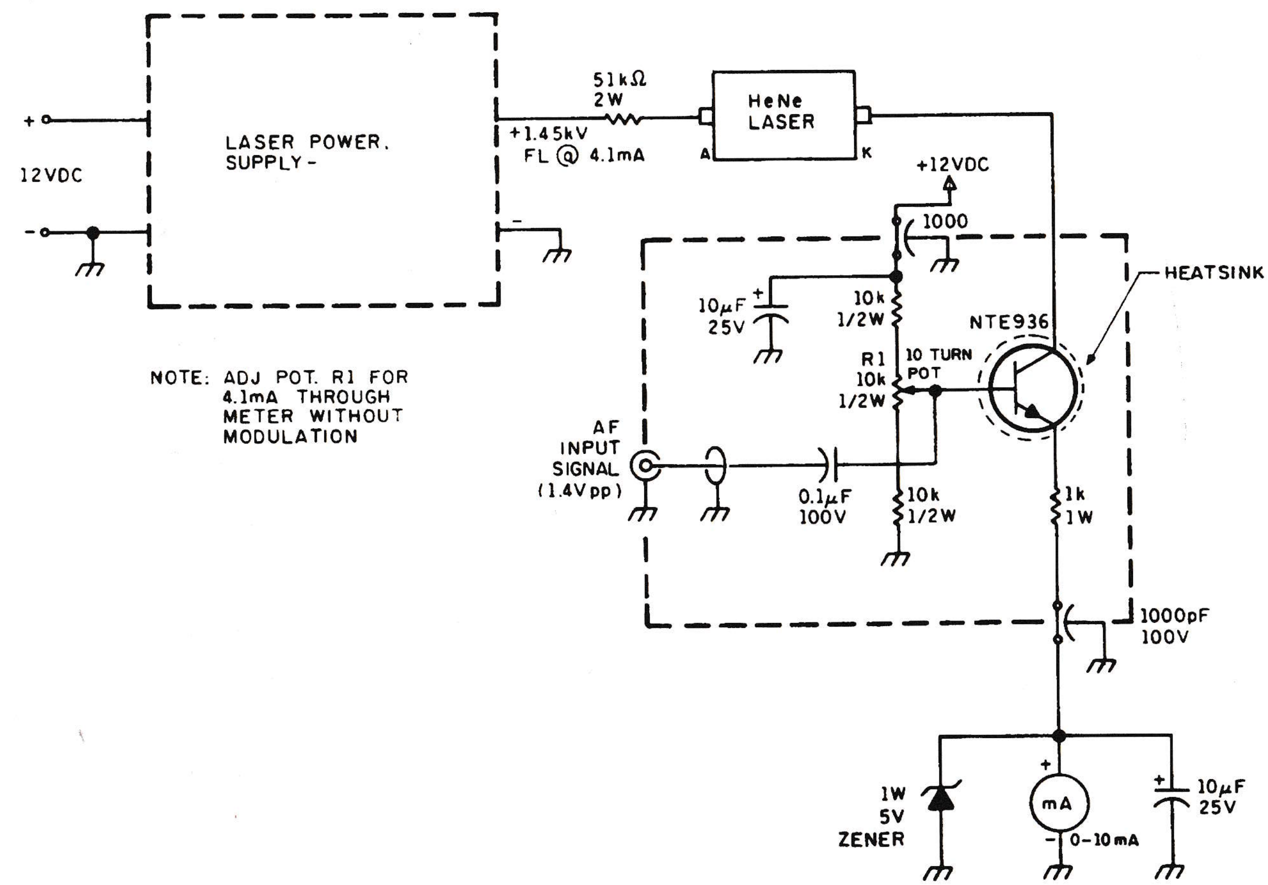
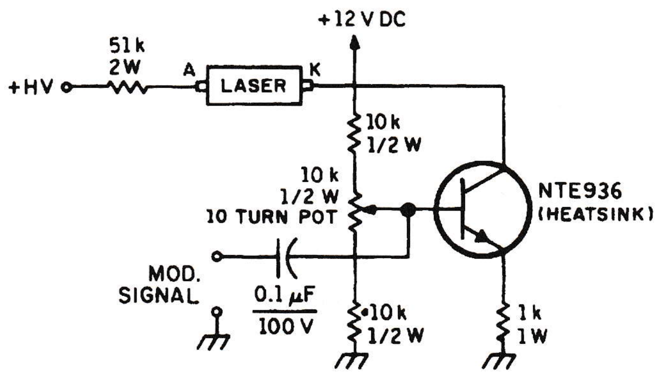


Fig. 45-11

A current modulator is connected to the cathode of the laser. The quiescent current is set at 4.1 mA, and the modulating signal is set to vary the current only by ±0.7 mA. A well-protected and bypassed 0- to 10-mA meter is used in the ground leg to set and monitor the laser's quiescent operating current. A 51-kΩ, 2-W ballast resistor close to the anode lead of the laser stabilizes the discharge. The voltage across the sustaining supply should be 1.45 kV dc at full load. Remember that the subcarrier frequency must be no greater than 200 kHz, and that the peak-to-peak signal voltage into the modulator must not exceed 1.4 V p-p. (There are any number of ICs that operate as voltage controlled oscillators to produce the FM subcarrier.)

LASER MODULATION WITH CURRENT SOURCE



COMMUNICATIONS QUARTERLY ***Fig. 45-12***

Because atmospheric turbulence can become bothersome, any low-frequency baseband intensity modulation will become corrupted over long distances. To overcome this limitation, the information must be frequency-modulated up to 200 kHz in this manner. To obtain a more desirable current modulator for subcarrier modulation, use a transistor as a variable-current source to vary the current through the laser. Here the quiescent operating current through the laser is set at 4.1 mA by a potentiometer in the base bias circuit. Assuming a 0.7-Vdc drop across the transistor base-emitter junction and a 1000-Ω resistor in the emitter lead, the quiescent bias voltage from base to ground is 4.8 Vdc. The input voltage must vary 1.4 V p-p to cause a ±0.7-mA current variation through the laser. To effect this, the modulation signal source must provide up to a 1.4-V p-p signal into the laser current modulator circuit.

