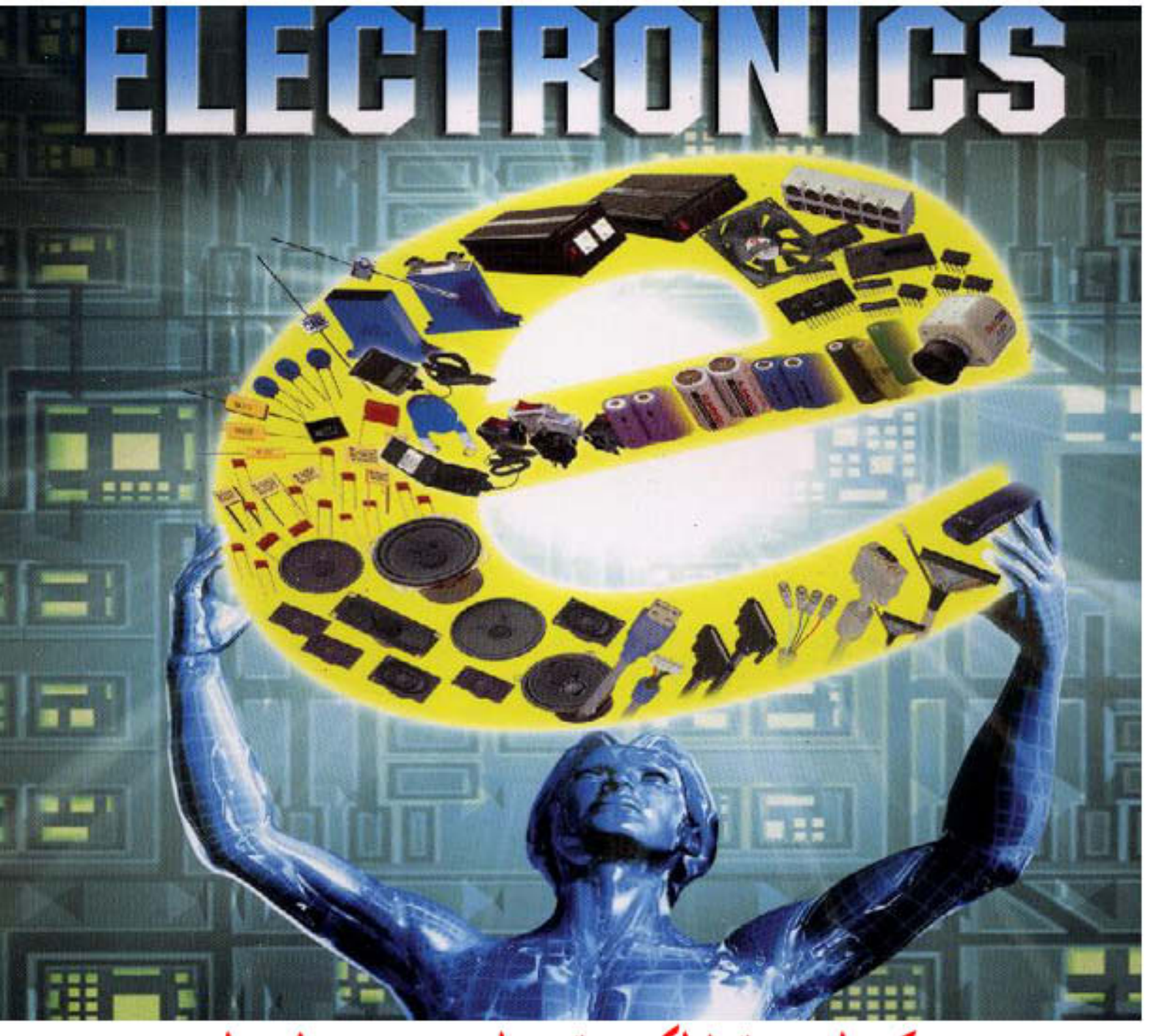
3138

Encyclopedia of ELECTRONIC CIRCUITS

Volume 2

Rudolf F. Graf



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Introduction

Encyclopedia of Electronic Circuits—Volume 2, a companion to Volume 1 published in 1985, contains well over 1400 not-previously covered circuits organized into 108 chapters. For each reference, circuits are listed at the beginning of each chapter. The extensive index further enhances the usefulness of this new work. The browser, as well as the serious researcher looking for a very specific circuit, will be richly rewarded by the context of this volume. A brief explanatory text accompanies almost every entry. The original source for each item is also given so that the reader requiring additional data will know where to find it.

I am most grateful to William Sheets for his many and varied contributions to this book, and to Mrs. Stella Dillon for her fine work at the word processor. These friends and associates of long standing have my sincere thanks for contributing to the successful completion of this book.

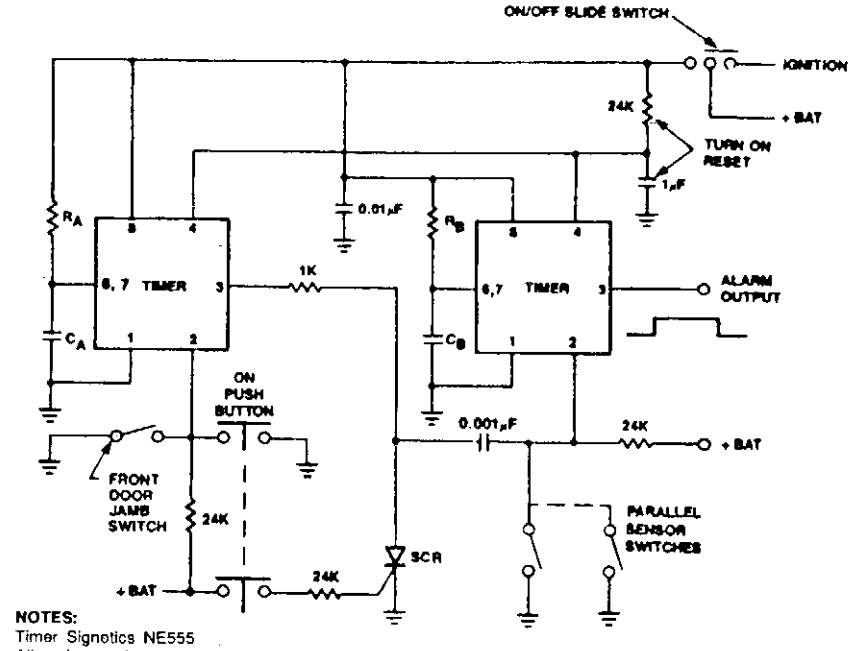
1

Alarm and Security Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Auto Burglar Alarm
Multiple Alarm Circuit
Differential-Voltage or Current Alarm
Trouble Tone Alert
Photoelectric Alarm System
Alarm Circuit

AUTO BURGLAR ALARM



NOTES:
Timer Signetics NE555
All resistor values in ohms

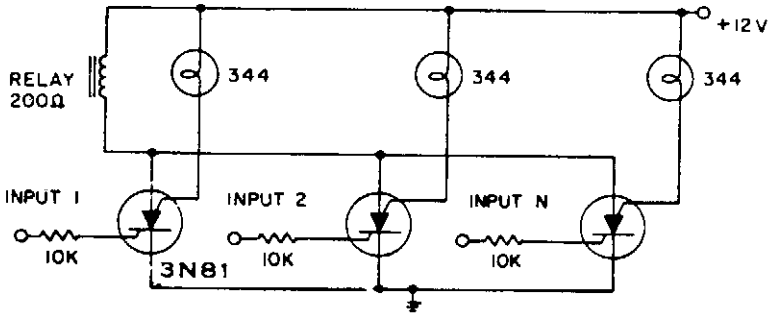
SIGNETICS

Fig. 1-1

Circuit Notes

Timer A produces a safeguard delay, allowing the driver to disarm the alarm and eliminating a vulnerable outside control switch. The SCR prevents timer A from triggering timer B, unless timer B is triggered by strategically-located sensor switches.

MULTIPLE ALARM CIRCUIT



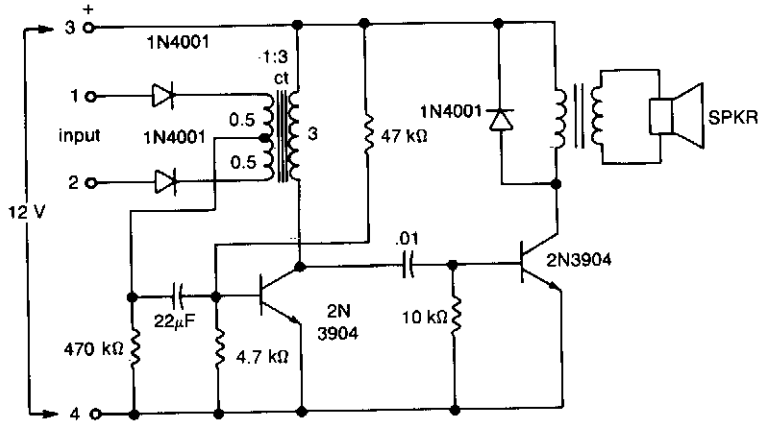
GE

Fig. 1-2

Circuit Notes

Any of several inputs pulls in the common alarm relay with lamps giving visual indication of triggering input. Low resistance lamps decrease input sensitivity.

DIFFERENTIAL VOLTAGE OR CURRENT ALARM



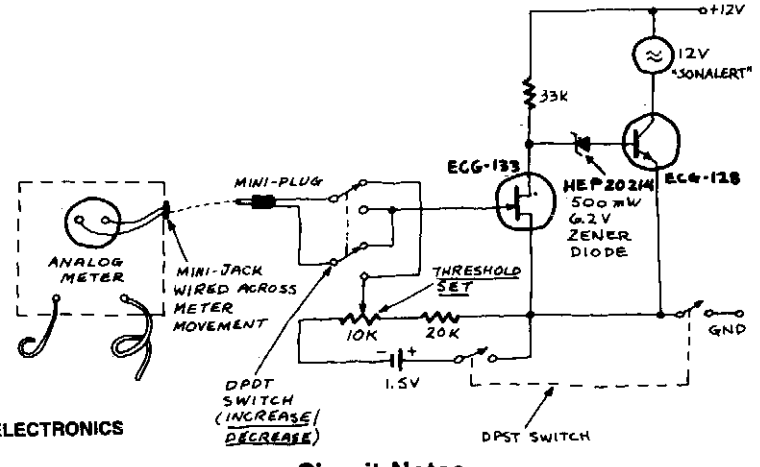
WILLIAM SHEETS

Fig. 1-3

Circuit Notes

The input may be dc or low frequency ac. The output is a distinctive series of audio beeps or a continuous tone, and occurs only when a selected polarity unbalance is present at the input.

TROUBLE TONE ALERT



RADIO ELECTRONICS

Fig. 1-4

Circuit Notes

The Trouble Tone Alert is intended for use with analog meters—just wire a "mini" earphone jack directly across the meter movement, plug it in, and you're all set. This device reacts the to the meter-movement driving voltage. It will respond to a change in ac or dc voltage, current, or in resistance. The circuit will respond to an increase or decrease selected by the DPDT switch and is adjusted with the threshold control until the tone from the Sonalert just disappears (with the meter in the circuit being tested, of course).

PHOTOELECTRIC ALARM SYSTEM

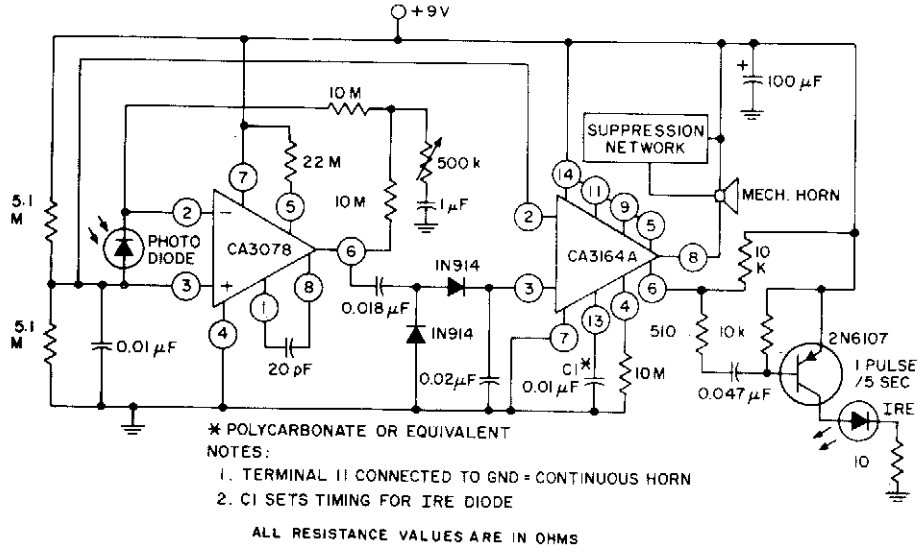


Fig. 1-5

GE/RCA

Circuit Notes

The CA3164A BiMOS detector alarm system and the CA3078 micropower op amp with a photodiode are used as an automatic switch for turning on a night light or sounding a mechanical horn.

ALARM CIRCUIT

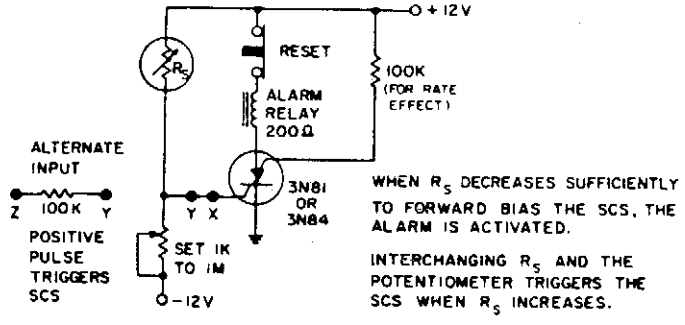


Fig. 1-6

GE

Circuit Notes

Temperature, light, or radiation sensitive resistors up to 1 megohm readily trigger the alarm when they drop below the value of the preset potentiometer. Alternately, 0.75 V at the input to the 100 kΩ triggers the alarm. Connecting SCS between ground and −12 V permits triggering on negative input to G_A.

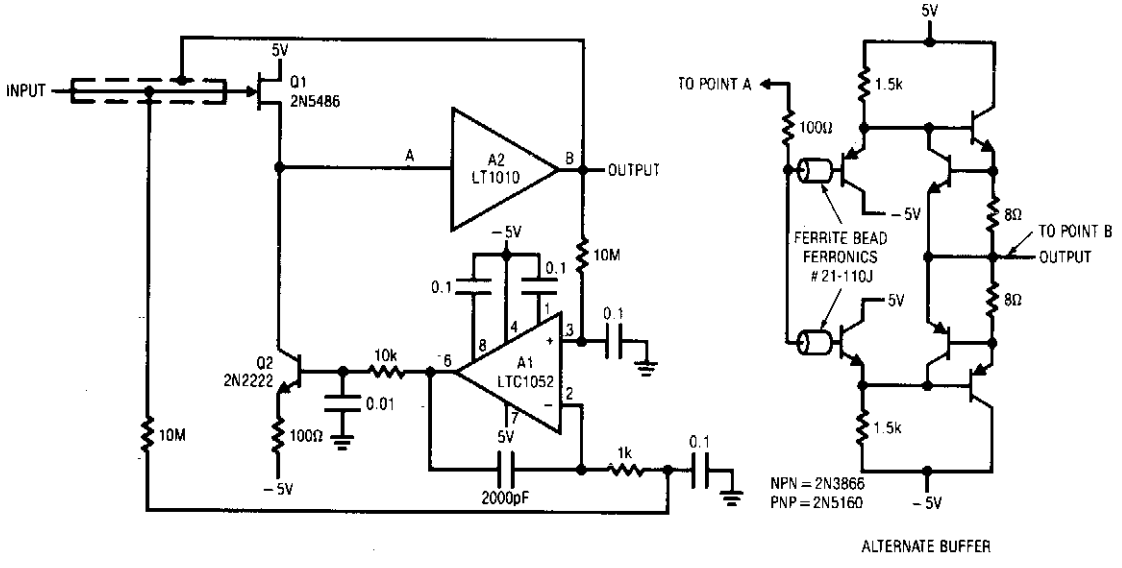
2

Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Stable Unity Gain Buffer with Good Speed and High Input Impedance
Chopper Stabilized Amplifier
Ultra-Low-Leakage Preamplifier
FET Input Amplifier
Ultra-High Z_{in} ac Unity Gain Amplifier
Logarithmic Amplifier
Composite Amplifier
Stereo Amplifier with Gain Control
Precision-Weighted Resistor Programmable-Gain Amplifier
Power GaAsFET Amplifier with Single Supply
Linear Amplifiers from CMOS Inverters
Current-Collector Head-Amplifier
Hi-Fi Compander
Two-Wire to Four-Wire Audio Converter
Thermocouple Amplifier
Low-Distortion Audio Limiter
Speech Compressor
Speaker Overload Protector
Audio Automatic Gain Control
Voltage Controlled Attenuator
High-Input-Impedance Differential Amplifier
Audio Q-Multiplier
Automatic Level Control
Pulse-Width Proportional-Controller Circuit
Op Amp Clamping

STABLE UNITY GAIN BUFFER WITH GOOD SPEED AND HIGH INPUT IMPEDANCE



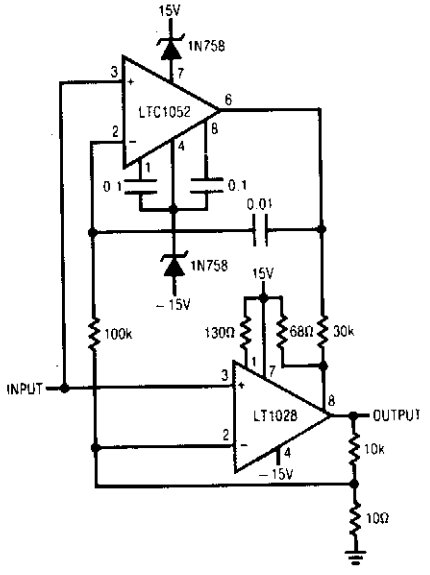
LINEAR TECHNOLOGY CORPORATION

Fig. 2-1

Circuit Notes

Q1 and Q2 constitute a simple, high speed FET input buffer. Q1 functions as a source follower, with the Q2 current source load setting the drain-source channel current. Normally, this open loop configuration would be quite drifty because there is no dc feedback. The LTC1052 contributes this function to stabilize the circuit by comparing the filtered circuit output to a similarly filtered version of the input signal. The amplified difference between these signals is used to set Q2's bias and hence Q1's channel current. This forces Q1's V_{GS} to whatever voltage is required to match the circuit's input and output potentials. The 2000 pF capacitor at A1 provides stable loop compensation. The RC network in A1's output prevents it from seeing high speed edges coupled through Q2's collector-base junction. A2's output is also fed back to the shield around Q1's gate lead, bootstrapping the circuit's effective input capacitance down to less than 1 pF. For very fast requirements, the alternate discrete component buffer shown will be useful. Although its output is current limited at 75 mA, the GHz range transistors employed provide exceptionally wide bandwidth, fast slewing and very little delay.

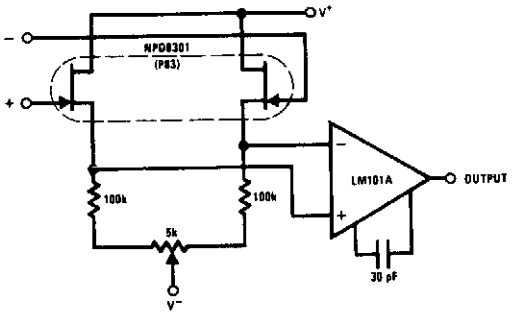
CHOPPER STABILIZED AMPLIFIER



LINEAR TECHNOLOGY CORP.