PHOTOMULTIPLIER LASER RECEIVER WITH VIDEO AMPLIFIER

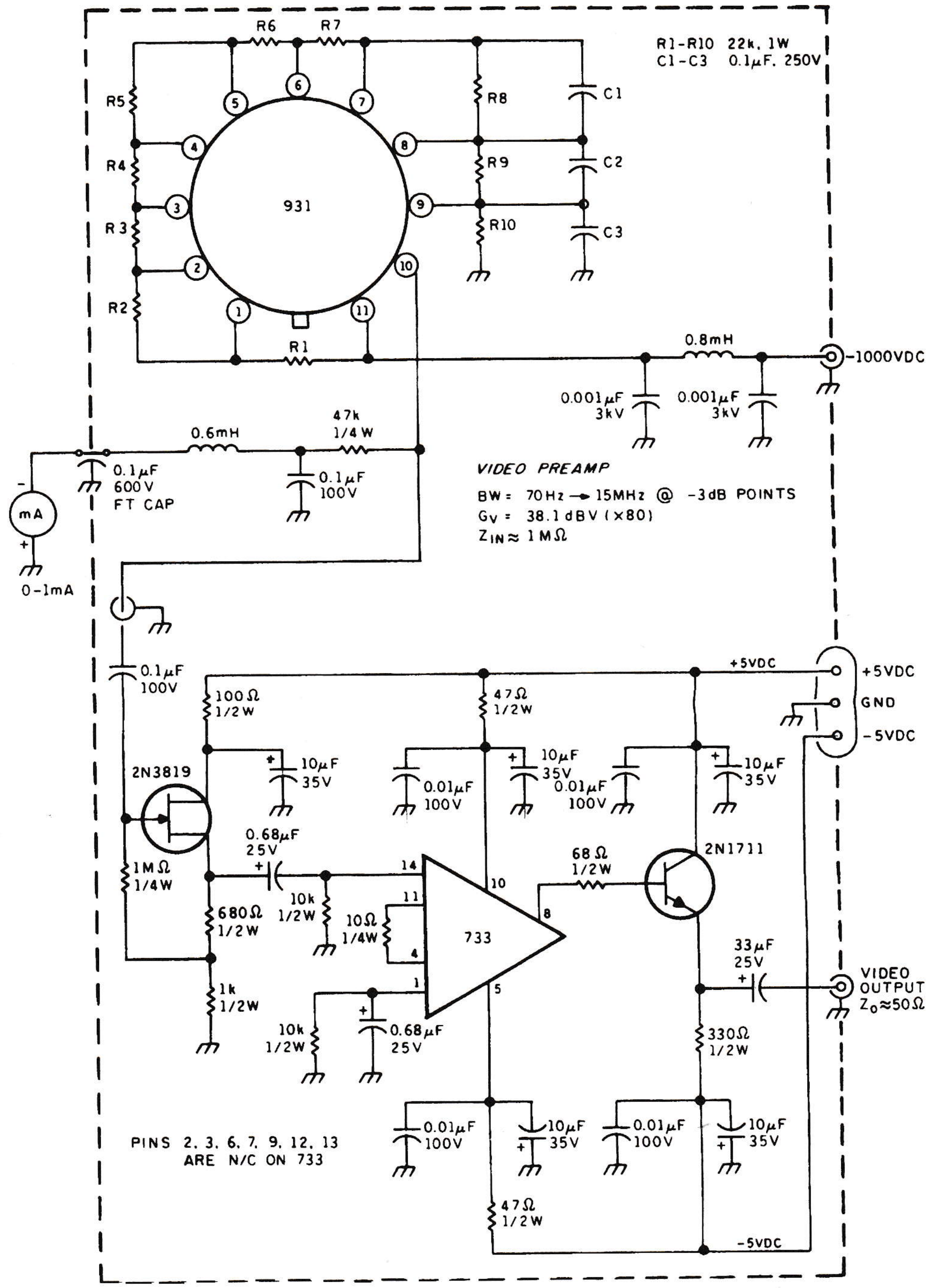


Fig. 45-13

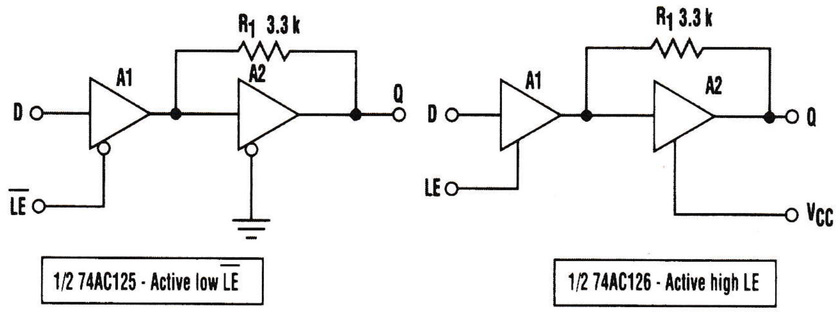
This is a circuit diagram for the inexpensive 931 side-looking PMT and its resistor divider network. This circuit maintains the proper voltages for the dynodes. The capacitors across the last three network resistors improve the frequency response of the PMT to modulated signals. A milliammeter in the PMT anode circuit registers the average PMT current under the medium to high illumination encountered during diagnostic tests. A PMT preamplifier uses a 733 (or equivalent) wideband amplifier. Because the demodulated signal can be either a direct baseband signal or a baseband signal modulated onto a subcarrier, the preamplifier must be capable of amplifying all the demodulated signal frequencies. The preamplifier is designed for a 70-Hz to 15-MHz bandwidth at the −3-dB points and for a gain of 80. Because the signal detection circuits will probably be placed in a separate housing, the video preamplifier is designed to drive a 50-Ω cable. With the high-current multiplication inherent in the PMT, the PMT noise will predominate over the video preamplifier IC noise.

46

Latch Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

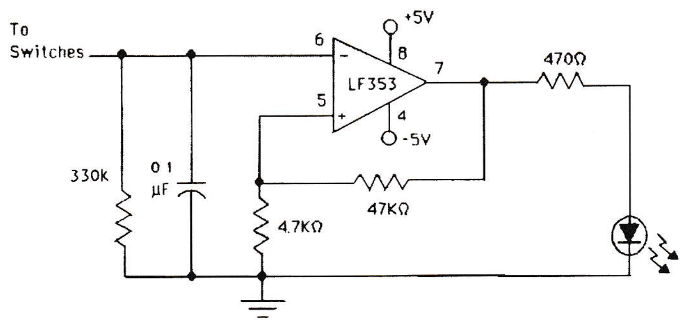
TRANSPARENT LATCH

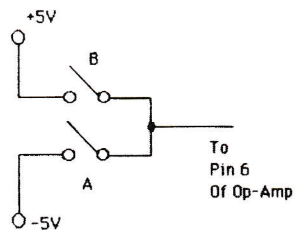


ELECTRONIC DESIGN *Fig. 46-1*

The circuit depicted here, which yields two independent latches per chip, is rather simple and inexpensive to build (see the figure). The second buffer A2 is wired with resistor R1 in the feedback path (ignore A1 for the moment). A2 with feedback resistor R1 is a stable latch. Because of the very low input current requirements of A2, there is hardly any voltage drop across R1. As a result, the input is the same as the output, and that is fed forward through the buffer, maintaining a stable level. When buffer A1 is enabled, the input of A2 is driven to the same level as D. Even if A2's output (Q) is at an opposite logic level, which can happen momentarily (for a gate delay) when D is opposite to Q, A1 is required to sink or source current through R1. After a gate delay, this signal propagates to Q. Both sides of R1 are now at the same logic level. When the LE signal goes inactive, A1 tristates and the latch will hold the level present at D one setup time prior to LE's going inactive.

PB SWITCH-ACTIVATED LATCH





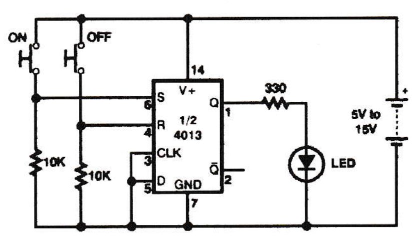
A	B	LED	
ON	OFF	ON	Last Input
OFF	ON	OFF	

TRUTH TABLE

NUTS AND VOLTS

Fig. 46-2

LATCHING SWITCH CIRCUIT

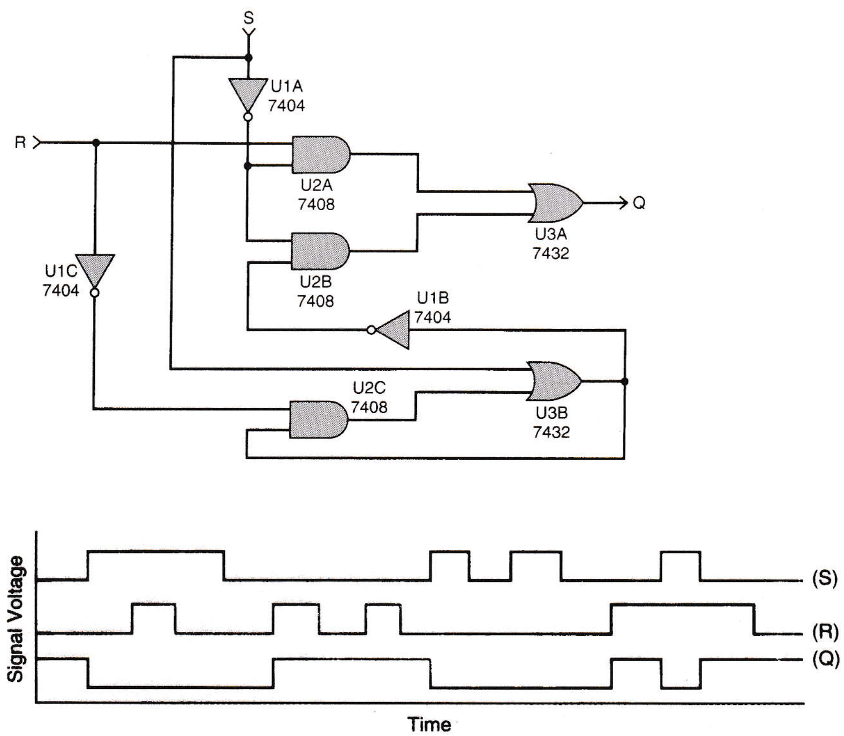


One button turns the LED on, the other turns it off. If you apply a positive pulse to the SET input, the output turns on. It remains on until a positive pulse is applied to the RESET input to turn it off. (TTL flip-flops such as the 74LS74 are actuated by negative, rather than positive pulses.)

ELECTRONICS NOW

Fig. 46-3

DIGITAL LATCH WITH SAFETY RESET FEATURE

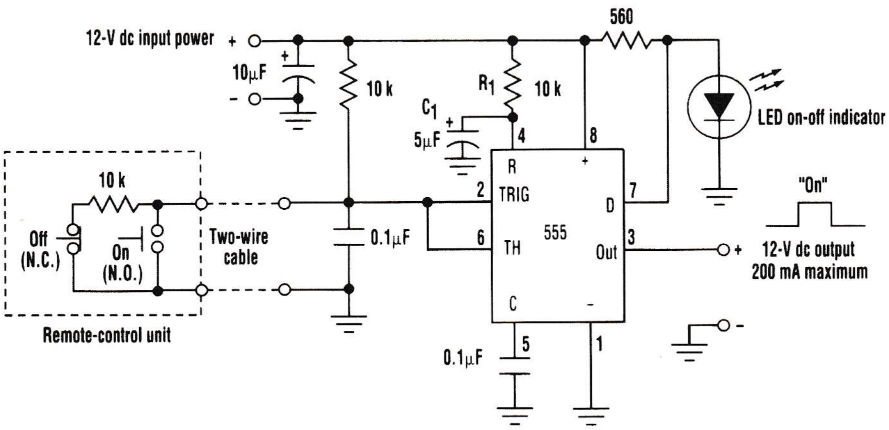


NASA TECH BRIEFS ***Fig. 46-4***

The time diagram illustrates the various modes of operation of the circuit. A high S input causes the output (Q) to go low. Thereafter, a high R input can reset Q to high, but only so long as S remains low.

The asynchronous digital latching circuit is designed for use in a safety-related application, like turning off power in response to an alarm signal. During normal operation in the absence of an alarm, the SET (S) and RESET input voltages are low or off, while the output voltage (Q) is high or on. The SET input constitutes the alarm signal: Whenever "S" goes high (on), Q goes low (off), and thereafter remains low, even when S goes low. Thus, for example, the circuit keeps a power supply turned off even when the alarm has been shut off. If a safe condition has been restored, then the circuit can be reset to Q high by applying a high (on) signal to the RESET (R) input terminal. However, regardless of the R input level, Q cannot be driven high as long as S remains high; that is, the circuit cannot be reset if the alarm signal is still on. Thus, the RESET signal cannot override the alarm signal and thereby provide a false indication of safety. Also, this does not go into oscillation when the SET and RESET inputs change simultaneously.

555 LATCH

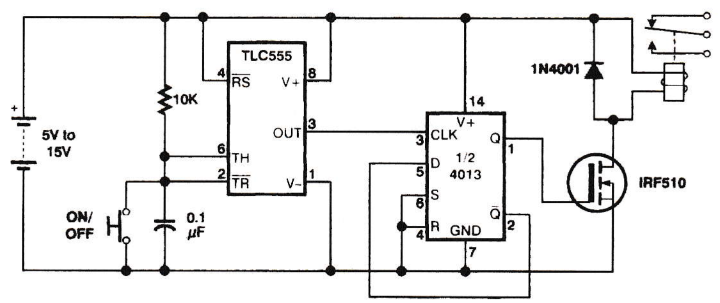


ELECTRONIC DESIGN

Fig. 46-5

Using a 555 chip in the memory mode, this push-button-controlled latch switch can source up to 200 mA of load current. Only one pair of wires is required to interface the ON and OFF push buttons to the control circuitry. The memory-mode feature of the 555 chip is implemented by connecting the trigger (pin 2) and threshold (pin 6) inputs together and applying one-half the supply voltage via a resistor network. Momentarily forcing the input low causes the output to go high, while forcing the input high causes the output to go low. To facilitate remote operation of the latch switch using one pair of wires, one resistor in the voltage-divider network is installed in the remote-control unit. Shorting this resistor out with the ON push button causes the output to go high. Conversely, opening this resistor with the OFF push button induces the output to go low. The R1-C1 network connected to the RESET input (pin 4) forces the latch to come up in the OFF state when power is first applied. The LED on-off indicator is kept off whenever the discharge output (pin 7) is conducting. When the output (pin 3) goes high.

ONE-BUTTON LATCHING SWITCH



ELECTRONICS NOW

Fig. 46-6

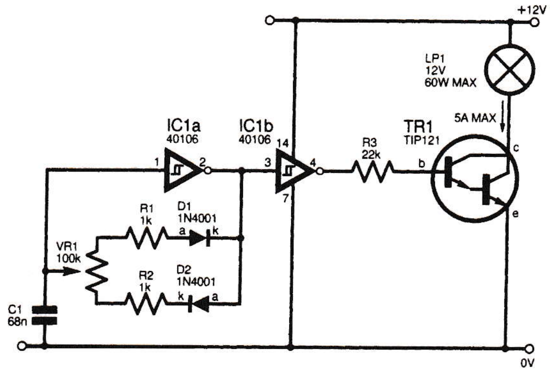
The SET and RESET inputs are grounded, the inverted (-Q) output is fed back to the D input, and the pulses go into the CLOCK input. Each positive pulse makes the flip-flop toggle from one state to the other. The TLC555 chip in the figure serves two purposes. It inverts the pulses so that you can get a positive pulse from a switch that is connected to ground. More importantly, it also "debounces" the switch. When you press a button, it doesn't just make contact once—the contacts "bounce," opening and closing three or four times. The 4013 would toggle once on each bounce, leading to unpredictable results. The TLC555 uses a resistor and capacitor to smooth out these fluctuations so that each press of the button produces only one pulse.