Fig. 2-2

FET INPUT AMPLIFIER



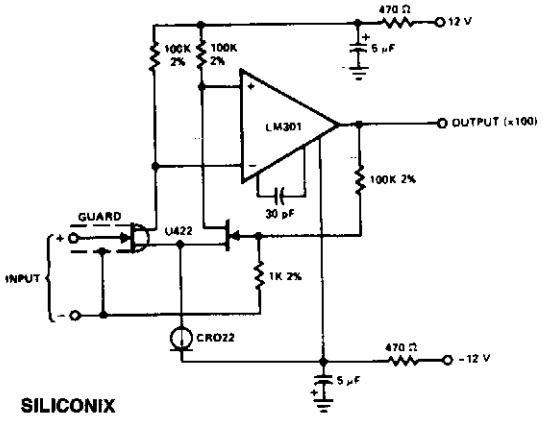
NATIONAL SEMICONDUCTOR CORP.

Circuit Notes

The NPD8301 monolithic-dual provides an ideal low offset, low drift buffer function for the LM101A op amp. The excellent matching characteristics of the NPD8301 track well over its bias current range, thus improving common-mode rejection.

Fig. 2-4

ULTRA-LOW-LEAKAGE PREAMPLIFIER



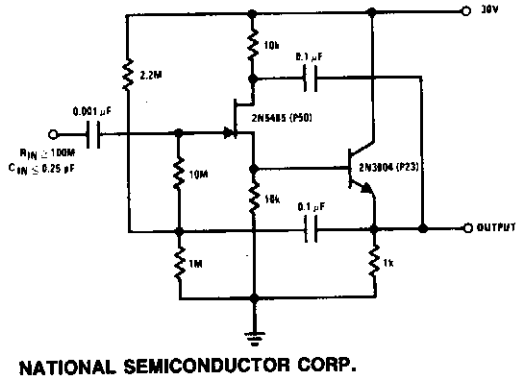
SILICONIX

Circuit Notes

The circuit has an input leakage of only 2 pA typical at 75°C and would be usable with 1 M ohm input resistance.

Fig. 2-3

ULTRA-HIGH Z_{in} AC UNITY GAIN AMPLIFIER



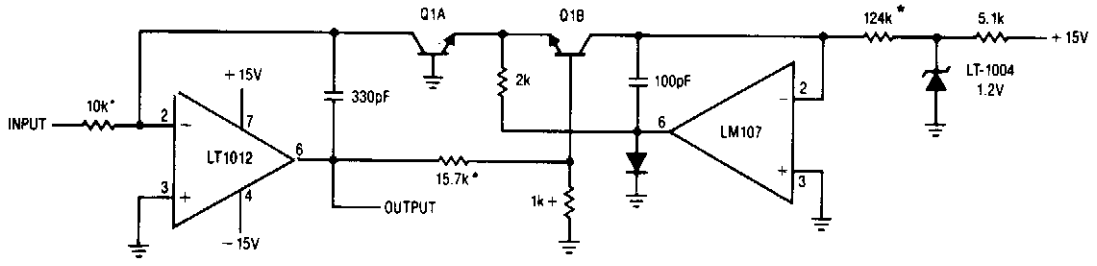
NATIONAL SEMICONDUCTOR CORP.

Circuit Notes

Nothing is left to chance in reducing input capacitance. The 2N5485, which has low capacitance in the first place, is operated as a source follower with bootstrapped gate bias resistor and drain.

Fig. 2-5

LOGARITHMIC AMPLIFIER



Low bias current and offset voltage of the LT1012 allow 4½ decades of voltage input logging.

LINEAR TECHNOLOGY CORP.

Fig. 2-6

COMPOSITE AMPLIFIER

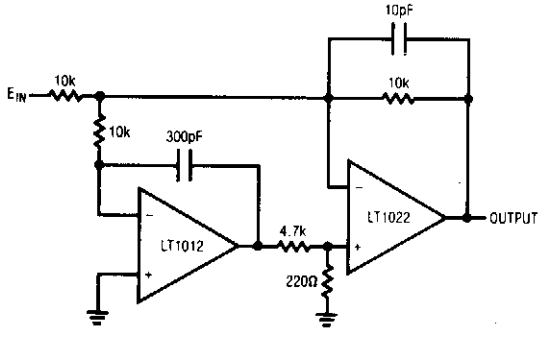


Fig. 2-7

LINEAR TECHNOLOGY CORPORATION

Circuit Notes

The circuit is made up of an LT1012 low drift device, and an LT1022 high speed amplifier. The overall circuit is a unity gain inverter, with the summing node located at the junction of three 10-k resistors. The LT1012 monitors this summing node, compares it to ground, and drives the LT1022's positive input, completing a dc stabilizing loop around the LT1022. The 10 k - 300 pF time constant at the LT1012 limits its response to low frequency signals. The LT1022 handles high frequency inputs while the LT1012 stabilizes the dc operating point. The 4.7 k - 220 ohm divider at the LT1022 prevents excessive input overdrive during start-up. This circuit combines the LT1012's 35 μV offset and 1.5 V/°C drift with the LT1022's 23 V/μs slew rate and 300 kHz full power bandwidth. Bias current, dominated by the LT1012, is about 100 pA.

STEREO AMPLIFIER WITH GAIN CONTROL

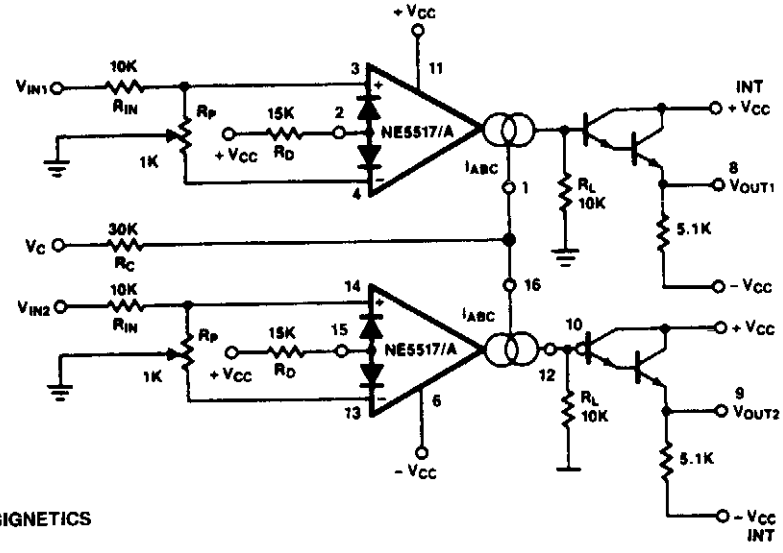


Fig. 2-8

Circuit Notes

Excellent tracking of typical 0.3 dB is easy to achieve. With the potentiometer, R_p, the offset can be adjusted. For ac-coupled amplifiers, the potentiometer may be replaced with two 5.1 k ohm resistors.

PRECISION-WEIGHTED RESISTOR PROGRAMMABLE-GAIN AMPLIFIER

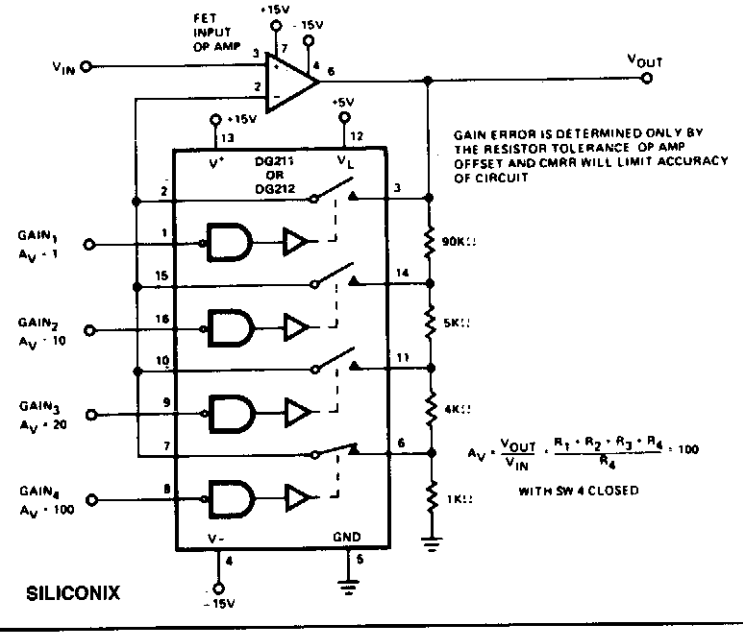
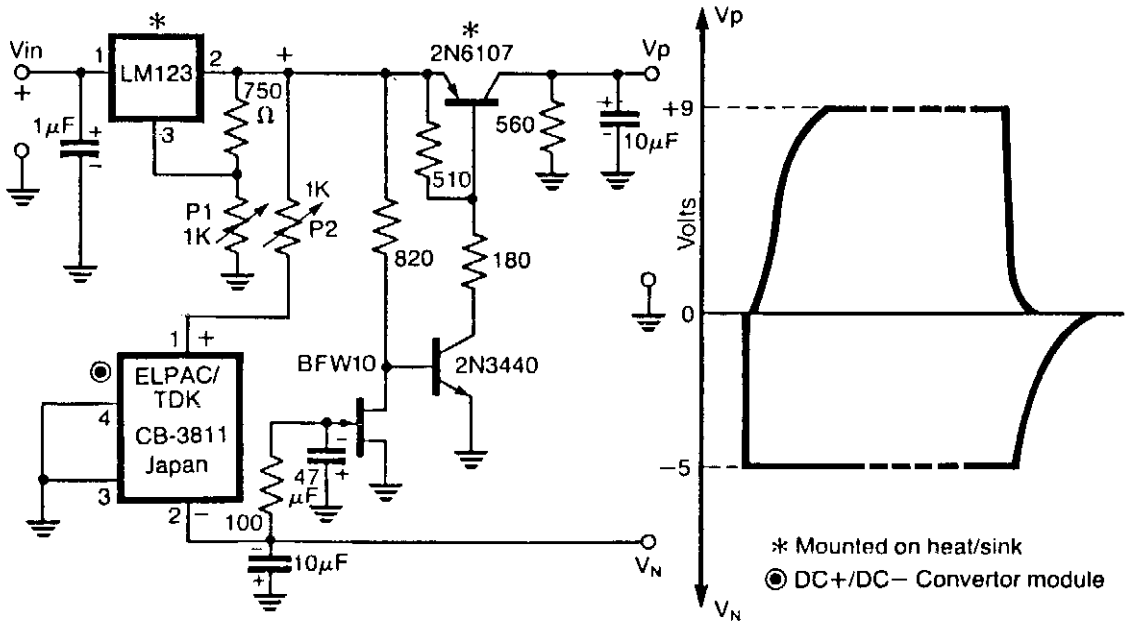


Fig. 2-9

POWER GaAsFET AMPLIFIER WITH SINGLE SUPPLY



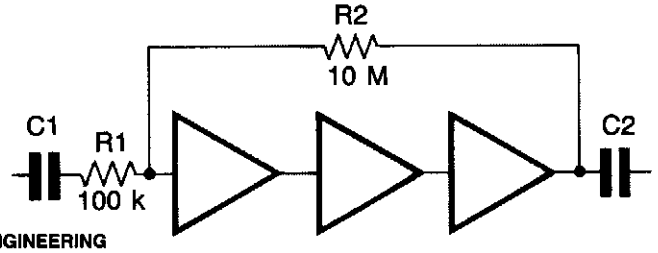
ELECTRIC ENGINEERING

Fig. 2-10

Circuit Notes

The dual regulator circuit operates from a positive supply, which when switched ON powers the gate first, and when switched OFF shuts off the drain first as shown in the figure. This circuit incorporates the LM123, a three terminal positive regulator and a dc+ to dc− converter, the outputs of which power the drains and gates of the power GaAsFETs in a power amplifier relay. The output of the three terminal regulator drives a dc+ to dc− converter whose output biases an N-channel JFET suitably so as to pull the base of the series pass transistor 2N6107 to a level to turn it on. The circuit will turn off the drain supply whenever the negative potential on the Gate fails.

LINEAR AMPLIFIERS FROM CMOS INVERTERS



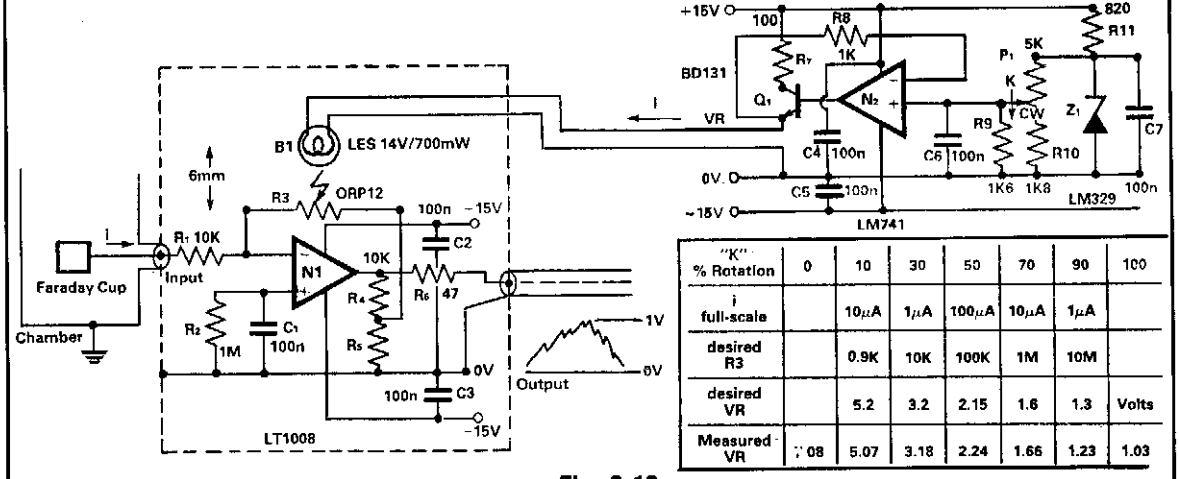
ELECTRONIC ENGINEERING

Fig. 2-11

Circuit Notes

CMOS inverters can be used as linear amplifiers if negative feedback is applied. Best linearity is obtained with feedback applied around three inverters which gives almost perfect linearity up to an output swing of 5 V p-p with a 10 V supply rail. The gain is set by the ratio of R1 and R2 and the values shown are typical for a gain of 100. The high frequency response with the values shown is almost flat to 20 kHz. The frequency response is determined by C1 and C2. This circuit is not suitable for low level signals because the signal-to-noise ratio is only approx. 50 dB with 5 V p-p output with the values shown.

CURRENT-COLLECTOR HEAD-AMPLIFIER



"K" % Rotation	0	10	30	50	70	90	100
i full-scale		10μA	1μA	100μA	10μA	1μA	
desired R3		0.9K	10K	100K	1M	10M	
desired VR		5.2	3.2	2.15	1.6	1.3	Volts
Measured VR	0.08	5.07	3.18	2.24	1.66	1.23	1.03

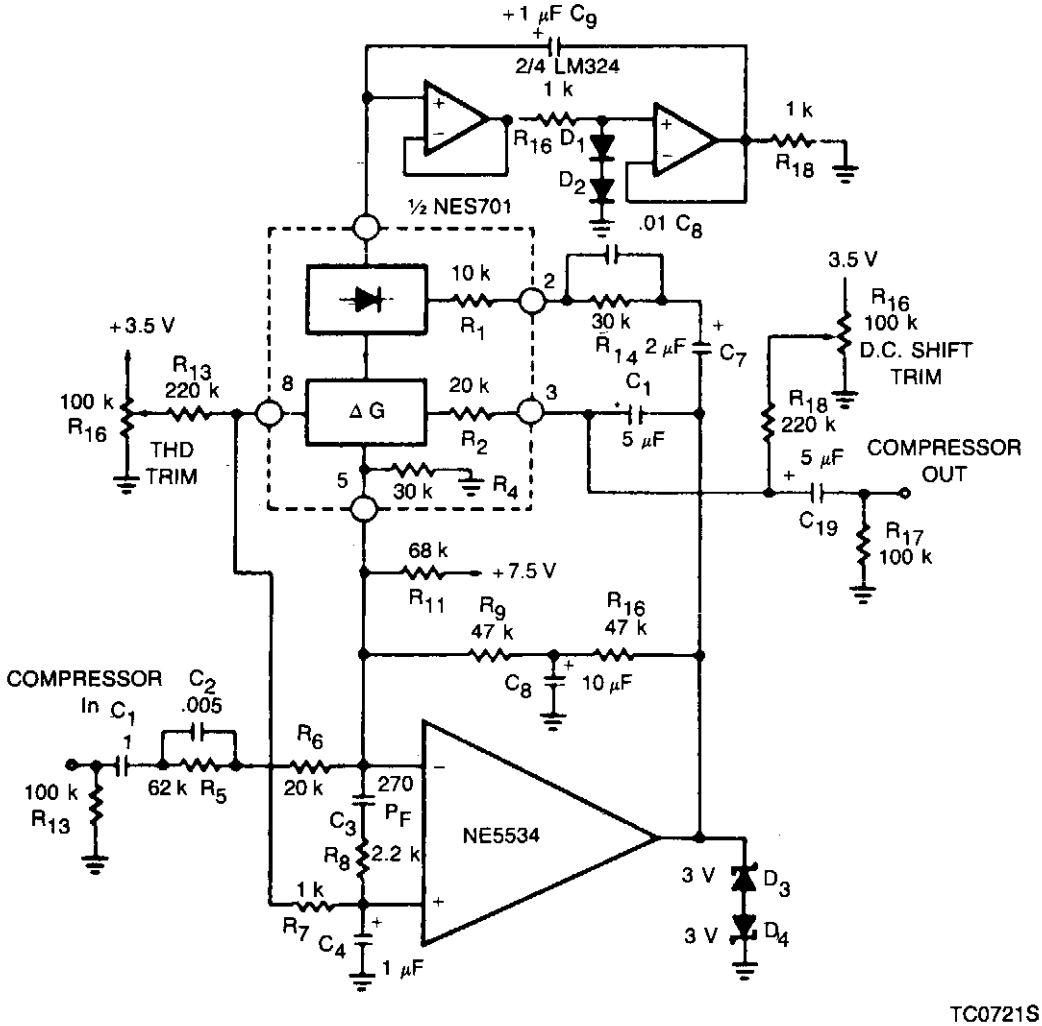
ELECTRONIC ENGINEERING

Fig. 2-12

Circuit Notes

To amplify small current signals such as from an electron-collector inside a vacuum chamber, it is convenient for reasons of noise and bandwidth to have a "head-amplifier" attached to the chamber. The op-amp N_1 is a precision bipolar device with extremely low bias current and offset voltage (1) as well as low noise, which allows the 100:1 feedback attenuator to be employed. The resistance of R_3 can be varied from above 10 M to below 1 k, and so the nominal 0 to 1 V-peak output signal corresponds to input current ranges of 1 nA to 10 μA.