47

Light-Control Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

DC LAMP DIMMER

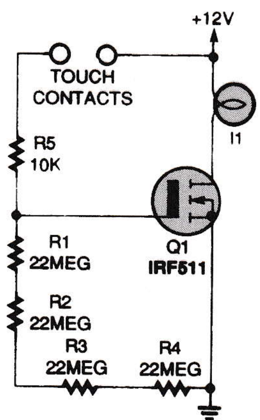


EVERYDAY PRACTICAL ELECTRONICS

Fig. 47-1

The dc lamp dimmer circuit will adjust the brightness of a dc lamp or the speed of a suitable dc motor. An oscillator is formed around Schmitt inverter IC1a, and is buffered by IC1b. This, in turn, drives TR1, which is a Darlington power transistor. The lamp LP1 forms the collector (c) load for the transistor, and this is driven at a nominal 330 Hz. The duty cycle is, however, fully variable using potentiometer VR1. Consequently, the power delivered to the load can be controlled between 5 and 95 percent, approximately. No flicker will be apparent at this frequency. The circuit consumes little power itself and is ideal for controlling car instrument panel lights.

INCANDESCENT LAMP TOUCH CONTROL

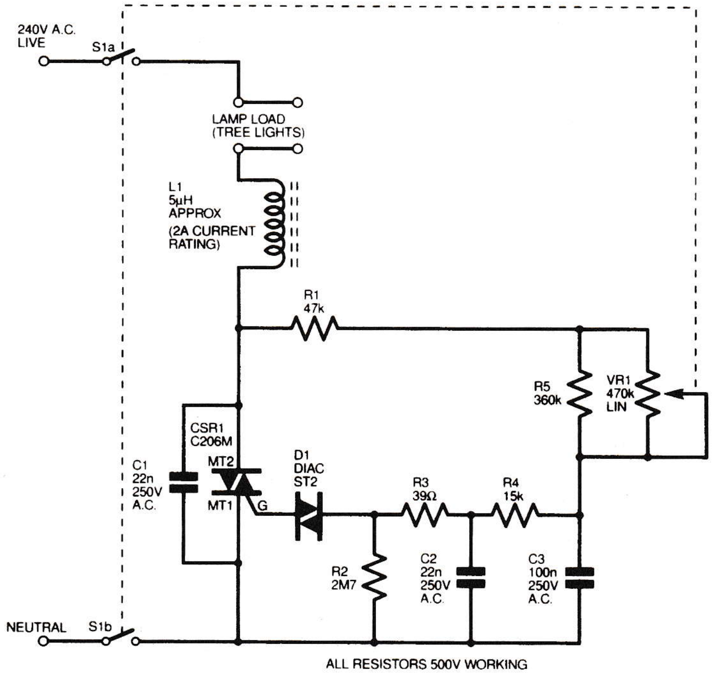


An IRF511 power MOSFET controls an incandescent lamp. This circuit is useful where low-voltage dc is used.

POPULAR ELECTRONICS

Fig. 47-2

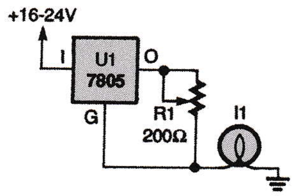
CHRISTMAS LIGHT DIMMER



EVERYDAY PRACTICAL ELECTRONICS ***Fig. 47-3***

This is based on a standard diac/triac dimmer configuration, which has been optimized for this application. Resistor R2 is, however, optional and was included only to help discharge the capacitors quickly. All capacitors are 260-Vac X2 rated (*this is very important*), all resistors are 500-V working voltage, and triac CSR1 was chosen for its low holding current so that it would function with small loads. Note that VR1 is fitted with a double-pole mains switch for complete isolation. Even with only one 20-lamp string plugged in (22 W), control is very smooth with barely any trace of hysteresis. This circuit can be built safely on a piece of matrix board, wired point-to-point, provided it is mounted in a *completely insulated plastic case* and that no external components (i.e., S1 and VR1) are metal. (*Do not build this circuit if you do not understand the safety requirements.*)

INCANDESCENT LAMP-LIFE EXTENDER

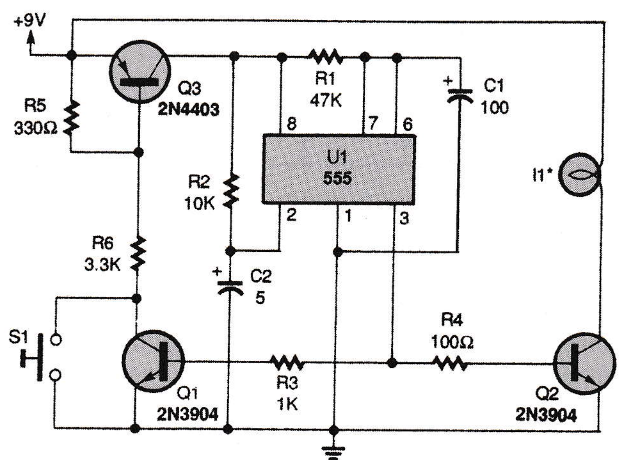


POPULAR ELECTRONICS

Fig. 47-4

The cold resistance of an incandescent lamp is normally very low compared to its operating resistance. Each time such a lamp is turned on, the initial current is several times greater than its rated operating current. The life-extender circuit will work with any incandescent lamp that operates at a voltage of 1.5 to 12 V and a current of 1 A or less. Look up the lamp's normal operating current and, using an ammeter, set R1 so that the normal operating current flows to the lamp. Now, each time the lamp is switched on, the initial current will be limited to its preset value.

TIMED NIGHT LIGHT

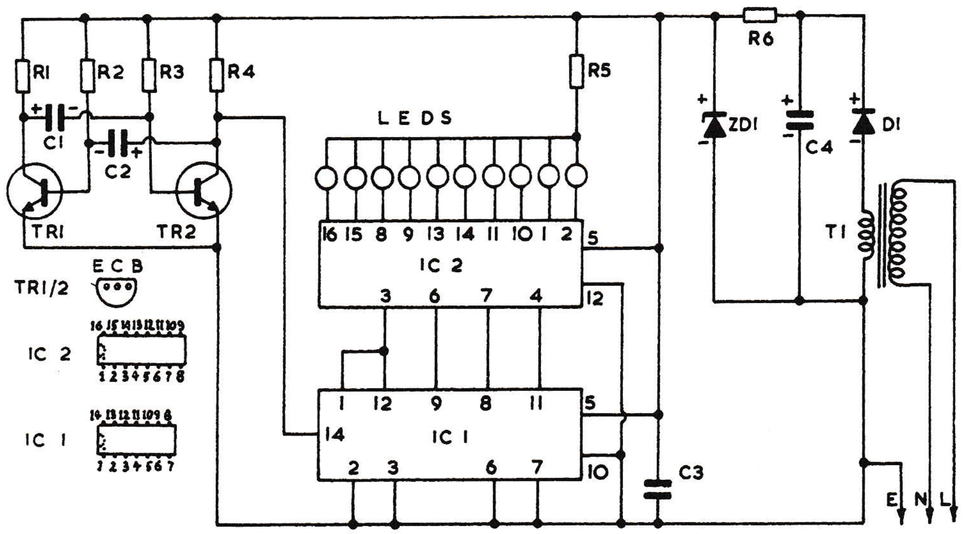


POPULAR ELECTRONICS

Fig. 47-5

A small lamp is turned on via switch Q2 for a predetermined time. S1 initiates the 555 timer cycle, holding on Q1 and switch Q3, supplying power to the circuit, and Q2, turning on the lamp. At the end of the cycle, power is removed from the circuit because Q3 is cut off. Therefore, no current is drawn from the battery during standby.

TWINKLE TREE



ELECTRONIC EXPERIMENTERS HANDBOOK

Fig. 47-6

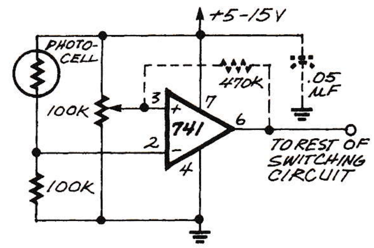
Twinkle Tree is an easy project for beginners to build, and its basic circuit has a number of useful applications. The circuit's visible action appears as a string of 10 LEDs (light-emitting diodes) flashing on, one at a time, in sequence, this being repeated so long as the circuit is powered. The LEDs can be used on a small table decoration, in the form of a Christmas tree, or around a picture frame; or they can be placed at various points on a hanging decoration, such as a bunch of mistletoe, or can be incorporated in other decorations or modes. The LED light display provides an interesting and novel twinkling effect.

48

Light-Controlled Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

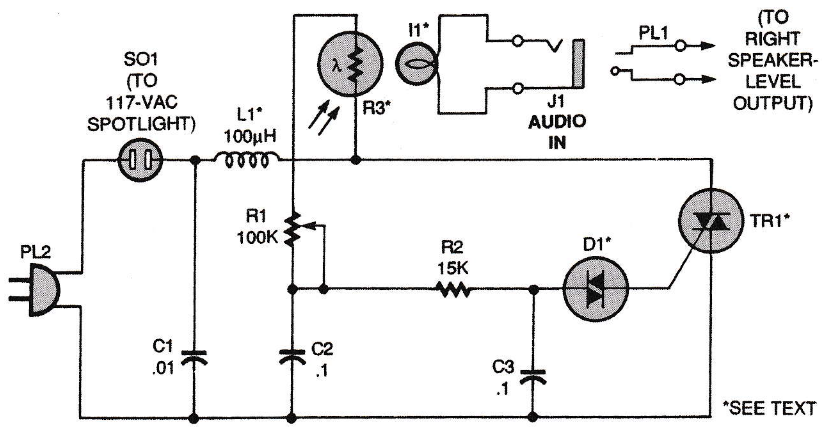
HYSTERESIS-STABILIZING LIGHT SENSOR



ELECTRONICS NOW *Fig. 48-1*

This light turn-on circuit originally had oscillation problems. The 470-kΩ resistor introduces the hysteresis to prevent the oscillation. You can experiment with the value of the resistor for best results in your application.

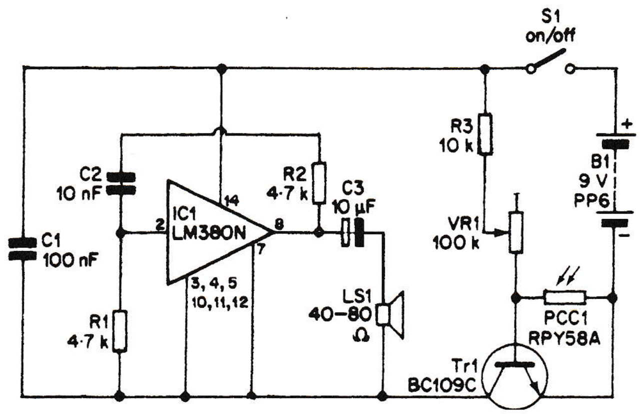
AUDIO-CONTROLLED LAMP DIMMER



POPULAR ELECTRONICS *Fig. 48-2*

The audio input to J1 lights low-voltage lamp I1. This light illuminates light-dependent resistor R3 (100 kΩ to 1 MΩ dark resistance) and modulates the intensity of the lamp plugged into SO1. L1 and C1 suppress RFI caused by the triac phase-control circuit. D1 and TR1 are a diac and a triac, respectively, and are the types used in common lamp dimmers.

DARK ALARM

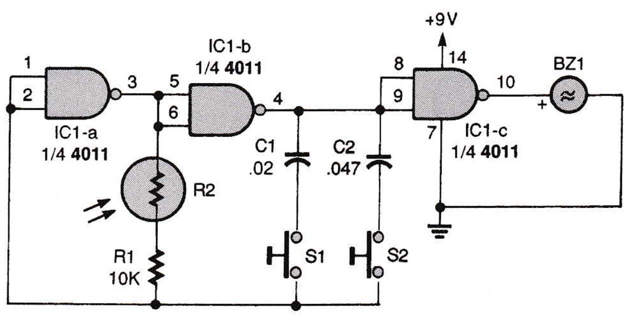


ELECTRONIC EXPERIMENTERS HANDBOOK

Fig. 48-3

This alarm senses darkness. Photocell PCC1 is normally irradiated with light and has a low resistance. In darkness, the resistance increases, and enough bias is available to turn on Tr1, activating the oscillator circuit consisting of IC1 and associated components. This produces a tone in speaker LS1.

MAGIC WAND

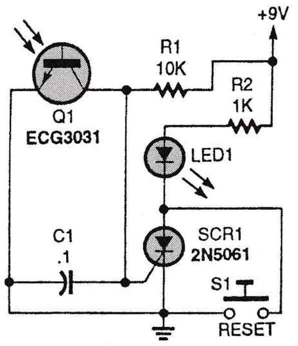


POPULAR ELECTRONICS

Fig. 48-4

This device is easy to use: Just aim the wand at a light source and listen to the tone. Two gates, IC1-a and IC1-b, of a quad two-input NAND gate are connected in an audio-oscillator circuit with R1, the photocell (R2), C1, and C2 setting the tone of BZ1. When the photocell is aimed at a light source, its resistance drops and the oscillator's output tone increases in frequency. When either of tone switches (S1 or S2) is pressed, the frequency range goes up. Pressing S2 shifts the oscillator into its highest-frequency range. Only one tone switch can be pressed at a time; if both switches are pressed simultaneously, the output stops.

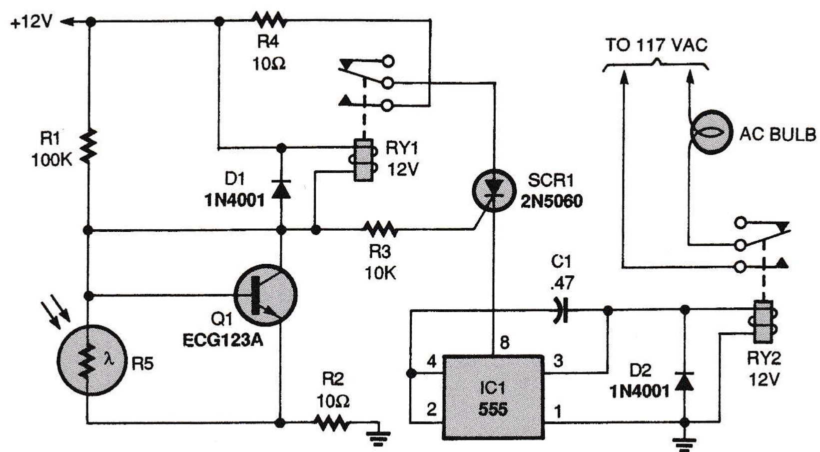
LOSS-OF-LIGHT DETECTOR



Light falling on Q1 causes Q1 to conduct. When the light fails, C1 charges toward +9 V, triggering SCR1, lighting LED1. S1 resets the SCR.