HI-FI COMPANDER



TC0721S

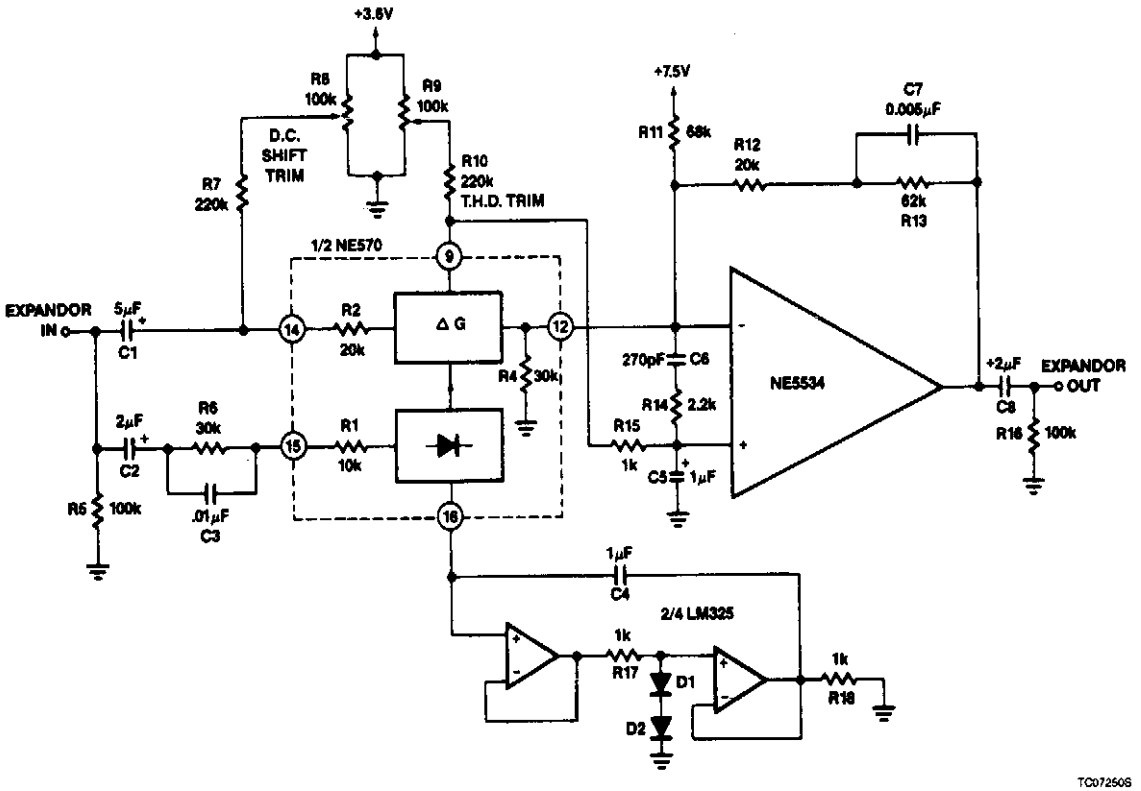
SIGNETICS

HI-FI Compressor With Pre-emphasis

Fig. 2-13(A)

Circuit Notes

This circuit for a high fidelity compressor uses an external op amp, and has a high gain and wide bandwidth. An input compensation network is required for stability. The rectifier capacitor (C_9) is not grounded, but is tied to the output of an op amp circuit. When a compressor is operating at high gain, (small input signal), and is suddenly hit with a signal, it will overload until



HI-FI Expandor With De-emphasis ***Fig. 2-13(B)***

it can reduce its gain. The time it takes for the compressor to recover from overload is determined by the rectifier capacitor C_9. The expandor to complement the compressor is shown in Fig. 2-13B. Here an external op amp is used for high slew rate. Both the compressor and expandor have unity gain levels of 0 dB. Trim networks are shown for distortion (THD) and dc shift. The distortion trim should be done first, with an input of 0 dB at 10 kHz. The dc shift should be adjusted for minimum envelope bounce with tone bursts.

TWO-WIRE TO FOUR-WIRE AUDIO CONVERTER

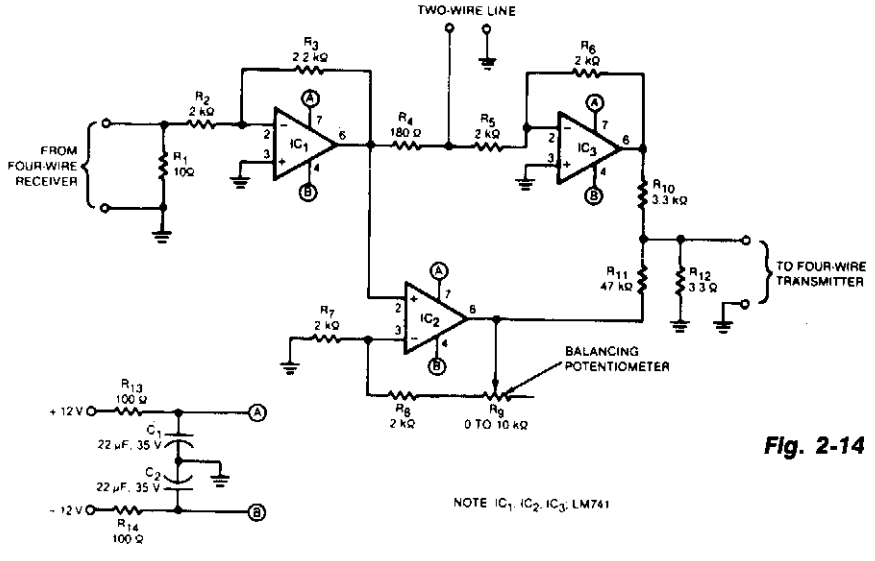


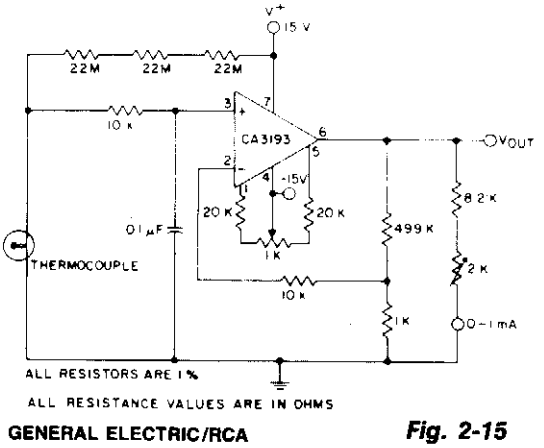
Fig. 2-14

NASA

Circuit Notes

This converter circuit maintains 40 dB of isolation between the input and output halves of a four-wire line while permitting a two-wire line to be connected. A balancing potentiometer, Rg, adjusts the gain of IC2 to null the feed-through from the input to the output. The adjustment is done on the workbench just after assembly by inserting a 1 kHz tone into the four-wire input and setting R_g for minimum output signal. An 82 ohm dummy-load resistor is placed across the two wire terminals.

THERMOCOUPLE AMPLIFIER



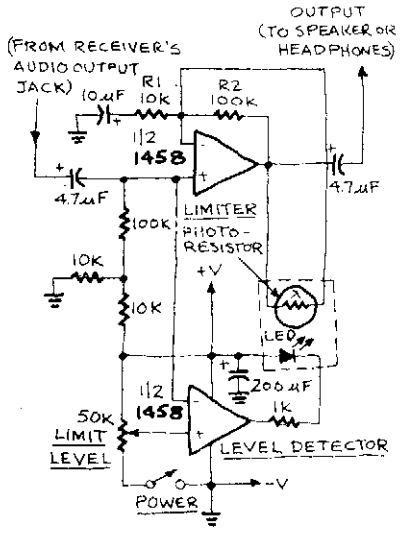
GENERAL ELECTRIC/RCA

Fig. 2-15

Circuit Notes

The circuit uses a CA3193 BiMOS precision op amp to amplify the generated signal 500 times. Three 22-megohm resistors will provide full-scale output if the thermocouple opens.

LOW-DISTORTION AUDIO LIMITER



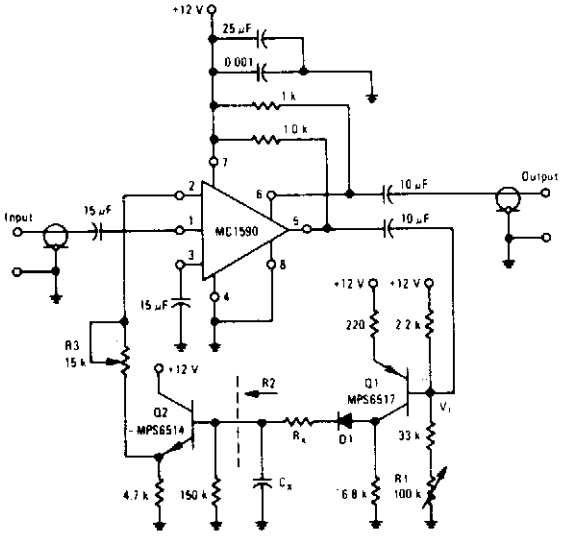
RADIO-ELECTRONICS

Fig. 2-16

Circuit Notes

The level at which the audio limiter comes into action can be set with the LIMIT LEVEL trimmer potentiometer. When that level is exceeded, the output from the LIMITER-DETECTOR half of the op-amp (used as a comparator) turns the LED which causes the resistance of the photoresistor to decrease rapidly. That in turn causes the gain of the LIMITER half of the op-amp to decrease. When the signal drops below the desired limiting level, the LED turns off, the resistance of the photoresistor increases, and the gain of the LIMITER op-amp returns to its normal level—that set by the combination of resistors R1 and R2. A dual-polarity power supply (±12 volts is desirable) is needed for the op-amp.

SPEECH COMPRESSOR



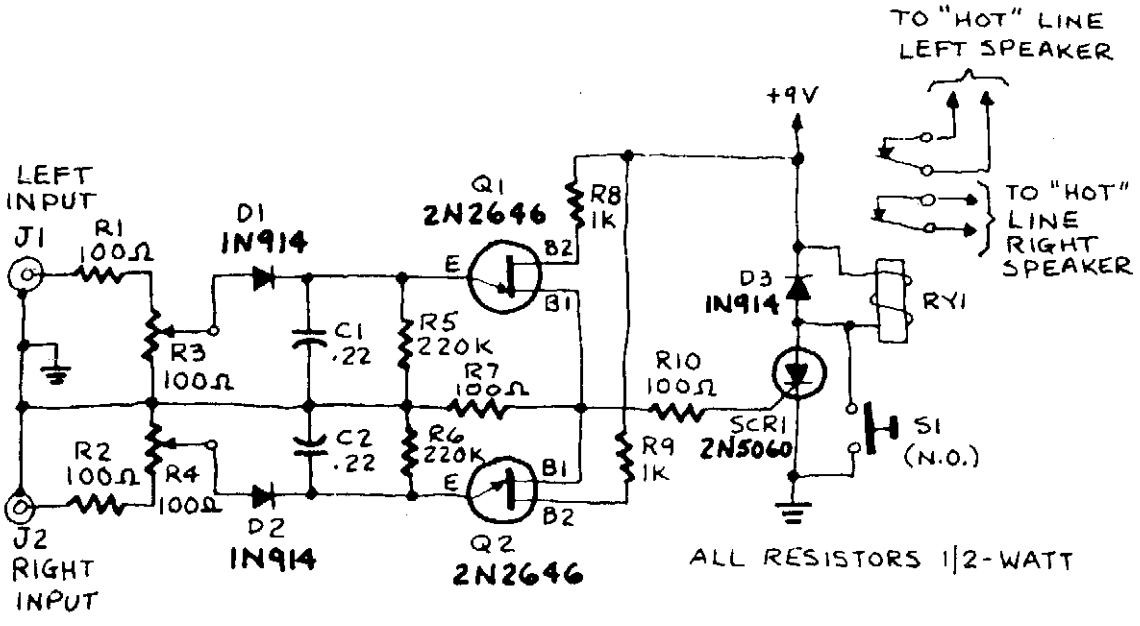
MOTOROLA INC.

Fig. 2-17

Circuit Notes

The amplifier drives the base of a pnp MPS6517 operating common-emitter with a voltage gain of approximately 20. The control R1 varies the quiescent Q point of this transistor so that varying amounts of signal exceed the level V_r. Diode D1 rectifies the positive peaks of Q1's output only when these peaks are greater than $V_r \simeq 7.0$ volts. The resulting output is filtered C_x, R_x. R_x controls the charging time constant or attack time. C_x is involved in both charge and discharge. R2 (150 K, input resistance of the emitter-follower Q2) controls the decay time. Making the decay long and attack short is accomplished by making R_x small and R2 large. (A Darlington emitter-follower may be needed if extremely slow decay times are required.) The emitter-follower Q2 drives the AGC Pin 2 of the MC1590 and reduces the gain. R3 controls the slope of signal compression.

SPEAKER OVERLOAD PROTECTOR



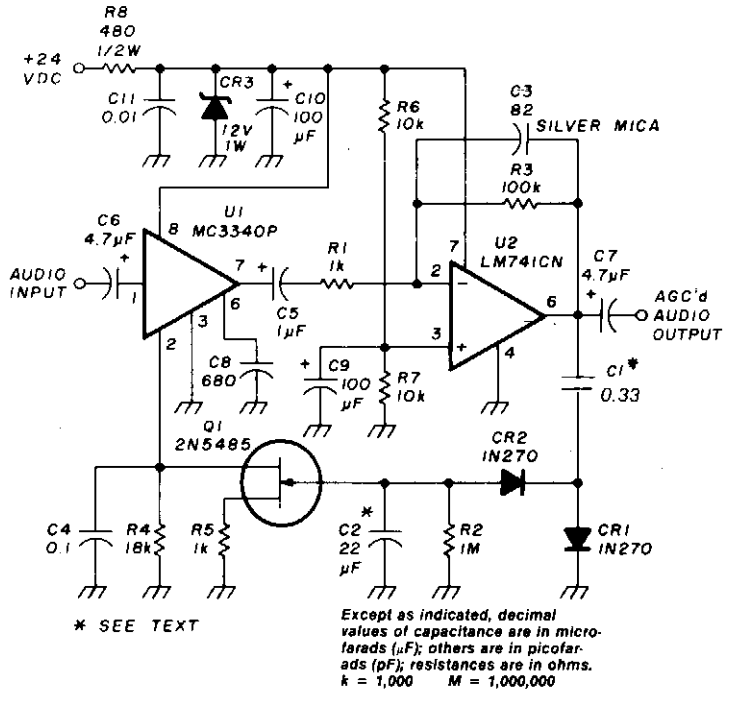
RADIO ELECTRONICS

Fig. 2-18

Circuit Notes

The input to the circuit is taken from the amplifier's speaker-output terminals or jacks. If the right-channel signal is sufficiently large to charge C1 to a potential that is greater than the breakdown voltage of Q1's emitter, a voltage pulse will appear across R7. Similarly, if the left-channel signal is sufficiently large to charge C2 to a potential that is greater than the breakdown voltage of Q2's emitter, a pulse will appear across R7. The pulse across R7 triggers SCR1, a sensitive gate SCR (I_{GT} less than 15 mA where I_{GT} is the gate trigger-current), that latches in a conducting state and energizes RY1. The action of the relay will interrupt both speaker circuits, and the resulting silence should alert you to the problem. Cut back the volume on your amplifier, then press and release S1 to reset the circuit and restore normal operation. The circuit can be adjusted to trip at any level from 15 to 150 watts RMS. To calibrate, deliberately feed an excessive signal to the right input of the speaker protector and adjust R3 until RY1 energizes. Do the same with the left channel, this time adjusting R4. The circuit is now calibrated and ready for use.

AUDIO AUTOMATIC GAIN CONTROL



HAM RADIO

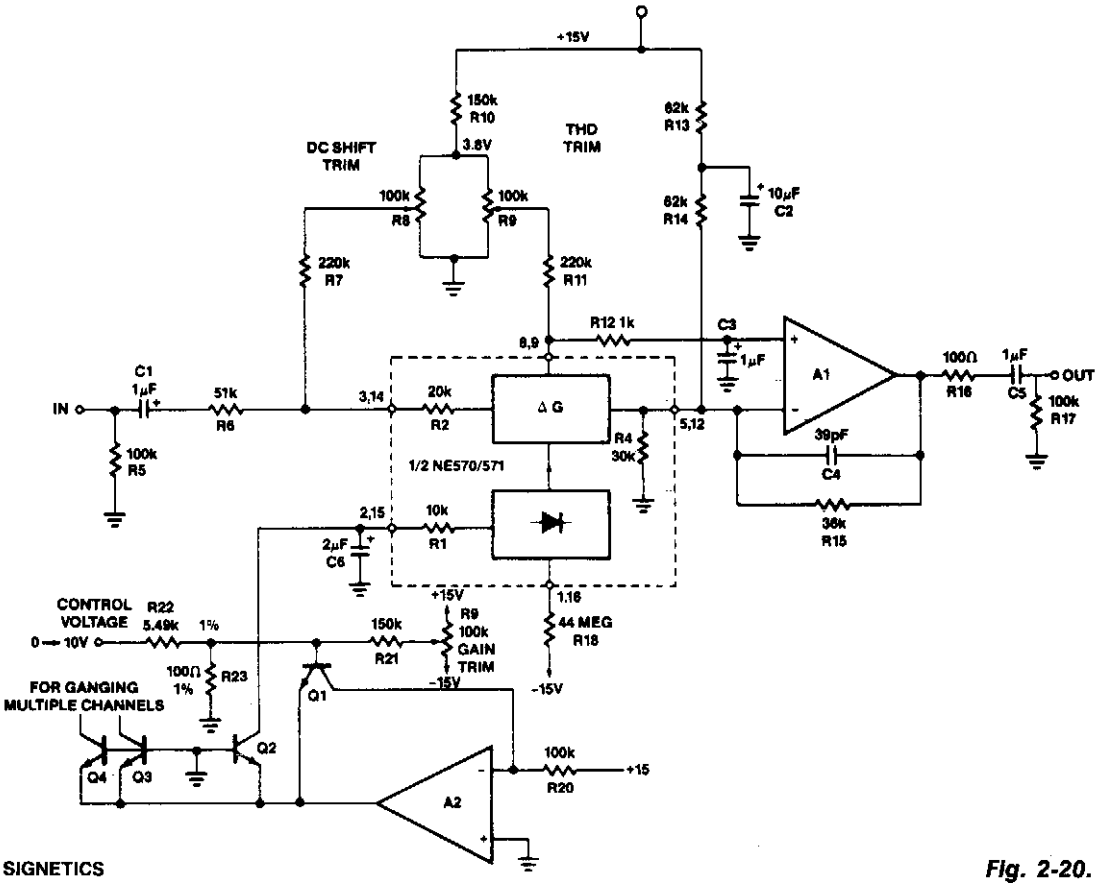
Fig. 2-19

Circuit Notes

An audio signal applied to U1 is passed through to the 741 operational amplifier, U2. After being amplified, the output signal of U2 is sampled and applied to a negative voltage doubler/rectifier circuit composed of diodes CR1 and CR2 along with capacitor C1. The resulting negative voltage is used as a control voltage that is applied to the gate of the 2N5485 JFET Q1. Capacitor C2 and resistor R2 form a smoothing filter for the rectified audio control voltage.

The JFET is connected from pin 2 of the MC3340P to ground through a 1 kilohm resistor. As the voltage applied to the gate of the JFET becomes more negative in magnitude, the channel resistance of the JFET increases causing the JFET to operate as a voltage controlled resistor. The MC3340P audio attenuator is the heart of the AGC. It is capable of 13 dB gain or nearly −80 dB of attenuation depending on the external resistance placed between pin 2 and ground. An increase of resistance decreases the gain achieved through the MC3340P. The circuit gain is not entirely a linear function of the external resistance but approximates such behavior over a good portion of the gain/attenuation range. An input signal applied to the AGC input will cause the gate voltage of the JFET to become proportionally negative. As a result the JFET increases the resistance from pin 2 to ground of the MC3340P causing a reduction in gain. In this way the AGC output is held at a nearly constant level.