POPULAR ELECTRONICS *Fig. 48-5*

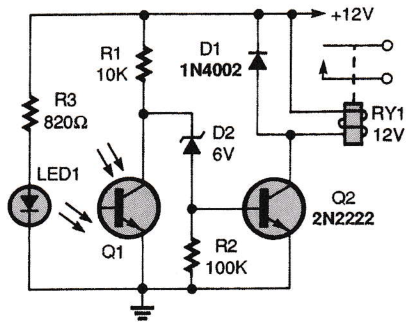
LIGHT-ACTIVATED SWITCH



POPULAR ELECTRONICS *Fig. 48-6*

When even a little light hits light-dependent resistor R5, transistor Q1 is turned off because the base has less resistance to the ground than to the positive rail. In that situation, the base is at a negative potential. When the sun sets, R5 is no longer illuminated, giving the transistor's base a high resistance to ground—higher than 100,000 Ω. With less resistance to positive potential, the base is biased, turning on Q1. Relay RY1 is then energized and pulls in, connecting the SCR1's anode to positive potential. The 555 timer, IC1, powers up, and its output goes high to approximately 10.67 V, which is sufficient to energize RY2. Relay RY2 then pulls in, keeping the ac bulb on the whole night, and turning it off again at sunrise.

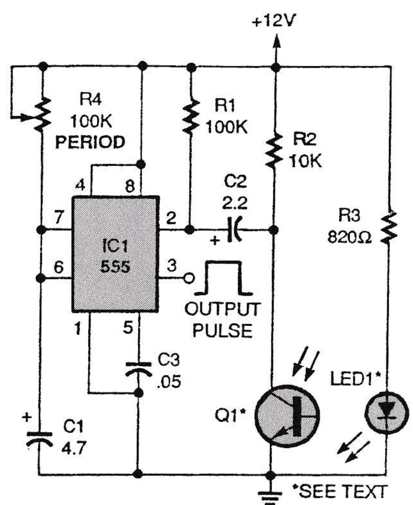
LIGHT BLOCK SENSOR



POPULAR ELECTRONICS ***Fig. 48-7***

In this circuit, relay RY1 remains open until the light source is blocked. As long as the IR light source is uninterrupted, phototransistor Q1's collector voltage is near zero. The voltage at zener diode D2's cathode is too low for conduction, keeping transistor Q2 off and the RY1 open.

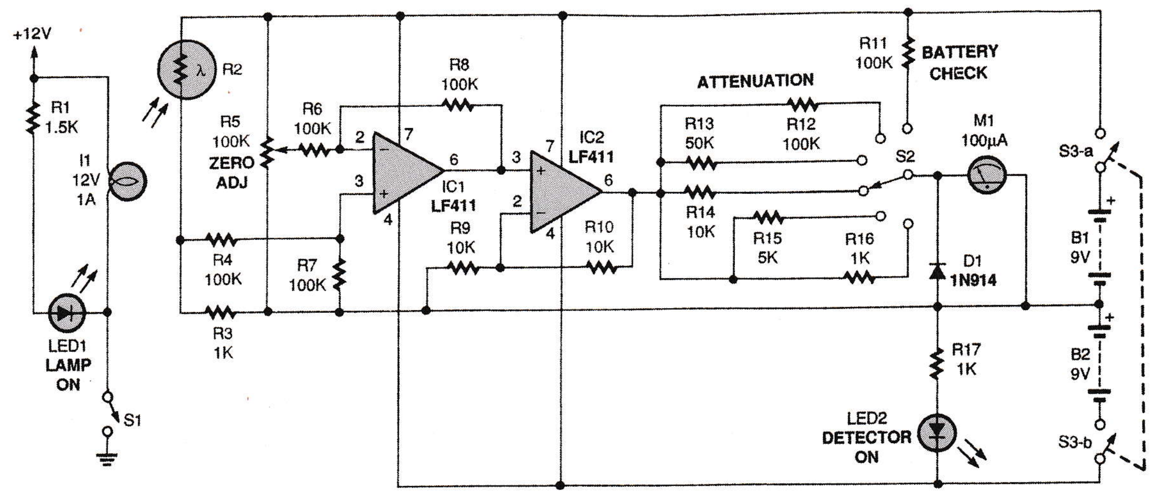
PROGRAMMABLE CONTROLLER LIGHT SENSOR INTERFACE



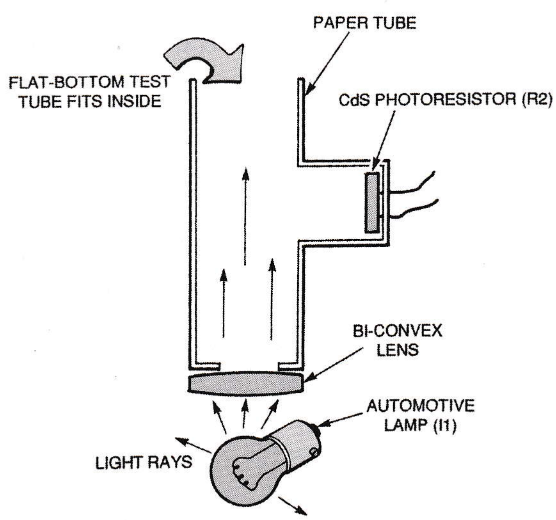
POPULAR ELECTRONICS ***Fig. 48-8***

As long as an object remains between LED1 and Q1, no output occurs. But as soon as the object moves out from between the LED and the phototransistor, IC1 (a 555 timer) is triggered, producing a timed output pulse. This circuit can be used to indicate when a part has moved away from a location.

ELECTRONIC TURBIDIMETER



A

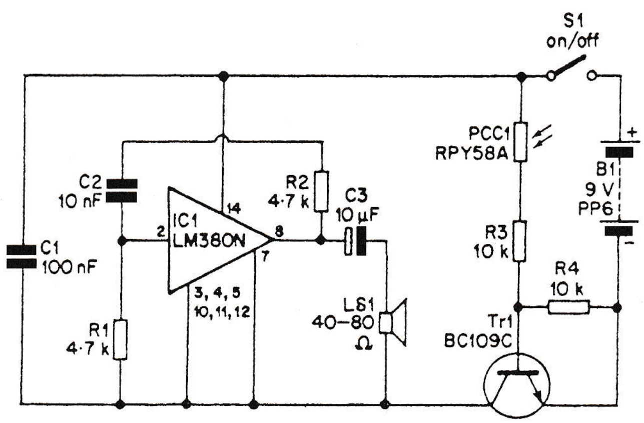


B

POPULAR ELECTRONICS ***Fig. 48-9***

A turbidimeter is a scientific instrument used to measure the cloudiness of solutions, such as water. If you measure the amount of scattered light, you can measure the turbidity. Paint the inside of the test-tube holder flat black. Glue a cadmium sulfide photoresistor (R2) to a cardboard disk, and glue the assembly to the open end of the 90° tube. Next, cement a biconvex lens (one that curves outward on both sides) to the bottom of the test-tube holder. Power up I1, a 12-V, 1-A automotive light bulb, and hold it at various distances below the biconvex lens. Find the distance at which a beam of light comes straight up the tube or converges slightly, but does not diverge. Mount the lamp and play it on the 100-μA meter; the resistors on the rotary switch set the sensitivity or attenuation. Power is supplied to the circuit by two 9-V batteries in a split supply; I run the lamp off a separate 12-V supply. If you want truly calibrated readings, you'll need to buy a turbidity standard solution, available in JTU (Jackson Turbidity Units) and NTU (Nephelos Turbidity Units), from a science-supply company. In operation, zero the meter to a clean-water blank, note the reading from a known standard, note the reading from the unknown sample, and calculate the sample's turbidity, based on the needle's position.