VOLTAGE-CONTROLLED ATTENUATOR



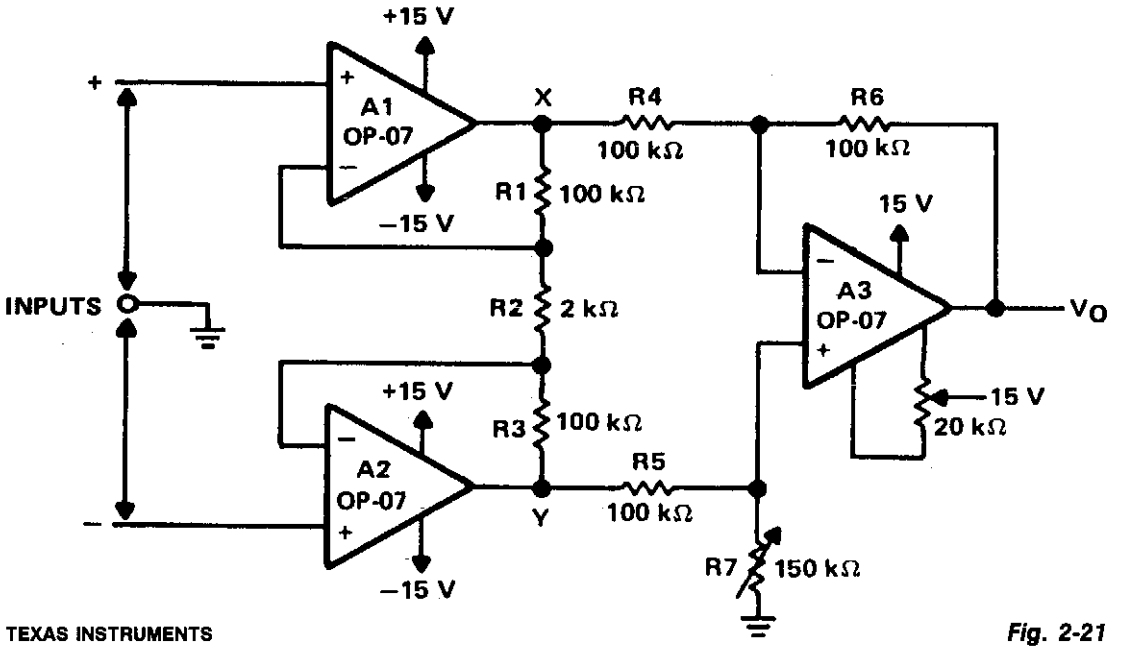
SIGNETICS

Fig. 2-20.

Circuit Notes

Op amp A_2 and transistors Q_1 and Q_2 form the exponential converter generating an exponential gain control current, which is fed into the rectifier. A reference current of 150 µA, (15 V and R_{20} = 100 k), is attenuated a factor of two (6 dB) for every volt increase in the control voltage. Capacitor C_6 slows down gain changes to a 20 ms time constant ($C_6 \times R_1$) so that an abrupt change in the control voltage will produce a smooth sounding gain change. R_{18} ensures that for large control voltages the circuit will go to full attentuation. The rectifier bias current would normally limit the gain reduction to about 70 dB. R_{16} draws excess current out of the rectifier. After approximately 50 dB of attentuation at a −6 dB/V slope, the slope steepens and attenuation becomes much more rapid until the circuit totally shuts off at about 9 V of control voltage. A_1 should be a low-noise high slew rate op amp. R_{13} and R_{14} establish approximately a 0 V bias at A_1's output.

HIGH-INPUT-IMPEDANCE DIFFERENTIAL AMPLIFIER



TEXAS INSTRUMENTS

Fig. 2-21

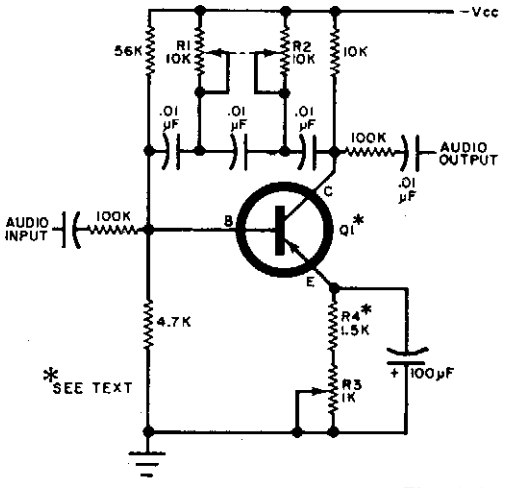
Circuit Notes

Operational amplifiers A1 and A2 are connected in a non-inverting configuration with their outputs driving amplifier A3. Operational amplifier A3 could be called a subtractor circuit which converts the differential signal floating between points X and Y into a single-ended output voltage. Although not mandatory, amplifier A3 is usually operated at unity gain and R4, R5, R6, and R7 are all equal.

The common-mode-rejection of amplifier A3 is a function of how closely the ratio R4:R5 matches the ratio R6:R7. For example, when using resistors with 0.1% tolerance, common-mode rejection is greater than 60 dB. Additional improvement can be attained by using a potentiometer (slightly higher in value than R6) for R7. The potentiometer can be adjusted for the best common-mode rejection. Input amplifiers A1 and A2 will have some differential gain but the common-mode input voltages will experience only unity gain. These voltages will not appear as differential signals at the input of amplifier A3 because, when they appear at equal levels on both ends of resistor R2, they are effectively canceled.

This type of low-level differential amplifier finds widespread use in signal processing. It is also useful for dc and low-frequency signals commonly received from a transducer or thermocouple output, which are amplified and transmitted in a single-ended mode. The amplifier is powered by ± 15 V supplies. It is only necessary to null the input offset voltage of the output amplifier A3.

AUDIO Q-MULTIPLIER



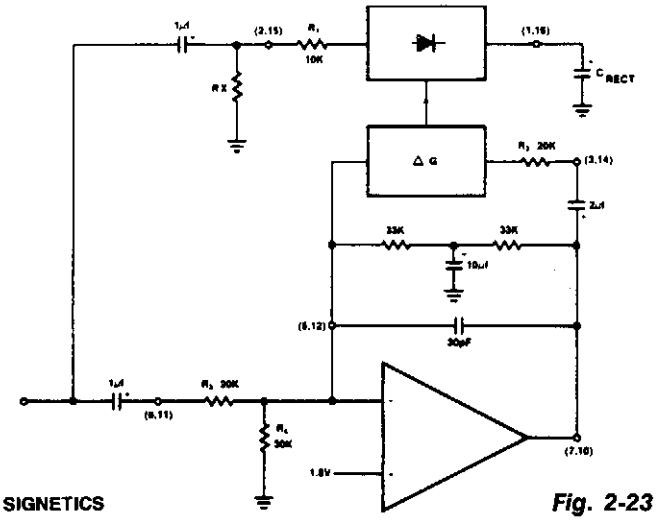
POPULAR ELECTRONICS

Fig. 2-22

Circuit Notes

This circuit is for selective tuning between two closely spaced audio tones. The selective frequency is dependent on the value of capacitors and resistors in the feedback circuit between the collector and base of Q1. With the values shown, the frequency can be "tuned" a hundred cycles or so around 650 Hz. R1 and R2 must be ganged. Emitter potentiometer R3 determines the sharpness of response curve. Any transistor having a beta greater than 50 can be used. Select a value for R4 so that the circuit will not oscillate when R3 is set for minimum bandwidth (sharpest tuning).

AUTOMATIC LEVEL CONTROL



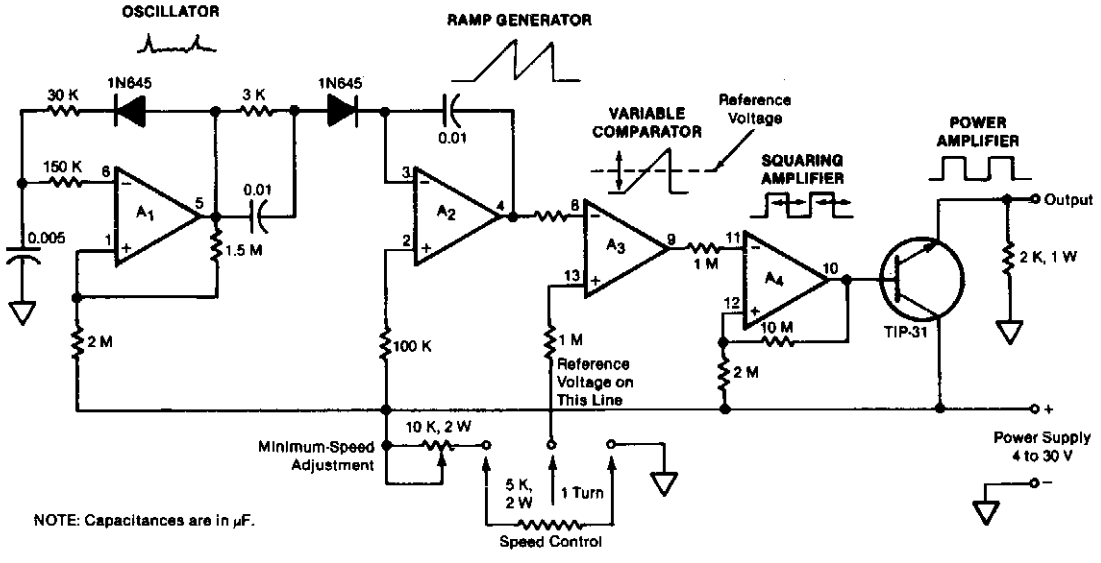
SIGNETICS

Fig. 2-23

Circuit Notes

The NE570 can be used to make a very high performance ALC compressor, except that the rectifier input is tied to the input. This makes gain inversely proportional to input level so that a 20 dB drop in input level will produce a 20 dB increase in gain. The output will remain fixed at a constant level. As shown, the circuit will maintain an output level of ±1 dB for an input range of +14 to −43 dB at 1 kHz. Additional external components will allow the output level to be adjusted.

PULSE-WIDTH PROPORTIONAL-CONTROLLER CIRCUIT



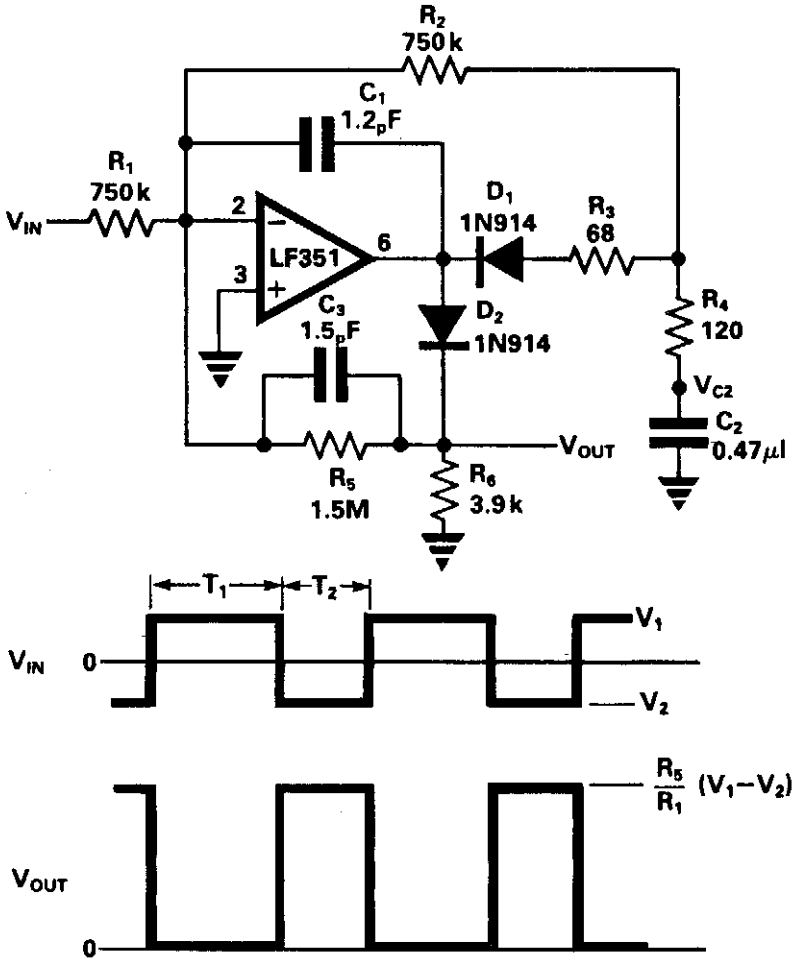
NASA

Fig. 2-24

Circuit Notes

The quad operational amplifier circuit yields full 0 to 100 percent pulse-width control. The controller uses an LM3900 that requires only a single supply voltage of 4 to 30 V. The pulse-repetition rate is set by a 1 kHz oscillator that incorporates amplifier A_1. The oscillator feeds ramp generator A_2, which generates a linear ramp voltage for each oscillator pulse. The ramp signal feeds the inverting input of comparator A_3; the speed-control voltage feeds the noninverting input. Thus, the output of the comparator is a 1 kHz pulse train, the pulse width of which changes linearly with the control voltage. The control voltage can be provided by an adjustable potentiometer or by an external source of feedback information such as a motor-speed sensing circuit. Depending on the control-voltage setting, the pulse duration can be set at any value from zero (for zero average dc voltage applied to the motor) to the full pulse-repetition period (for applied motor voltage equal to dc power-supply voltage). An amplifier stage (A_4) with a gain of 10 acts as a pulse-squaring circuit. A TIP-31 medium-power transistor is driven by A_4 and serves as a separate power-amplifier stage.

OP AMP CLAMPING



ELECTRONIC ENGINEERING

Fig. 2-25

Circuit Notes

The circuit clamps the most positive value of the input pulse signal to the zero base level. Additionally, the circuit inverts and amplifies the input signal by the factor of R_5/R_1. The waveforms are shown in the bottom of Fig. 2-24.

3

Analog-to-Digital Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Successive Approximation A/D Converters
4 Digit (10,000 Count) A/D Converter
16-Bit A/D Converter
Inexpensive, Fast 10-Bit Serial Output A/D
10-Bit A/D Converter
High Speed 12-Bit A/D Converter
Successive Approximation A/D Converter
Cyclic A/D Converter
Differential Input A/D System

SUCCESSIVE APPROXIMATION A/D CONVERTERS

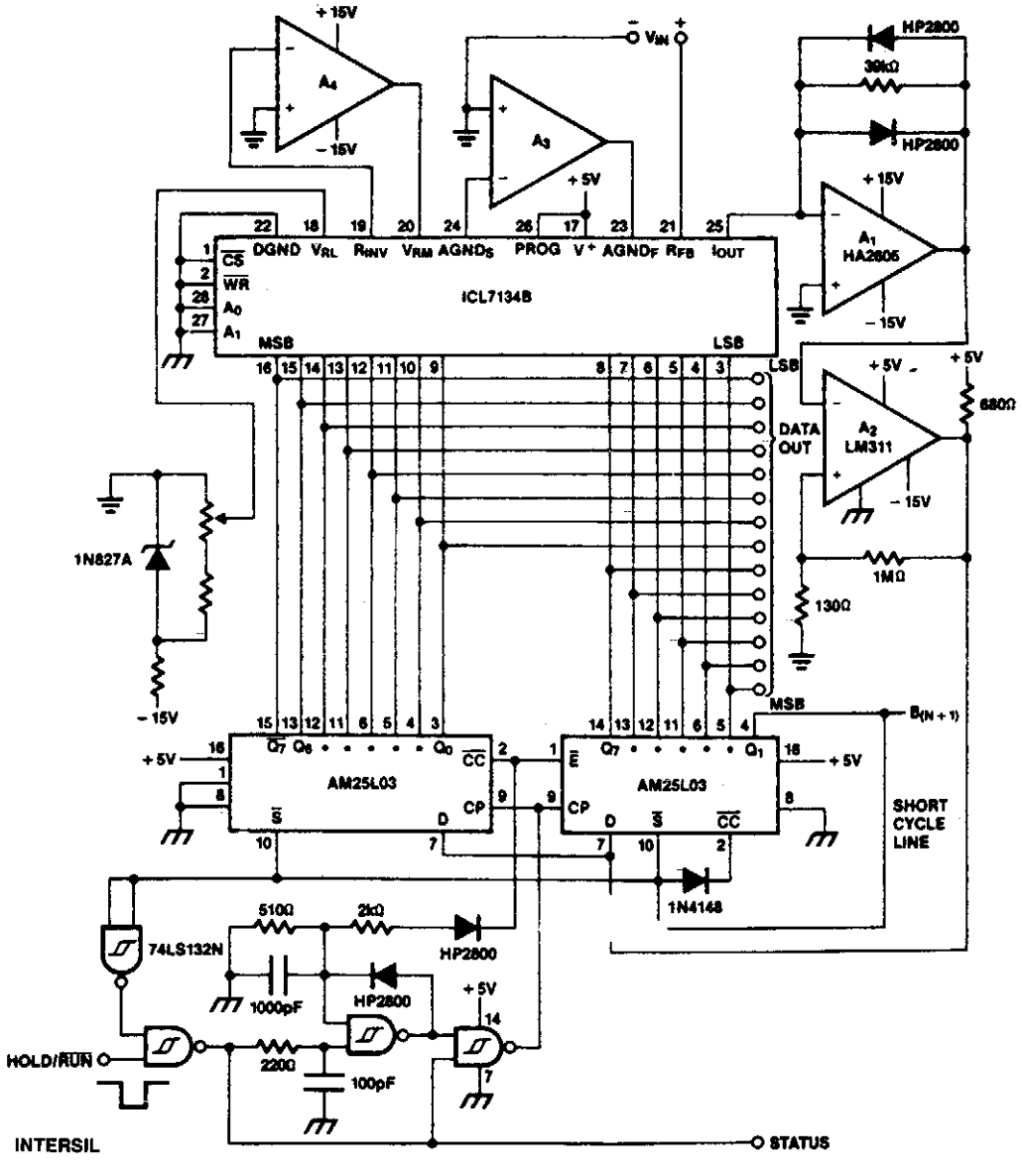


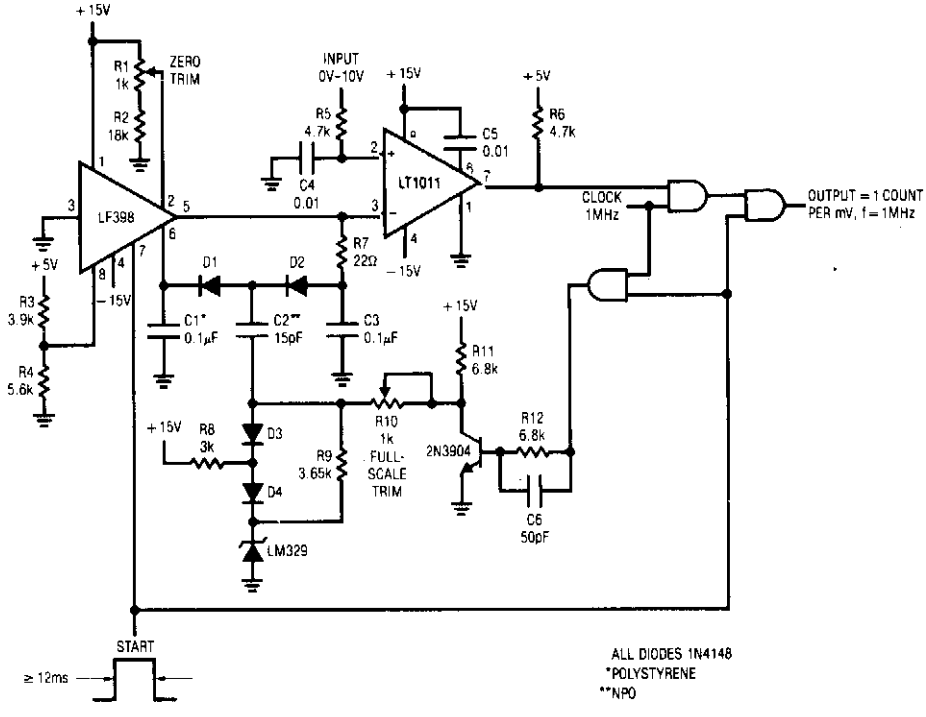
Fig. 3-1

SUCCESSIVE APPROXIMATION A/D CONVERTERS , Continued.

Circuit Notes

The ICL7134B-based circuit is for a bipolar-input high-speed A/D converter, using two AM25L03s to form a 14-bit successive approximation register. The comparator is a two-stage circuit with an HA2605 front-end amplifier, used to reduce settling time problems at the summing node (see A020). Careful offset-nulling of this amplifier is needed, and if wide temperature range operation is desired, an auto-null circuit using an ICL7650 is probably advisable (see A053). The clock, using two Schmitt trigger TTL gates, runs at a slower rate for the first 8 bits, where settling-time is most critical than for the last 6 bits. The short-cycle line is shown tied to the 15th bit; if fewer bits are required, it can be moved up accordingly. The circuit will free-run if the HOLD/RUN input is held low, but will stop after completing a conversion if the pin is high at that time. A low-going pulse will restart it. The STATUS output indicates when the device is operating, and the falling edge indicates the availability of new data. A unipolar version can be constructed by typing the MSB (D13) on an ICL7134U to pin 14 on the first AM25L03, deleting the reference inversion amplifier A4, and tying V_{RFM} to V_{RFL}.

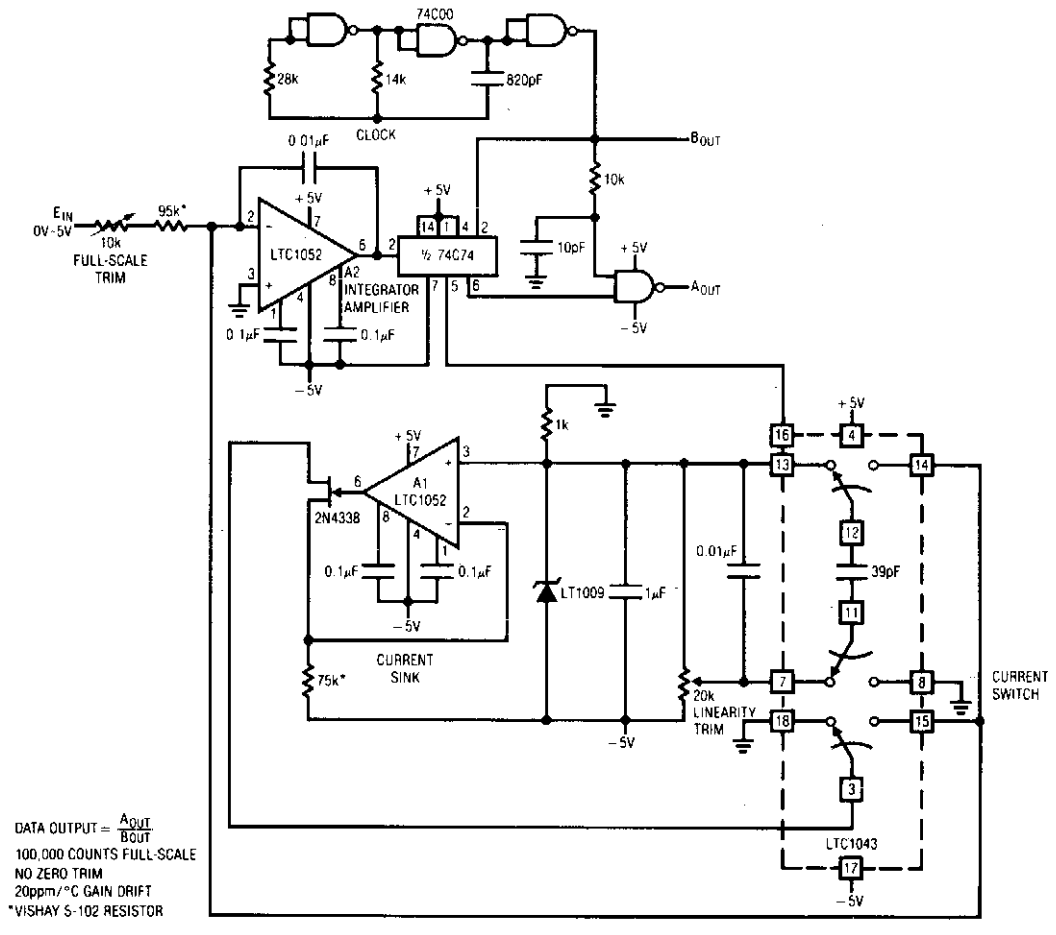
4 DIGIT (10,000 COUNT) A/D CONVERTER



LINEAR TECHNOLOGY CORP.

Fig. 3-2

16-BIT A/D CONVERTER



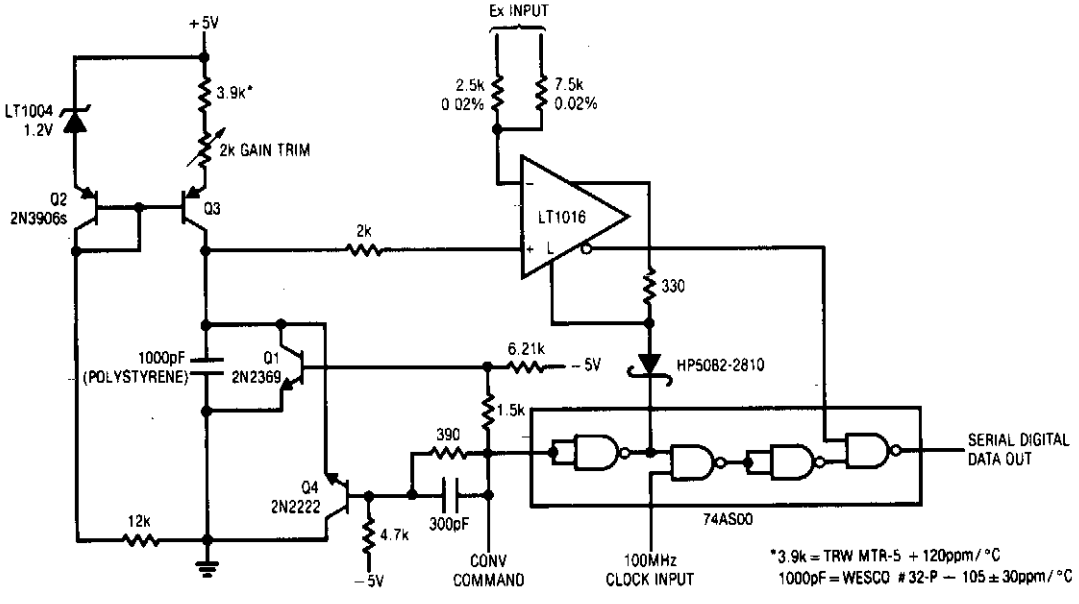
LINEAR TECHNOLOGY

Fig. 3-3

Circuit Notes

The A/D converter, made up of A2, a flip-flop, some gates and a current sink, is based on a current balancing technique. Once again, the chopper-stabilized LTC1052's 50 nV/°C input drift is required to eliminate offset errors in the A/D.

INEXPENSIVE, FAST 10-BIT SERIAL OUTPUT A/D



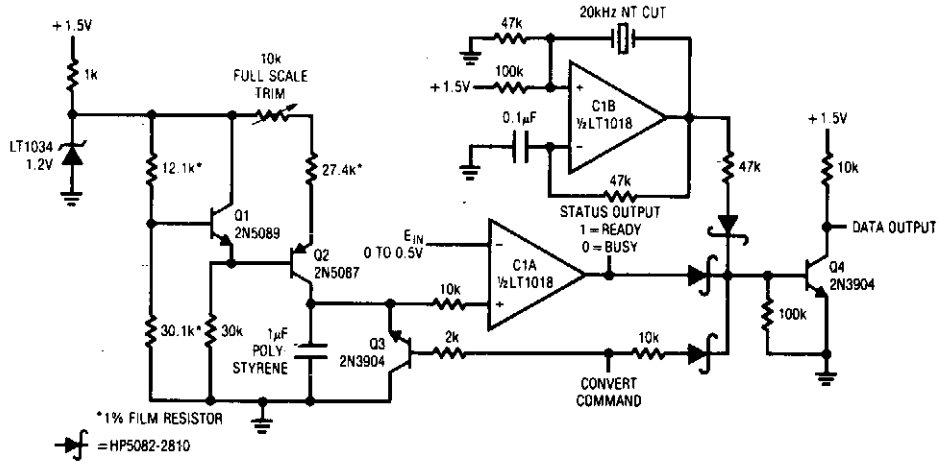
LINEAR TECHNOLOGY CORPORATION

Fig. 3-4

Circuit Notes

Everytime a pulse is applied to the convert command input, Q1 resets the 1000 pF capacitor to 0 V. This resetting action takes 200 ns of the falling edge of the convert command pulse, the capacitor begins to charge linearly. In precisely 10 microseconds, it charges to 2.5 V. The 10 microseconds ramp is applied to the LT1016's positive input. The LT1016 compares the ramp to Ex, the unknown, at its negative input. For a 0 V - 2.5 V range, Ex is applied to the 2.5 k ohm resistor. From a 0 V - 10 V range, the 2.5 k ohm resistor is grounded and Ex is applied to the 7.5 k ohm resistor. Output of the LT1016 is a pulse whose width is directly dependent on the value of Ex. This pulse width is used to gate a 100 MHz clock. The 100 MHz clock pulse bursts that appear at the output are proportional to Ex. For a 0 V - 10 V input, 1024 pulses appear at full-scale, 512 at 5.00 V, etc.

10-BIT A/D CONVERTER



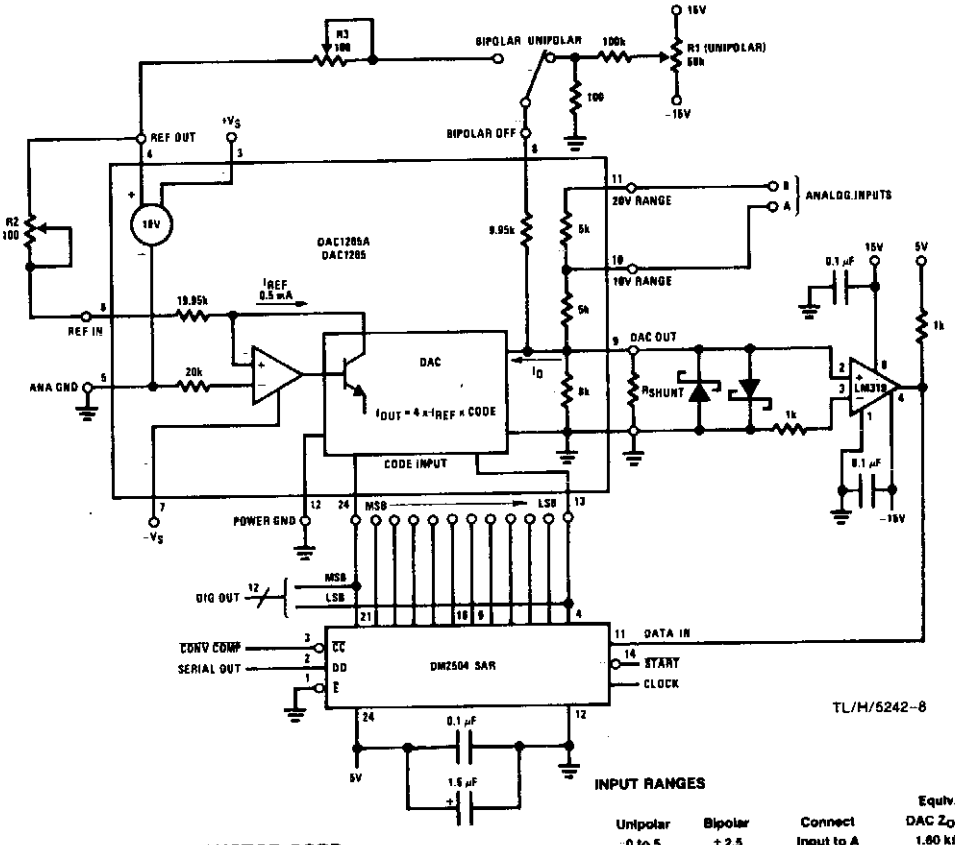
LINEAR TECHNOLOGY CORPORATION

Fig. 3-5

Circuit Notes

The converter has a 60 ms conversion time, consumes 460 μA from its 1.5 V supply and maintains 10 bit accuracy over a 15°C to 35°C temperature range. A pulse applied to the convert command line causes Q3, operating in inverted mode, to discharge through the 10 kΩ diode path, forcing its collector low. Q3's inverted mode switching results in a capacitor discharge within 1 mV of ground. During the time the ramps' value is below the input voltage, CIA's output is low. This allows pulses from C1B, a quartz stabilized oscillator, to modulate Q4. Output data appears at Q4's collector. When the ramp crosses the input voltages value C1A's output goes high, biasing Q4 and output data ceases. The number of pulses at the output is directly proportional to the input voltage. To calibrate apply 0.5 V to the input and trim the 10 kΩ potentiometer for exactly 1000 pulses out each time the convert command line is pulsed.

HIGH SPEED 12-BIT A/D CONVERTER



INPUT RANGES

Unipolar	Bipolar	Connect	Equiv. DAC Z_{OUT}
0 to 5	±2.5	Input to A B to DAC OUT	1.60 kΩ
0 to 10	±5	Input to A	2.35 kΩ
0 to 20	±10	Input to B	3.08 kΩ

NATIONAL SEMICONDUCTOR CORP.

Fig. 3-6

Circuit Notes

This system completes a full 12-bit conversion in 10 µs unipolar or bipolar. This converter will be accurate to ± ½ LSB of 12 bits and have a typical gain TC of 10 ppm/°C. In the unipolar mode, the system range is 0 V to 9.9976 V, with each bit having a value of 2.44 mV. For the true conversion accuracy, an A/D converter should be trimmed so that given bit code output results from input levels from ½ LSB below to ½ LSB above the exact voltage which that code represents. Therefore, the converter zero point should be trimmed with an input voltage of 1.22 mV; trim R1 until the LSB just begins to appear in the output code (all other bits "0"). For full-scale, use an input voltage of 9.9963 V (10 V-1 LSB-½ LSB); then trim R2 until the LSB just begins to appear (all other bits "1"). The bipolar signal range is −5.0 V to 4.9976 V. Bipolar offset trimming is done by applying a −4.9988 V input signal and trimming R3 for the LSB transition (all other bits "0"). Full-scale is set by applying 4.9963 V and trimming R2 for the LSB transition (all other bits "1").

SUCCESSIVE APPROXIMATION A/D CONVERTER

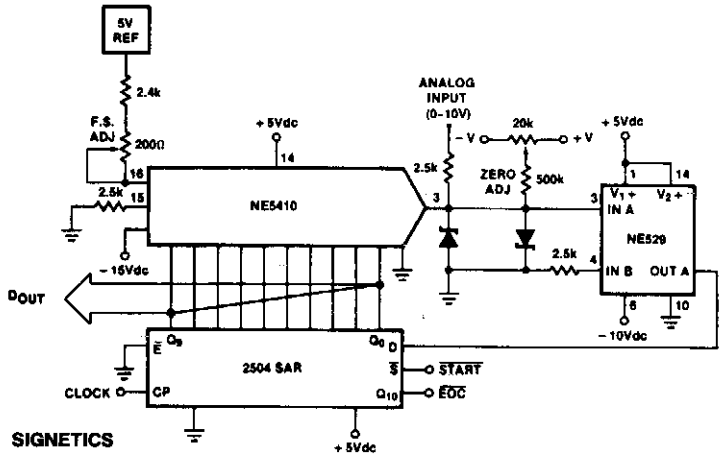
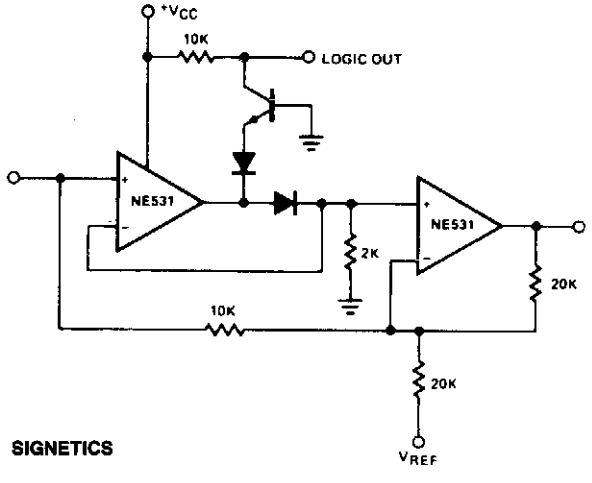


Fig. 3-7

Circuit Notes

The 10-bit conversion time is 3.3 μs with a 3 MHz clock. This converter uses a 2504 12-bit successive approximation register in the short cycle operating mode where the end of conversion signal is taken from the first unused bit of the SAR (Q_{10}).

CYCLIC A/D CONVERTER



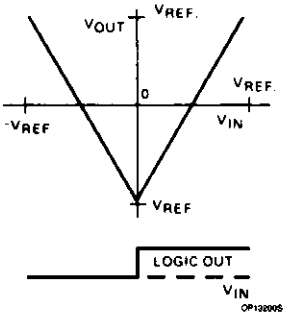
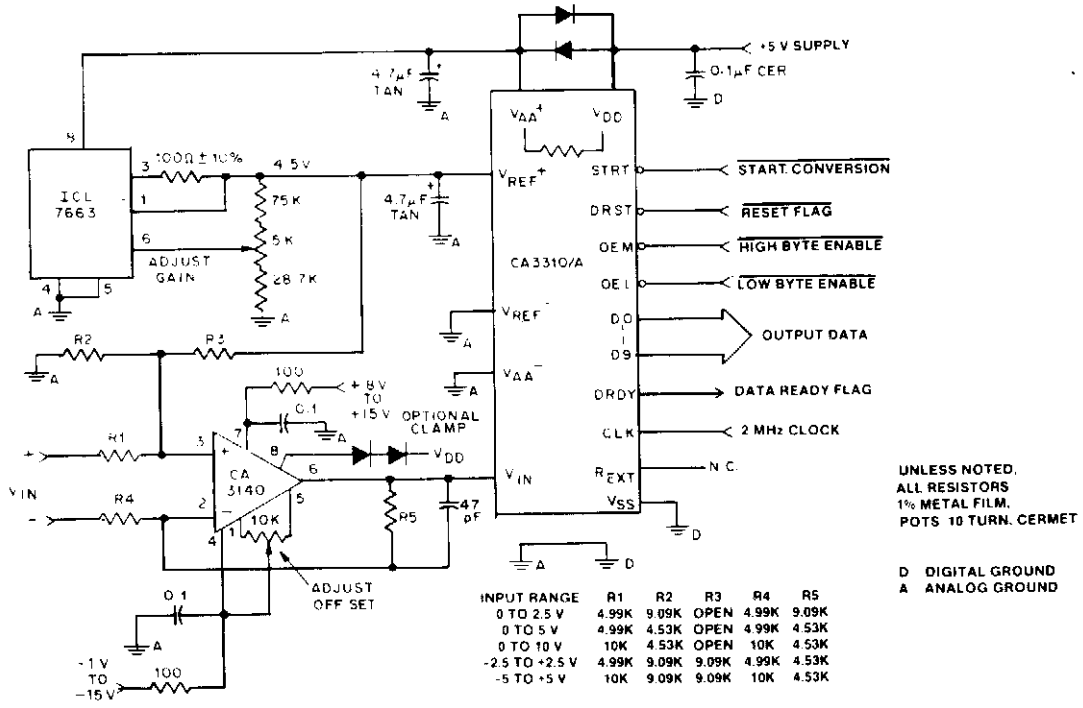


Fig. 3-8

Circuit Notes

The cyclic converter consists of a chain of identical stages, each of which senses the polarity of the input. The stage then subtracts V_{REF} from the input and doubles the remainder if the polarity was correct. The signal is full-wave rectified and the remainder of $V_{IN} - V_{REF}$ is doubled. A chain of these stages gives the gray code equivalent of the input voltage in digitized form related to the magnitude of V_{REF}. Possessing high potential accuracy, the circuit using NE531 devices settles in 5 μs.