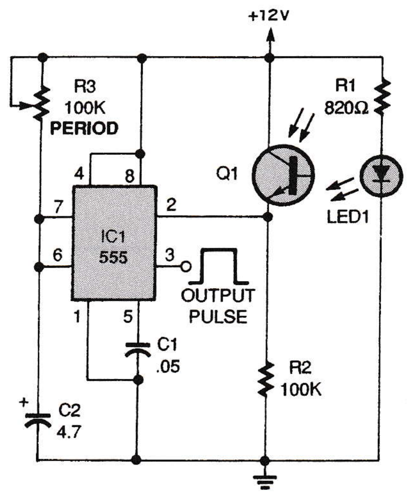
LIGHT ALARM



ELECTRONIC EXPERIMENTER'S HANDBOOK ***Fig. 48-10***

The circuit is activated if Tr1 is switched on by a suitable base current and voltage. The voltage and current available at the base of Tr1 are dependent on two main factors: the resistance provided by R4 and the setting of control VR1. If VR1 is set at maximum value, photocell PCC1 needs to have a resistance of about 10,000 Ω to bias Tr1 into conduction and activate the audio alarm circuit, of which IC1 is a primary part. Fixed resistor R4 has been used across the base-emitter terminals of the switching transistor so that the sensitivity of the circuit is preset. R4 can be raised somewhat in value if increased sensitivity is required. The audio-alarm generator uses an LM380N (IC1) in a simple audio-oscillator circuit, and drives high-impedance loudspeaker LS1 via coupling capacitor C3. Provided the losses through this coupling are less than the voltage gain provided by the amplifier, this will give sufficient positive feedback to sustain oscillation. The values for R1, R2, and C2 shown in the circuit diagram give considerably more feedback than is needed to just sustain oscillations, and the circuit oscillates strongly, producing a square-wave output at a frequency in the region of 1 kHz (1000 Hz).

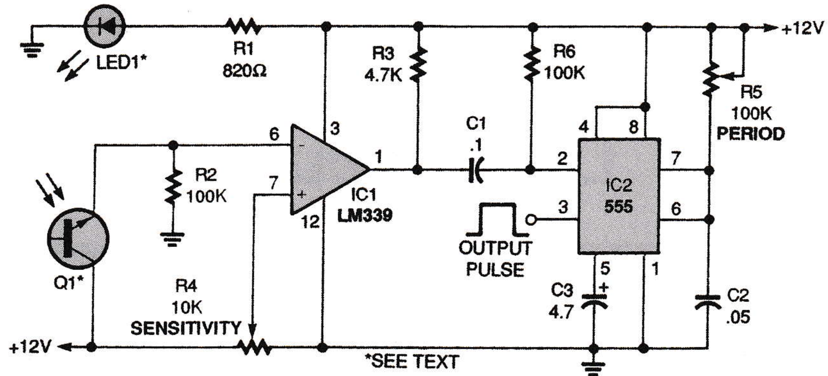
PROGRAMMABLE CONTROLLER PHOTOELECTRIC INTERFACE



POPULAR ELECTRONICS *Fig. 48-11*

This circuit is an IR sensor with a timed positive output pulse that will operate with older and slower PLCs. Some of the early PLCs have scan times of 15 ms or more. The sensor's extended output pulse can be set, by R3, to a time period longer than the controller's scan time. The sensor can also be used as a stand-alone circuit to operate a counter, a valve, an indicator, or any other electrically controlled device. As long as nothing is blocking the IR light source, the emitter of phototransistor Q1 is high and 555 timer IC1 is set in the READY state. When an object blocks the light source, Q1 turns off, sending a negative pulse to the trigger input at pin 2 of IC1, producing a timed output pulse. R_3 and C_2 determine the length of the output pulse. Larger values produce longer output pulses.

REFLECTIVE SENSOR



POPULAR ELECTRONICS *Fig. 48-12*

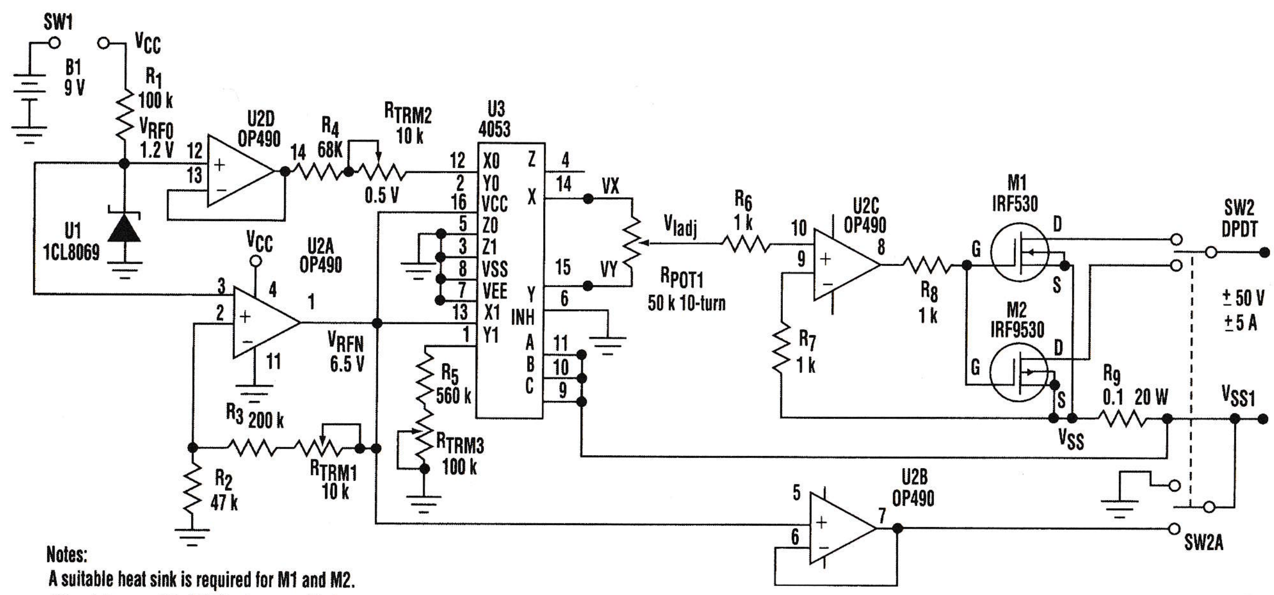
The IR sensor circuit shown produces an output when LED1's light is reflected from an object back to phototransistor Q1. The LED and phototransistor should be mounted parallel to each other and aimed in the same direction. Without a reflective object, the voltage at the input of the LM339 comparator (IC1) is at or near ground level and the output at pin 1 of IC1 is high. When the phototransistor detects a reflected light signal, the voltage at Q1's emitter goes high, causing the comparator's output to go low. The 555 timer (IC2) is then triggered, and produces a timed output pulse at pin 3. The circuit's sensitivity is set by R4 and its output time period by R5. Note that in this circuit, all of the unused input and output pins of the LM339 must be tied to circuit ground.