DIFFERENTIAL INPUT A/D SYSTEM



INPUT RANGE	R1	R2	R3	R4	R5
0 TO 2.5 V	4.99K	9.09K	OPEN	4.99K	9.09K
0 TO 5 V	4.99K	4.53K	OPEN	4.99K	4.53K
0 TO 10 V	10K	4.53K	OPEN	10K	4.53K
-2.5 TO +2.5 V	4.99K	9.09K	9.09K	4.99K	4.53K
-5 TO +5 V	10K	9.09K	9.09K	10K	4.53K

GENERAL ELECTRIC/RCA

Fig. 3-9

Circuit Notes

Using a CA3140 BiMOS op amp provides good slewing capability for high bandwidth input signals, and can quickly settle energy that the CA3310 outputs at its V_{IN} terminal. The CA3140 can also drive close to the negative supply rail. If system supply sequencing or an unknown input voltage is likely to cause the op amp to drive above the V_{DD} supply, a diode clamp can be added from pin 8 of the op amp to the V_{DD} supply.

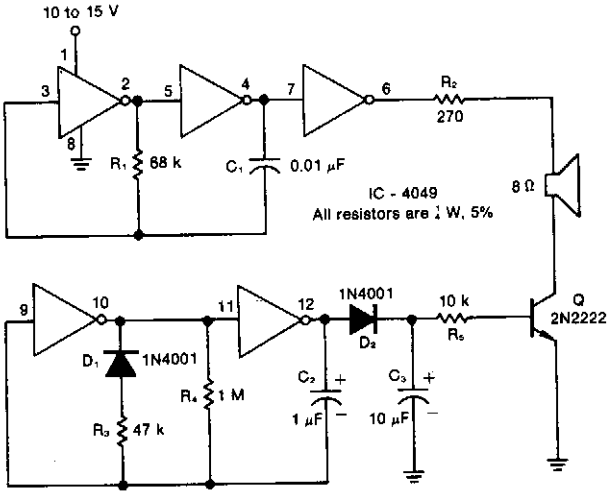
4

Annunciators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low-Cost Chime Circuit
Electronic Bell
Sliding-Tone Doorbell

LOW-COST CHIME CIRCUIT



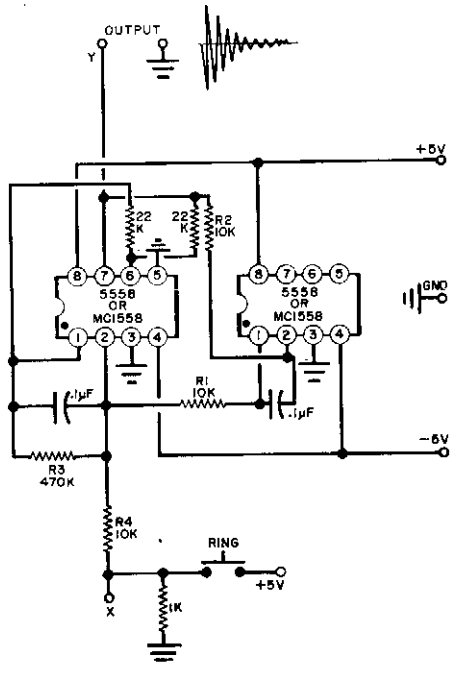
ELECTRONIC DESIGN

Fig. 4-1

Circuit Notes

Resistor R1, capacitor C1, and two inverters form a square wave generator, which produces the basic tone. The generator is followed by an inverter that acts as both a buffer and a driver for the speaker. Resistor R2, which has a minimum value of 100 ohms, limits the current and controls the volume. Diode D1, capacitor C2, resistors R3 and R4, and two inverters create the pulse generator that determines the turn-on and decay times of the chime. The decay circuit—formed by D2, C3, R5, and Q—reduces the amplitude of the chime tone exponentially as a function of time.

ELECTRONIC BELL



POPULAR ELECTRONICS

Fig. 4-2

Circuit Notes

Coarse frequency is controlled by the capacitors which must be kept nearly identical in value to each other for best results. Fine tuning is accomplished with R1 and R2. The decay time is controlled by R3.

SLIDING-TONE DOORBELL

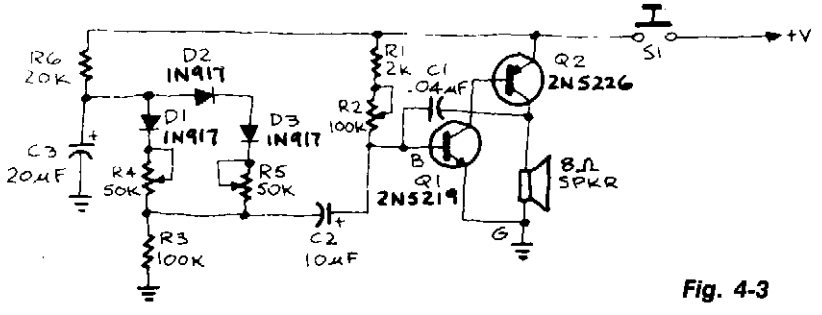


Fig. 4-3

RADIO ELECTRONICS

Circuit Notes

When the doorbell is pushed, you'll hear a low tone that will "slide up" to a higher frequency. The frequency of the AF oscillator is determined by coupling capacitor, C1 and the value of the resistance connected between the base of Q1 and ground. That resistance, R_{BG} is equal to (R1 + R2) R3. First, assume that S1 is closed and R2 has been adjusted to produce a pleasant, low-frequency tone. Capacitor C3 will charge through R6 until it reaches such a voltage that it will cause diode D1 to conduct. When that happens, the value of R_{BG} is paralleled by R4. Thus, because the total resistance R_{BG} decrease, the output tone slides up in frequency. Capacitor C3 will continue to charge until the voltage across D2 and D3 causes those diodes to conduct. Then R_{BG} is paralleled also by R5, the total resistance again decreases, and the oscillator's frequency again increases.

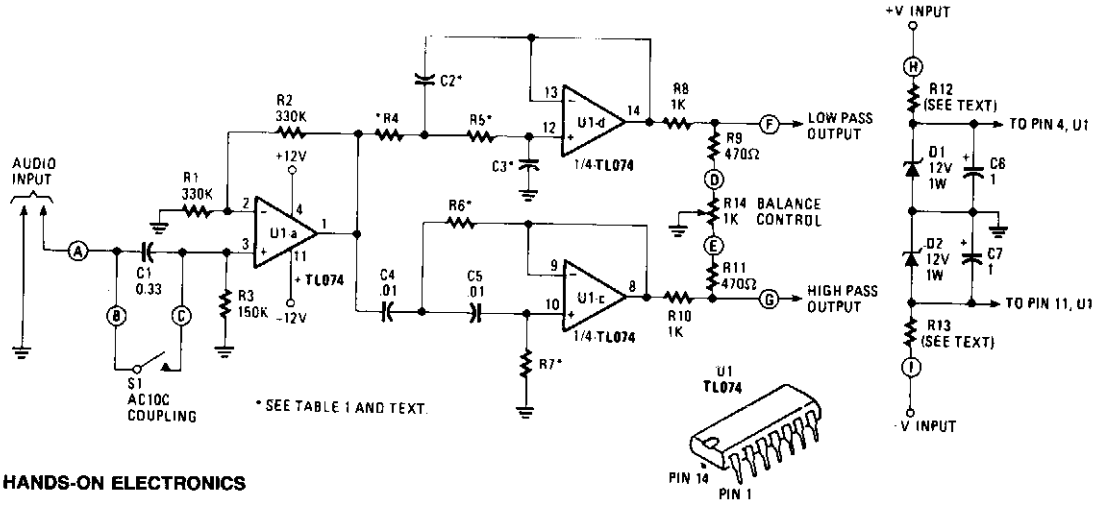
5

Audio Mixers, Crossovers and Distribution Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Electronic Crossover Circuit
Sound Mixer Amplifier
Microphone Mixer
Low Distortion Input Selector for Audio Use
Audio Distribution Amplifier
Four Channel Four Track Mixer

ELECTRONIC CROSSOVER CIRCUIT



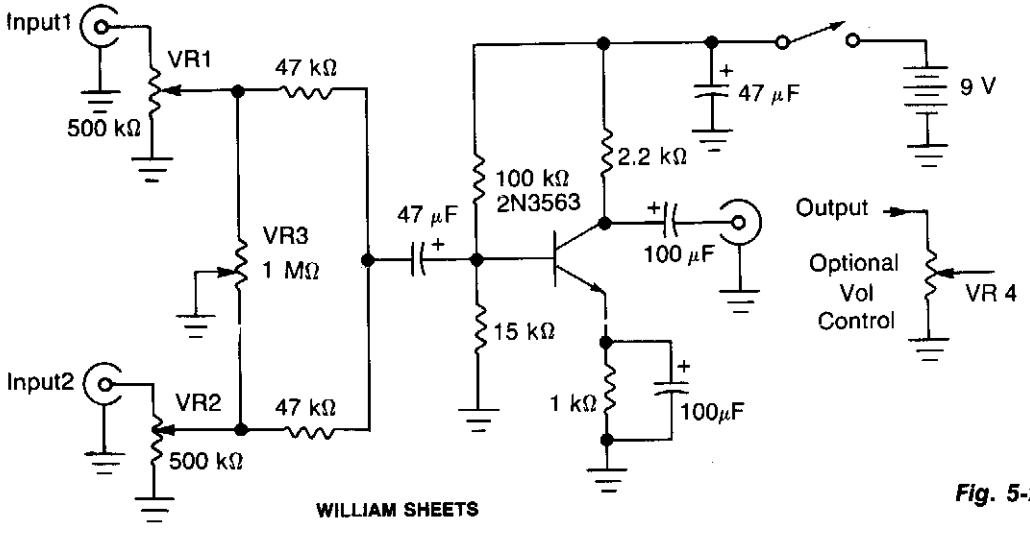
HANDS-ON ELECTRONICS

Fig. 5-1

Circuit Notes

An audio source, such as a mixer, preamplifier, equalizer, or recorder, is fed to the Electronic Crossover Circuit's input. That signal is either ac- or dc-coupled, depending on the setting of switch S1, to the non-inverting input of buffer-amplifier U1a, one section of a quad, BIFET, low-noise TL074 op amp made by Texas Instruments. That stage has a gain of 2, and its output is distributed to both a lowpass filter made by R4, R5, C2, C3, and op-amp U1d, and a highpass filter made by R6, R7, C4, C5, and op amp U1c. Those are 12 dB/octave Butterworth-type filters. The Butterworth filter response was chosen because it gives the best compromise between damping and phase shift. Values of capacitors and resistors will vary with the selected crossover at which your unit will operate. The filter's outputs are fed to a balancing network made by R8, R9, R10, R11 and balance potentiometer R14. When the potentiometer is at its mid-position, there is unity gain for the passbands of both the high and low filters. Dc power for the Electronic Crossover Circuit is regulated by R12, R13, D1, and D2, and decoupled by C6 and C7.

SOUND MIXER/AMPLIFIER



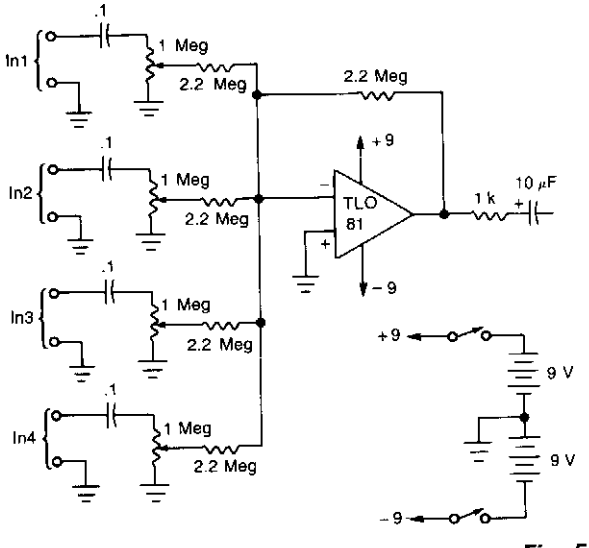
WILLIAM SHEETS

Fig. 5-2

Circuit Notes

Both input signals can be independently controlled by VR1 and VR2. The balance control VR3 is used to fade out one signal while simultaneously fading in the other. The transistor provides gain and the combined output signal level is controlled by VR4 (optional).

MICROPHONE MIXER



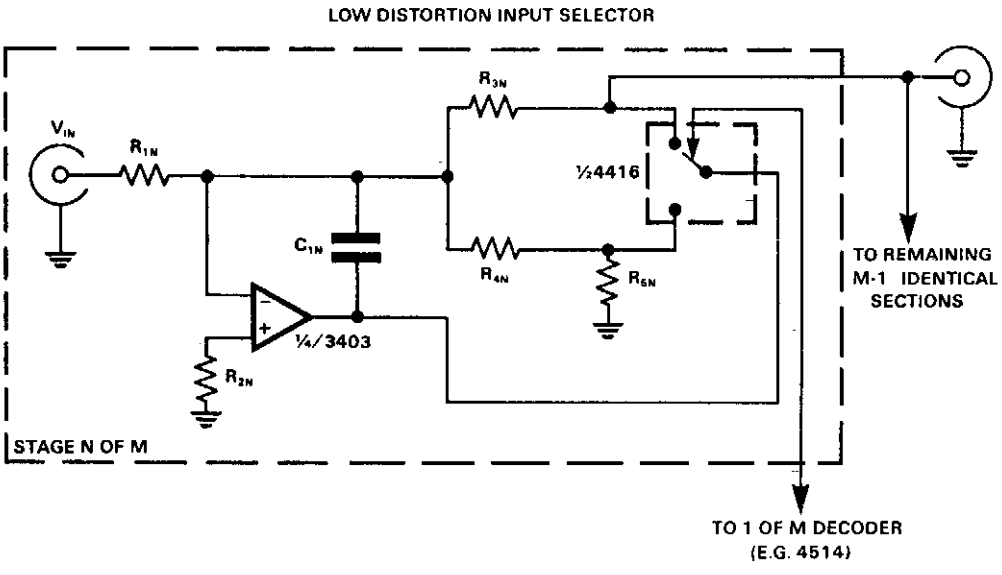
WILLIAM SHEETS

Fig. 5-3

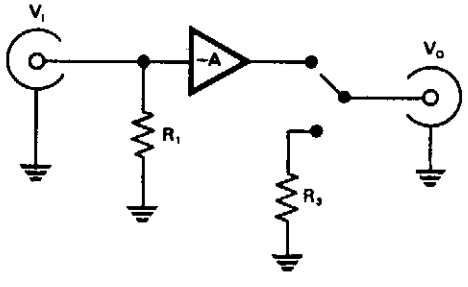
Circuit Notes

A TL081 op amp is used as a high-to-low impedance converter and signal mixer. The input impedance is approximately 1 megohm and the output impedance is about 1 kilohm. Two 9-volt batteries are used as the power source. Battery life should be several hundred hours with alkaline batteries.

LOW DISTORTION INPUT SELECTOR FOR AUDIO USE



EQUIVALENT CIRCUIT OF EACH STAGE:



$$R_{3N}=R_{4N}=AR_{1N}$$

$$R_{2N}=(R_{1N}+R_5)//R_{3N}//(R_{4N}+R_{5N})$$

$$\frac{1}{2\pi f_{MAX}} \gg R_{3N}C_{1N} \gg r_s$$

$$R_{5N}=\frac{1}{R_L^{-1}+\sum_{i=1}^{M}(R_{3i})^{-1}}$$

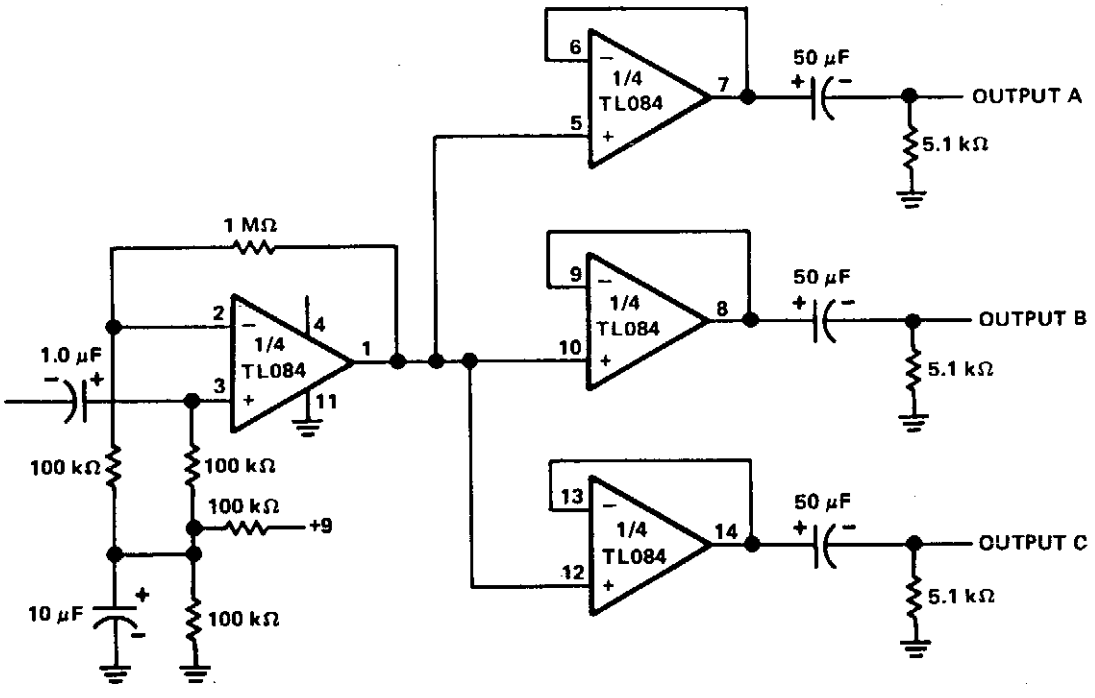
ELECTRONIC ENGINEERING

Fig. 5-4

Circuit Notes

CMOS switches are used directly to select inputs in audio circuits, this can introduce unacceptable levels of distortion, but if the switch is included in the feedback network of an op amp, the distortion due to the switch can be almost eliminated. The circuit uses a 4416 CMOS switch, arranged as two independent SPDT switches. If switching transients are unimportant, R5 and C1 can be omitted, and R4 can be shorted out. However, a feedback path must be maintained, even when a channel is switched out, in order to keep the inverting input of the op amp at ground potential, and prevent excessive crosstalk between channels.

AUDIO DISTRIBUTION AMPLIFIER



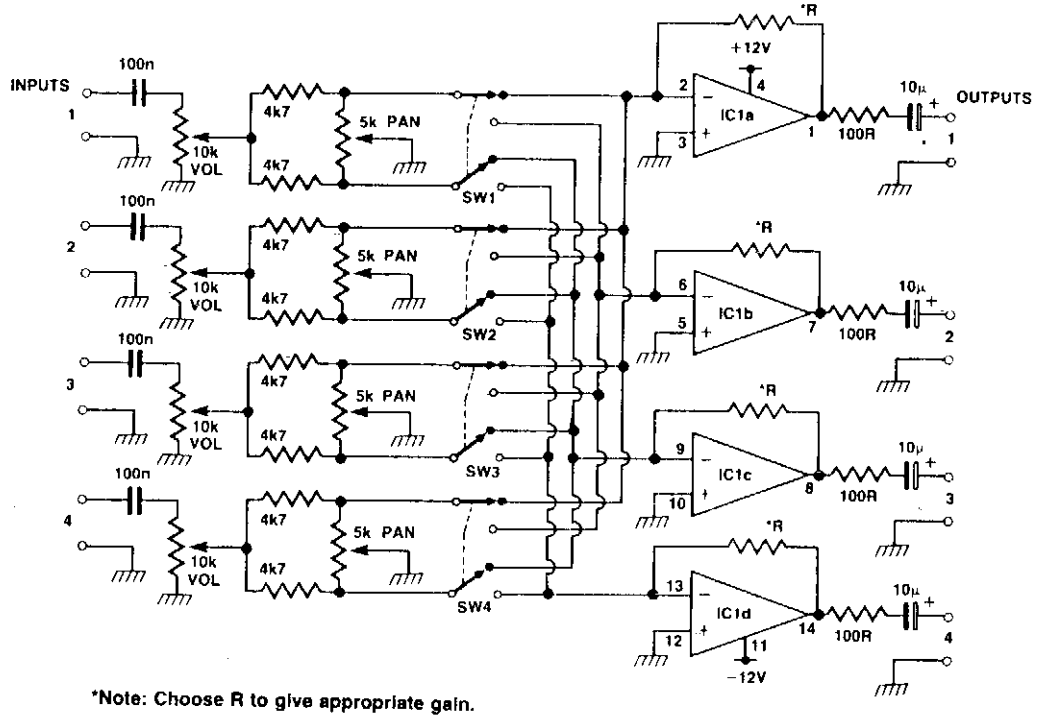
TEXAS INSTRUMENTS

Fig. 5-5

Circuit Notes

The three channel output distribution amplifier uses a single TL084. The first stage is capacitively coupled with a 1.0 μF electrolytic capacitor. The inputs are at ½ V_{CC} rail or 4.5 V. This makes it possible to use a single 9 V supply. A voltage gain of 10 (1 M ohm/100 k ohm) is obtained in the first stage, and the other three stages are connected as unity-gain voltage followers. Each output stage independently drives an amplifier through the 50 μF output capacitor to the 5.1 k ohm load resistor. The response is flat from 10 Hz to 30 kHz.

FOUR CHANNEL FOUR TRACK MIXER



*Note: Choose R to give appropriate gain.

ELECTRONICS TODAY INTERNATIONAL

Fig. 5-6

Circuit Notes

This circuit can be used as a stereo mixer as well as a four track. The quad op-amp IC gives a bit of gain for each track. The pan control allows panning between tracks one and two with the switch in the up position, and with the switch in the down position, it makes possible panning between tracks three and four. Extra channels can be added. A suitable op amp for IC1 is TL074 or similar.

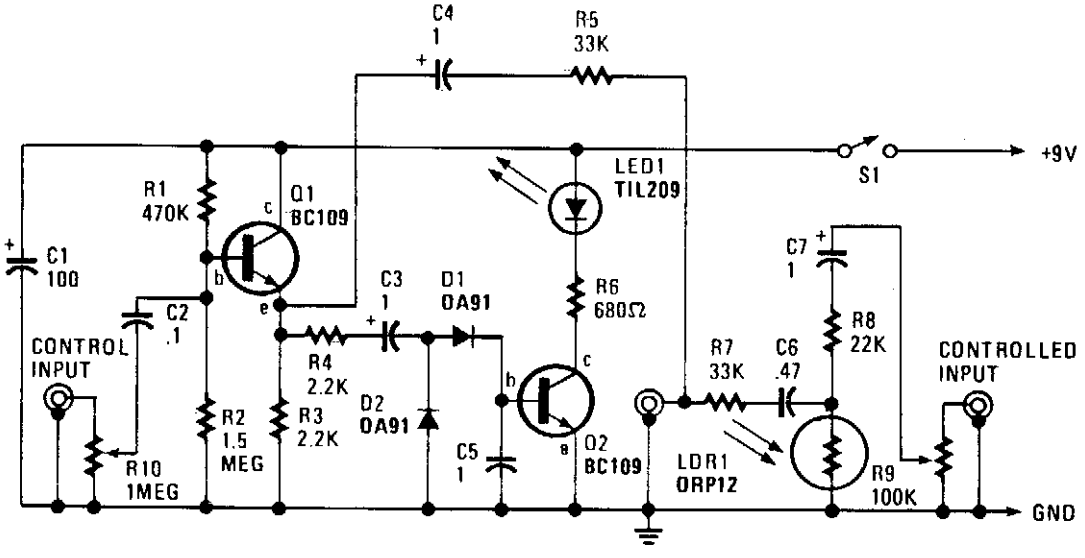
6

Audio Signal Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Auto Fade
Transistor Headphone Amplifier
Stereo Preamplifier
Audio Compressor
Micropower High-Input-Impedance 20-dB Amplifier
Stereo Preamplifier
Microphone Preamplifier
Volume, Balance, Loudness & Power Amps
Balance and Loudness Amplifier

AUTO FADE



TAB BOOKS INC.

Fig. 6-1

Circuit Notes

The automatic fader drops the level of the background music when the narration comes up. The control input goes through R10, a preset audio level control, to the input of an emitter-follower buffer stage (Q1). The buffer offers a high input impedance and makes sure that the source impedance is low enough to drive the rectifier and smoothing circuit, which consist of D1, D2, and C5. The smoothed output drives a simple LED circuit. R8 and LDR1 form an input attenuator across which the output is fed via C6 and C7 to the output jack. The output at the emitter of Q1 couples to this socket through C4 and R5. R5 and R7 are a passive mixer. With 200 mV or less at the input, there isn't sufficient voltage across C5 to make Q2 turn on. Over 200 mV, Q2 does turn on to a limit, and the LED gets power. That makes the LDR's resistance fall, and signal loss through the attenuator increases. Increase the input to 350 mV rms, and you get a signal reduction of better than 20 dB.

TRANSISTOR HEADPHONE AMPLIFIER

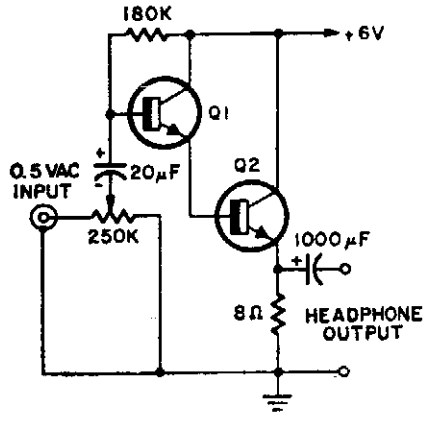
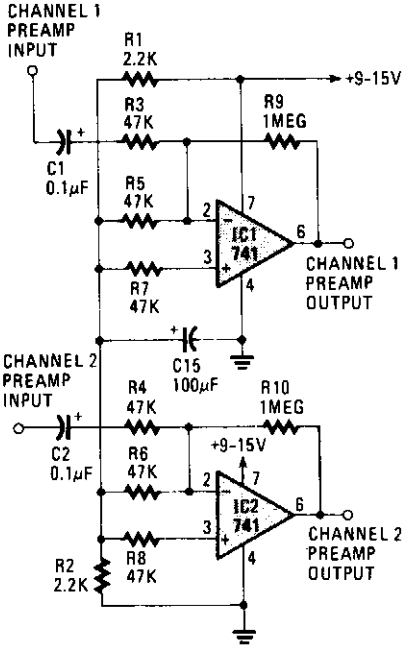


Fig. 6-2

RADIO ELECTRONICS

STEREO PREAMPLIFIER



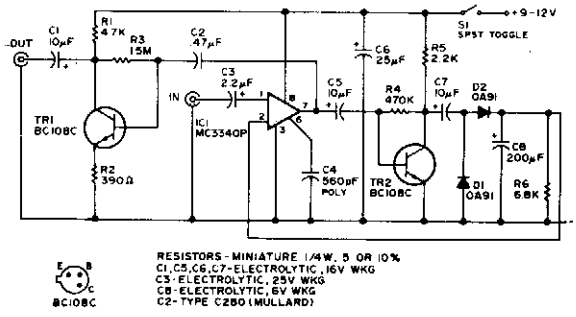
HANDS-ON ELECTRONICS

Circuit Notes

The circuit provides better than 20-dB gain in each channel. A better op-amp type will give a better noise figure and bandpass. In this circuit the roll-off is acute at 20,000 Hertz.

Fig. 6-2

AUDIO COMPRESSOR



Circuit Notes

A MC3340P is used as a variable gain amplifier. The output of TR2 is rectified and controls the gain of IC1.

73 MAGAZINE

Fig. 6-3

MICROPOWER HIGH-INPUT-IMPEDANCE 20-dB AMPLIFIER

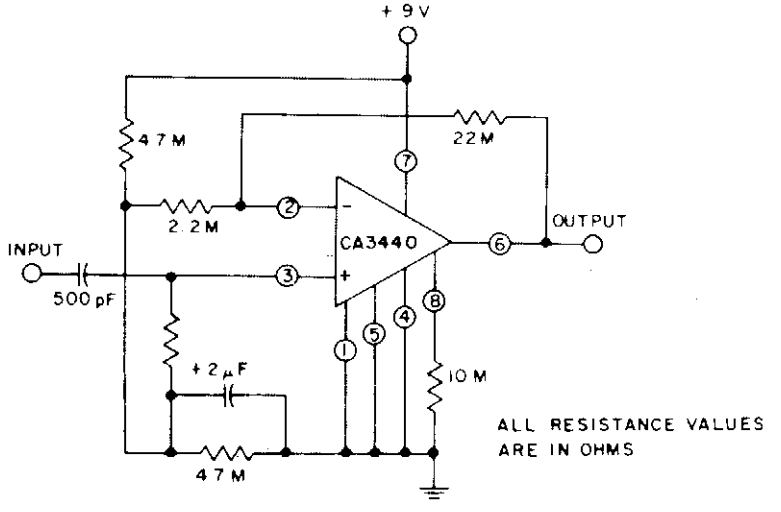


Fig. 6-4

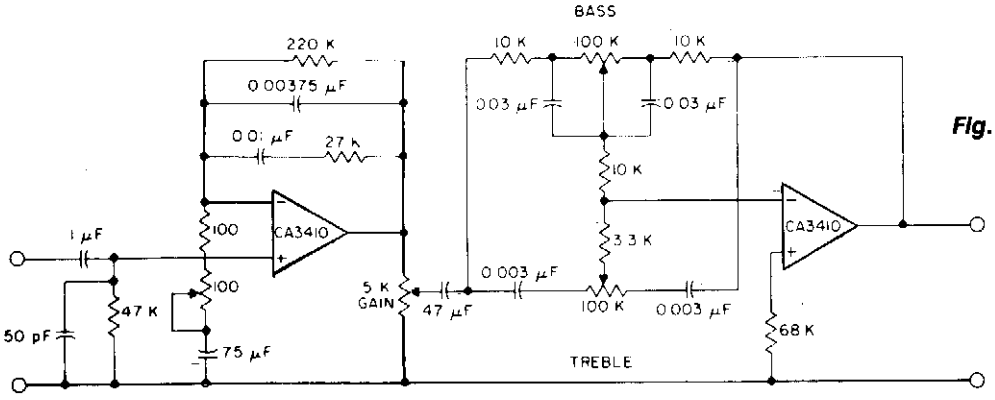
$R_{in} > 20$ M
STAND-BY POWER = 90μW 3 kHz UPPER 3 dB BANDWIDTH
GAIN = 20 db
BW: 20-Hz TO 3-KHz
SR = 0.016 V/μs

GENERAL ELECTRIC/RCA

Circuit Notes

This circuit takes advantage of low power drain, high input impedance, and the excellent frequency capability of the CA3440. Only a 500-pF input coupling capacitor is needed to achieve a 20 Hz, −3 dB low-frequency response.

STEREO PREAMPLIFIER



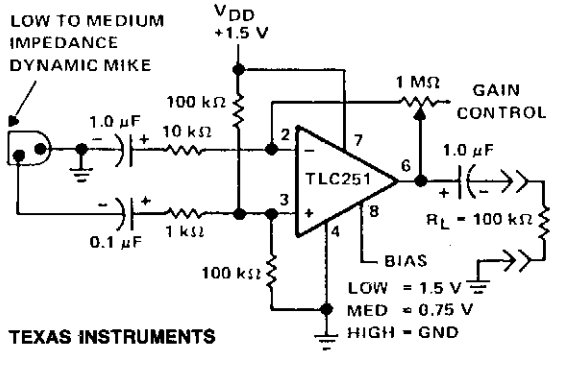
GENERAL ELECTRIC/RCA

Fig. 6-5

Circuit Notes

This circuit has RIAA playback equalization, tone controls, and adequate gain to drive a majority of commercial power amplifiers, using the CA3410 BiMOS op amp. Total harmonic distortion, when driven to provide a 6-V output, is less than 0.035% in the audio-frequency range of 150 Hz to 40 kHz. Complete stereo preamplifier consists of duplicating this circuit using the two remaining CA3410 amplifiers.

MICROPHONE PREAMPLIFIER



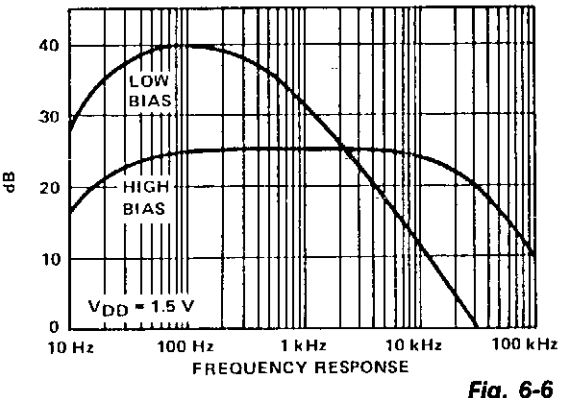
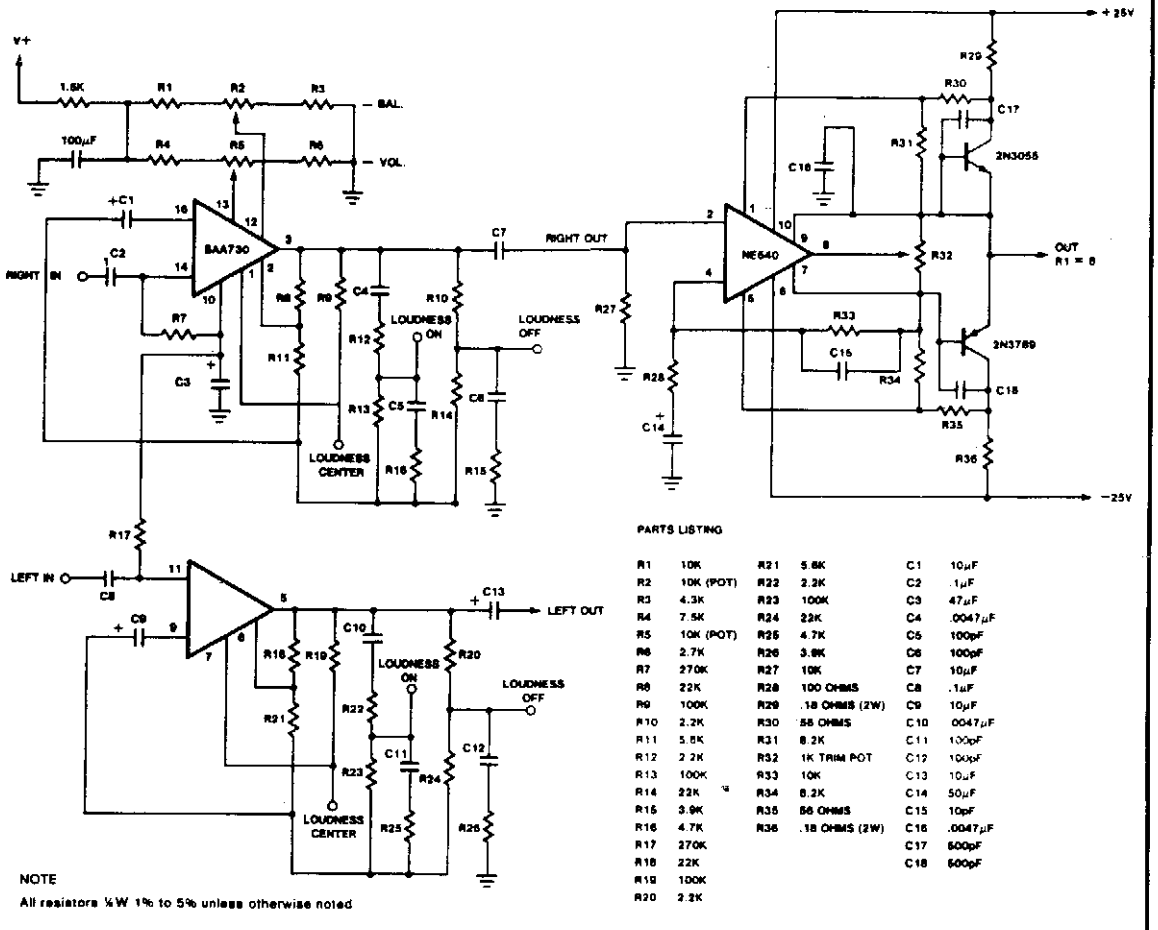


Fig. 6-6

Circuit Notes

A microphone preamplifier using a :om CMOS op amp complete with its own battery, is small enough to be put in a small mike case. The amplifier operates from a 1.5-V mercury cell battery at low supply currents. This preamplifier will operate at very low power levels and maintain a reasonable frequency response as well. The TLC251 operated in the low bias mode (operating at 1.5 V) draws a supply current of only 10 μA and has a −3 dB frequency response of 27 Hz to 4.8 kHz. With pin 8 grounded, which is designated as the high bias condition, the upper limit increases to 25 kHz. Supply current is only 30 μA under those conditions.