49

Load Circuits

The sources of the following circuits are contained in the Sources section, which begins on page 1043. The figure number in the box of each circuit correlates to the entry in the Sources section.

BIDIRECTIONAL ACTIVE LOAD



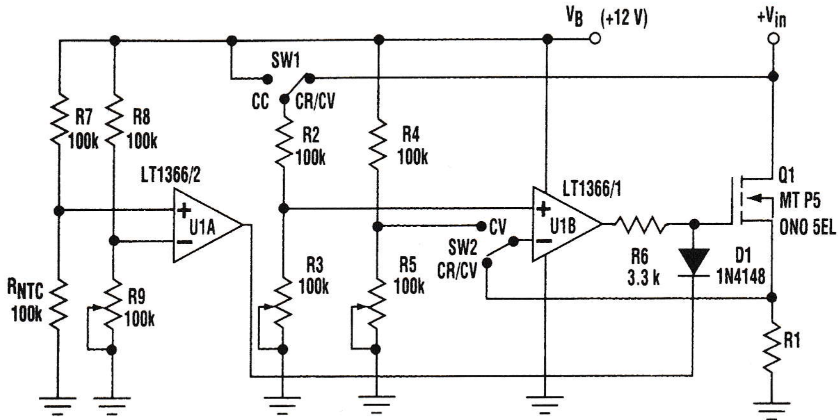
Notes:
A suitable heat sink is required for M1 and M2.
All resistors are 5%, 1/4 W unless specified.

ELECTRONIC DESIGN

Fig. 49-1

The design presented here is a single +9-V, battery-operated, bidirectional active load that can sink and source current. This is a low-power design, consuming only about 140 μA. The power MOSFET selected (IRF530 N-channel and IRF9530 P-channel), after derating, accepts a maximum load of ±50 V at ±5 A. The key to the design is having two different sets of voltage levels at V_X, V_Y, and V_{SS1}. One set is for current-sinking test, while the other is for current-sourcing test.

ACTIVE LOAD RESISTOR



ELECTRONIC DESIGN *Fig. 49-2*

The design idea presented here concerns an active power resistor. It can be used as a load resistor when probing or servicing power supplies. The circuit can work in three different modes. It can act as a constant resistor (mode CR), or as a constant current from a supply of any positive voltage. Finally, it can be in constant-voltage mode (CV), in which the circuit loads the voltage across supply terminals to a constant value adjusted by the user. The power MOSFET transistor Q1 works as a resistive component. The transistor gate is controlled by an op amp (U1B). The feedback voltage, which can be selected by switch SW2, is connected to the amplifier's inverting input. In CC and CR modes, the feedback voltage is the voltage between the source resistor (R1) terminals, which is proportional to the amplifier supply voltage (V_B) through a voltage-divider circuit (R4-R5). The amplifier noninverting input is controlled by a control voltage. The control-voltage input can be selected by switch SW1. In CR and CV modes, that input is the resistor input voltage (V_{in}), and in CC mode, it is the amplifier supply voltage (V_B). The control voltage is set in voltage divider R2-R3 to a proper value to control the amplifier. The other op amp (U1A) protects the MOSFET transistor. It is controlled by a resistor bridge circuit consisting of three resistors (R7 through R9) and an NTC resistor. The NTC is in contact with the transistor cooling element. With moderate element temperatures, the output voltage of U1A is high, and thus has no effect on the transistor gate because of the reverse-biased diode (D1). If the element temperature becomes high, the amplifier output has a zero value, which takes the transistor gate voltage to zero through the diode. The amplifiers also can be powered directly from the resistor input voltage. This circuit can work in CR mode without any control voltage. In CC and CV modes, however, it should be controlled by the external voltage CC/CV input. The circuit acts in CR mode as a resistor that has a resistance of

$$R=[(R_2+R_3)/R_3]\times R_1$$

The required resistance value can be adjusted by potentiometer R3. In CC mode, the circuit sinks a current

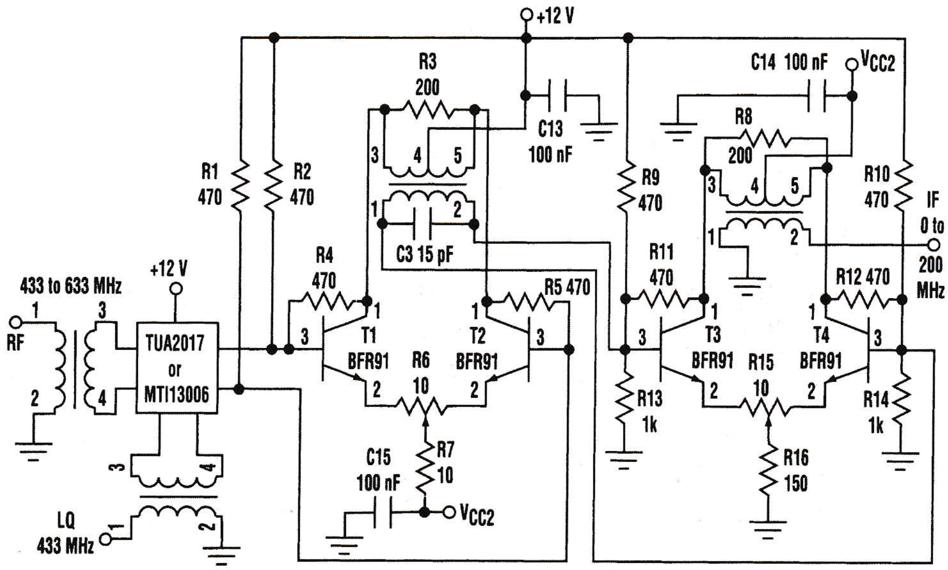
$$I=R_3/[(R_2+R_3)R_1]\times V_B$$

The sink current also can be adjusted by potentiometer R3. In CV mode, the voltage between the terminals of the resistor is:

$$V=[(R_2+R_3)/R_3][R_5/(R_4+R_5)]\times V_B$$

The required constant voltage can be adjusted by both R3 and R5.

MIXER LOAD DIFFERENTIAL AMPLIFIER



ELECTRONIC DESIGN

Fig. 49-3

Two different amplifier stages are used as the load of the mixer TUA2017 or MTI13006 (products developed by Siemens Co.). The LQ frequency is fixed at 433 MHz, because a commercial SAW oscillator at 433 MHz can be used as a local oscillator. If the RF frequency is swept from 433 to 633 MHz, then the IF frequency also is swept from 0 to 200 MHz. Compared to a passively loaded mixer, this combination of mixer and differential amplifiers exhibits excellent performance. The relative amplitude (amplitude difference between the fundamental frequency and the largest distortion frequency) is typically at least 50 dB. But it is only 30 dB (or less) in the case of a passive load mixer. Resistors R4, R5, R11, and R12 are used as the shunt feedback to obtain the flat output amplitude when the RF frequency is swept. The amplitude ripple with the IF bandwidth (0 to 200 MHz) is less than 0.5 dB.