VOLUME, BALANCE, LOUDNESS & POWER AMPS



NOTE
All resistors ¼W 1% to 5% unless otherwise noted

PARTS LISTING

R1	10K	R21	5.6K	C1	10μF
R2	10K (POT)	R22	2.2K	C2	.1μF
R3	4.3K	R23	100K	C3	47μF
R4	7.5K	R24	22K	C4	.0047μF
R5	10K (POT)	R25	4.7K	C5	100pF
R6	2.7K	R26	3.9K	C6	100pF
R7	270K	R27	10K	C7	10μF
R8	22K	R28	100 OHMS	C8	.1μF
R9	100K	R29	.18 OHMS (2W)	C9	10μF
R10	2.2K	R30	56 OHMS	C10	.0047μF
R11	5.6K	R31	8.2K	C11	100pF
R12	2.2K	R32	1K TRIM POT	C12	100pF
R13	100K	R33	10K	C13	10μF
R14	22K	R34	8.2K	C14	50μF
R15	3.9K	R35	56 OHMS	C15	10pF
R16	4.7K	R36	.18 OHMS (2W)	C16	.0047μF
R17	270K			C17	500pF
R18	22K			C18	500pF
R19	100K				
R20	2.2K				

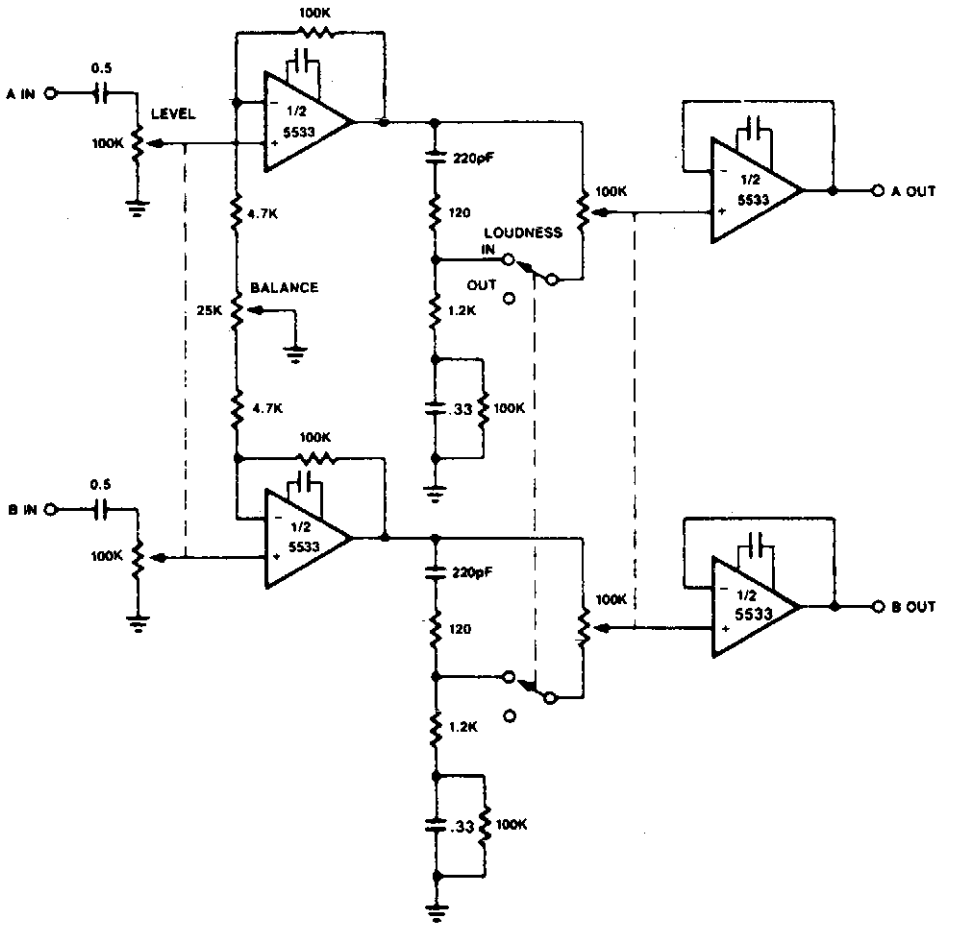
SIGNETICS

Fig. 6-7

Circuit Notes

This circuit should prove suitable as a design example for audio sound application.

BALANCE AND LOUDNESS AMPLIFIER



NOTE:
All resistor values are in ohms.

SIGNETICS

Fig. 6-7

Circuit Notes

The circuit shows a combination of balance and loudness controls. Due to the non-linearity of the human hearing system, the low frequencies must be boosted at low listening levels. Balance, level, and loudness controls provide all the listening controls to produce the desired music response.

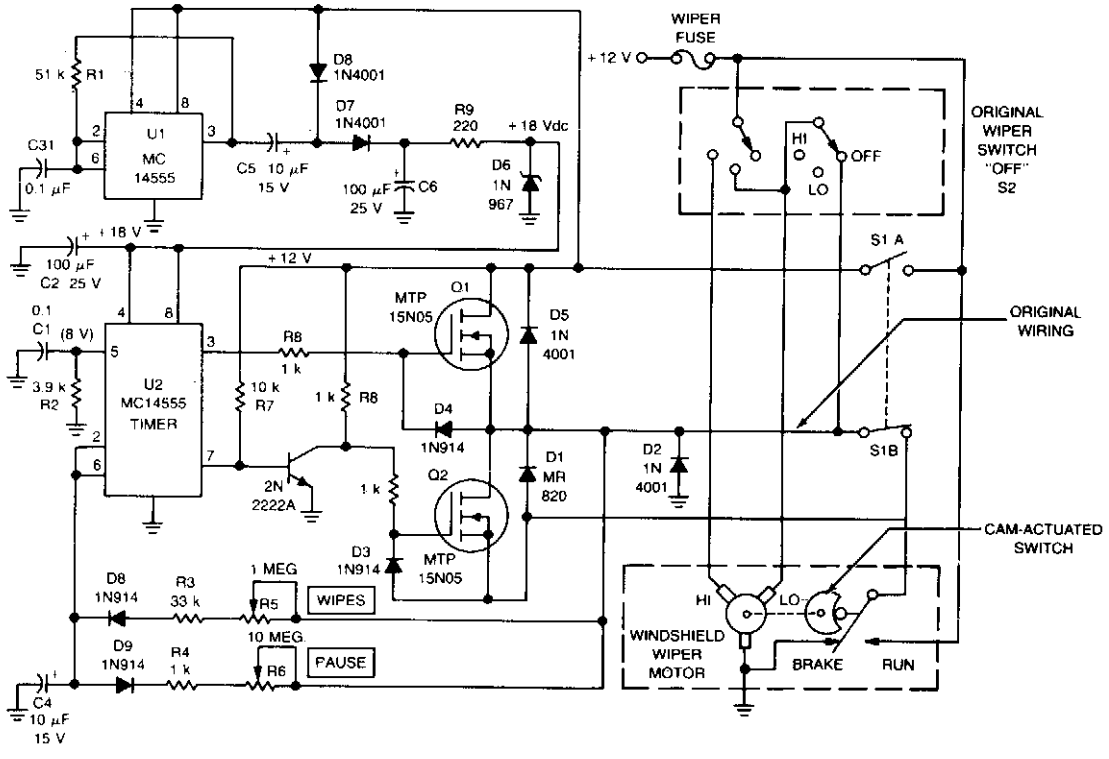
7

Automotive Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Intermittent Windshield Wiper with Dynamic Braking
Immobilizer
Automotive Exhaust Emissions Analyzer
Glow Plug Driver
Garage Stop Light
Bar-Graph Voltmeter
Delayed-Action Windshield Wiper Control
Slow-Sweep Wiper Control
Automotive Lights On Warning
PTC Thermistor Automotive Temperature Indicator
Road Ice Alarm
Headlight Dimmer
Ice Formation Alarm
Delay Circuits for Headlights
Ignition Timing Light
Digi-Tach
Car-Wiper Control
Automatic Headlight Dimmer

INTERMITTENT WINDSHIELD WIPER WITH DYNAMIC BRAKING



MOTOROLA

Fig. 7-1

Circuit Notes

The circuit provides a delayed windshield wiping, and dynamic braking of wiper blades when they reach the rest position. This prevents the blades from overshooting, which might cause them to stop at a point where they interfere with the drivers' vision.

With the original wiper switch off, switch S1A turns on the delay circuit and S1B disconnects the original automotive wiring. When S1 is turned off, the original wiring controls the system and the delay circuit is bypassed.

Turning S1 on applies the +12-V battery to U1 which is a voltage doubler that produces +18 V. This higher voltage supply is necessary to ensure reliable turn on of Q1 by multivibrator U2. This arrangement provides about +18 V to the gate of Q1, whose source is +12 V minus the V_{DS} drop of Q1.

Q1 remains on for a time determined by the WIPES potentiometer. The interval between wipes is controlled by the PAUSE control. When C1 drops below +4 V, U2 fires, turning Q1 on and restarting the cycle.

IMMOBILIZER

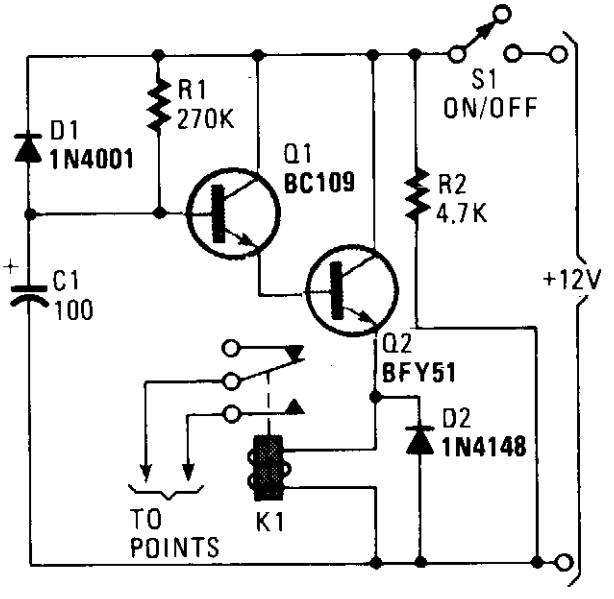


Fig. 7-2

HANDS-ON ELECTRONICS

Circuit Notes

A flip of S1 puts the circuit into action. Power for the circuit is picked up from the ignition switch, and the circuit receives no power until the ignition switch is closed. When power is turned on, capacitor C1 is not charged and the emitter-follower Darlington pair (formed by Q1 and Q2) are cutoff, thus no power is applied to the relay (K1), which serves as Q1's emitter load. The relay's normally-open contacts are connected across the vehicle's points. (At this time, the relay contacts are open and have no effect on the ignition system). C1 charges by way of R1, causing the voltage at the base of Q1 to rise steadily. That creates a similar rise in the voltage at the emitter of Q2. A Darlington pair is used to provide a high input-impedance, buffer stage so that the voltage across C2 is free to rise almost to the full supply potential. Loading effects do not limit the charge potential to just a few volts. Eventually, the voltage applied to the relay becomes sufficient to activate it. The contacts close and short out the points. The ignition system now doesn't act properly and the vehicle is disabled. If the ignition is switched off, power is removed from the circuit and diode D1, which was previously reverse-biased, is now forward biased by the charge on C1. D1 allows C1 to rapidly discharge through R2 (and any other dc paths across the supply lines). The circuit is ready to operate when the ignition is again turned on. The engine will operate, but not for very long. The values of R1 and C1 provides a delay of about 25 to 30 seconds. Increase R1's value to provide a longer delay.

AUTOMOTIVE EXHAUST EMISSIONS ANALYZER

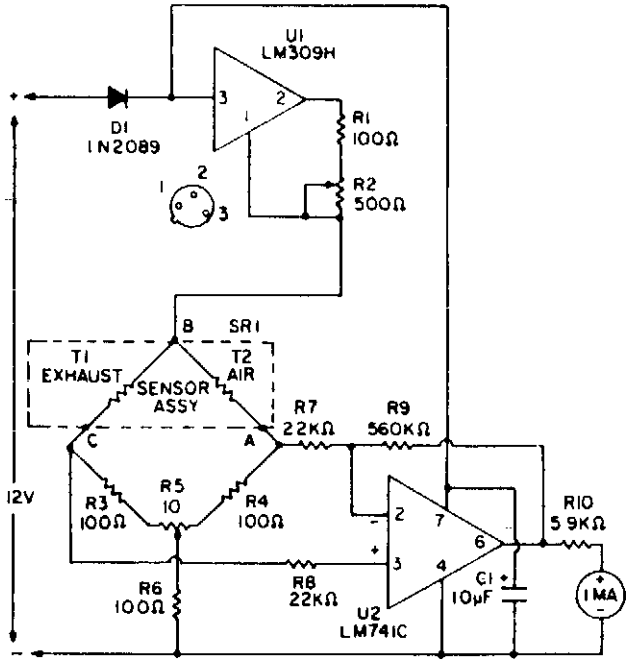


Fig. 7-3

C1—10-uF, 25 volt electrolytic capacitor
D1—Silicon diode, general purpose, 1N2069 or equivalent
M1—0.1 ma ammeter
R1, R3, R4, R6—100-ohm 10% ¼ watt resistor (All resistors are 10%, ¼ watt)
R2—500-ohm potentiometer, PC mount
R5—10-ohm potentiometer, front panel mount
R7, R8—22,000-ohm resistor
R9—570,000-ohm resistor
R10—5,900-ohm resistor
SR1—thermister sensor assembly, part number 100-1648 from Heathkit model CI-1080, Heath Company Benton Harbor, MI, 49022
U1—LM309H 5-volt regulator integrated circuit
U2—LM741TC op amp intergrated circuit

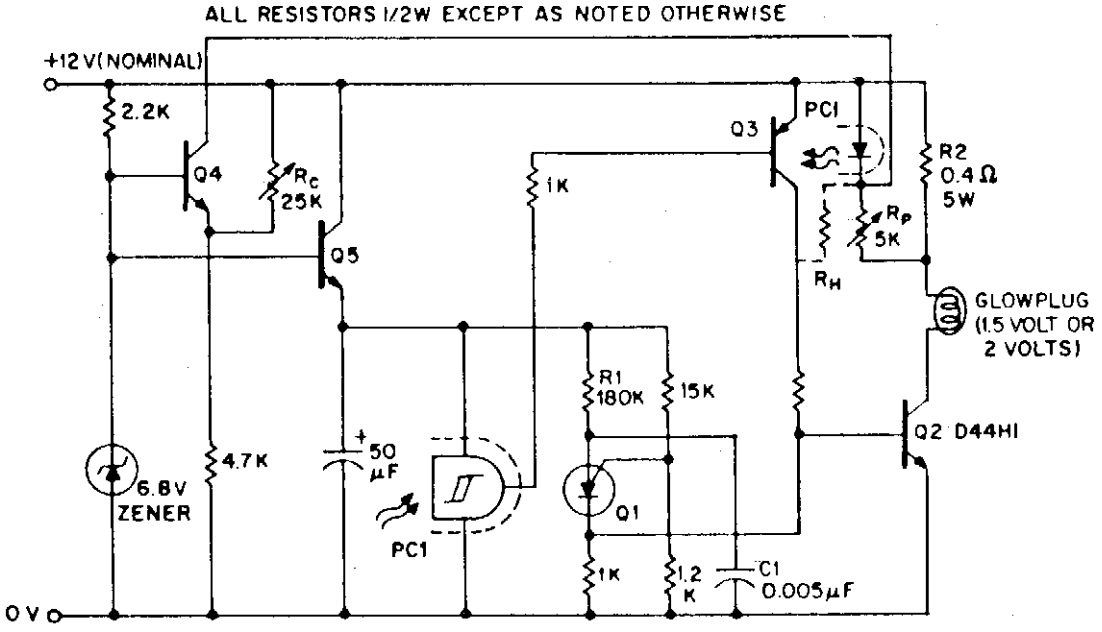
Misc.—PC Board, screws, cabinet, solder, hardware, 10-foot plastic tube, etc.

TAB BOOKS, INC.

Circuit Notes

A bridge circuit contains two 100-ohm resistors (R3 and R4), and two thermistors (T1 and T2). At room temperature the resistance of T1 and T2 is about 2000 ohms. When they are each heated to 150°C by a 10 mA current, the resistance value decreases to 100 ohms. Thus, the four elements comprise a bridge circuit. A characteristic of CO is that it conducts heat away from a thermistor at a different rate than air. One thermistor, T1, is exposed to the automobile exhaust while the other, T2, is isolated in a pure air environment. The difference in thermal conduction unbalances the bridge. A voltage difference is caused between points A and C. A differential amplifier, U1, amplifies this difference and drives the meter with sufficient current to read out the percentage of CO and the air-fuel ratio. A front panel balance control, R5, balances the bridge and calibrates the instrument. Calibration is performed when both thermistors are exposed to the outside air.

GLOW PLUG DRIVER



R_C – SUPPLY VOLTAGE COMPENSATION TRIM. PC1 – H11L2.
R_P – PLUG TEMPERATURE ADJUST. Q3 – D29E2
Q1 – 2N6028 Q4, Q5 – D33D1

GENERAL ELECTRIC

Fig. 7-4

Circuit Notes

Model airplanes, boats, and cars use glow plug ignitions for their miniature (0.8cc to 15cc) internal combustion engines. Such engines dispense with the heavy on-board batteries, H.T. coil, and "condenser" required for conventional spark ignition, while simultaneously developing much higher RPM (hence power) than the compression ignition (diesel) motors. The heart of a glow plug is a platinum alloy coil heated to incandescence for engine starting by an external battery, either 1.5 volts or 2 volts. Supplementing this battery, a second 12-volt power supply is frequently required for the engine starter, together with a third 6 volt type for the electrical fuel pump.

Rather than being burdened by all these multiple energy sources, the model builder would prefer to carry (and buy) a single 12-volt battery, deriving the lower voltages from this by use of suitable electronic step-down transformers (choppers). The glow driver illustrated does this and offers the additional benefit of (through negative feedback) maintaining constant plug temperature independent of engine flooding, or battery voltage while the starter is cranking.

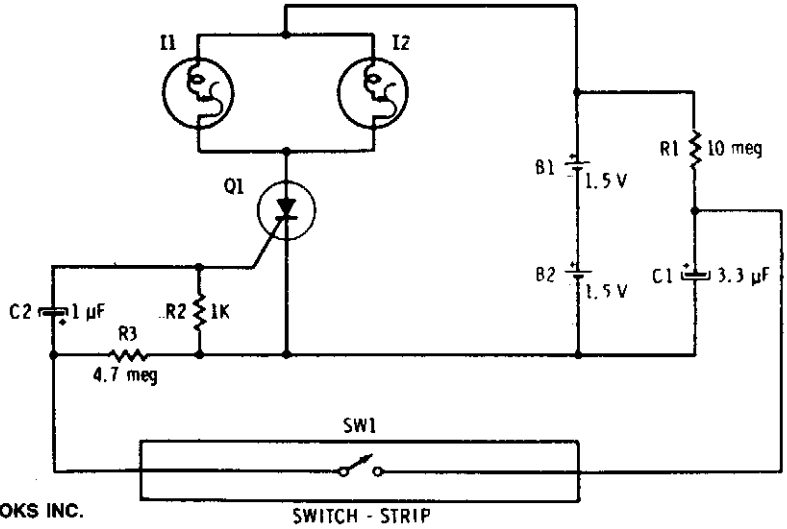
In this circuit, the PUT relaxation oscillator Q1 turns on the output chopper transistor Q2 at a fixed repetition rate determined by R1 and C1. Current then flows through the glow plug and the parallel combination of the current sense resistor R2 and the LED associated with the H11L Schmitt trigger. With the plug cold (low resistance), current

is high, the H11L is biased "on", and Q3 conducts to sustain base drive to Q2. Once the plug has attained optimum operating temperature, which can be monitored by its ohmic resistance, the H11L is programmed (via R_p) to switch off, removing base drive from Q3 and Q2.

However, since the H11L senses glow plug current, not resistance, this is only valid if supply voltage is constant, which is not always the case. Transistor Q4 provides suitable compensation in this case; if battery voltage falls (during cold cranking, for instance), the collector current of Q4 rises, causing additional current to flow through the LED, thus delaying the switch-off point for a given plug current. The circuit holds plug temperature relatively constant, with the plug either completely dry or thoroughly "wet", over an input voltage range of 8 to 16 volts. A similar configuration can be employed to maintain constant temperature for a full size truck diesel glow plug (28-volts supply, 12-volts glow plug); in this case, since plug temperature excursions are not so great, a hysteresis expansion resistor R_H may be required.

Fig. 7-4 Continued

GARAGE STOP LIGHT



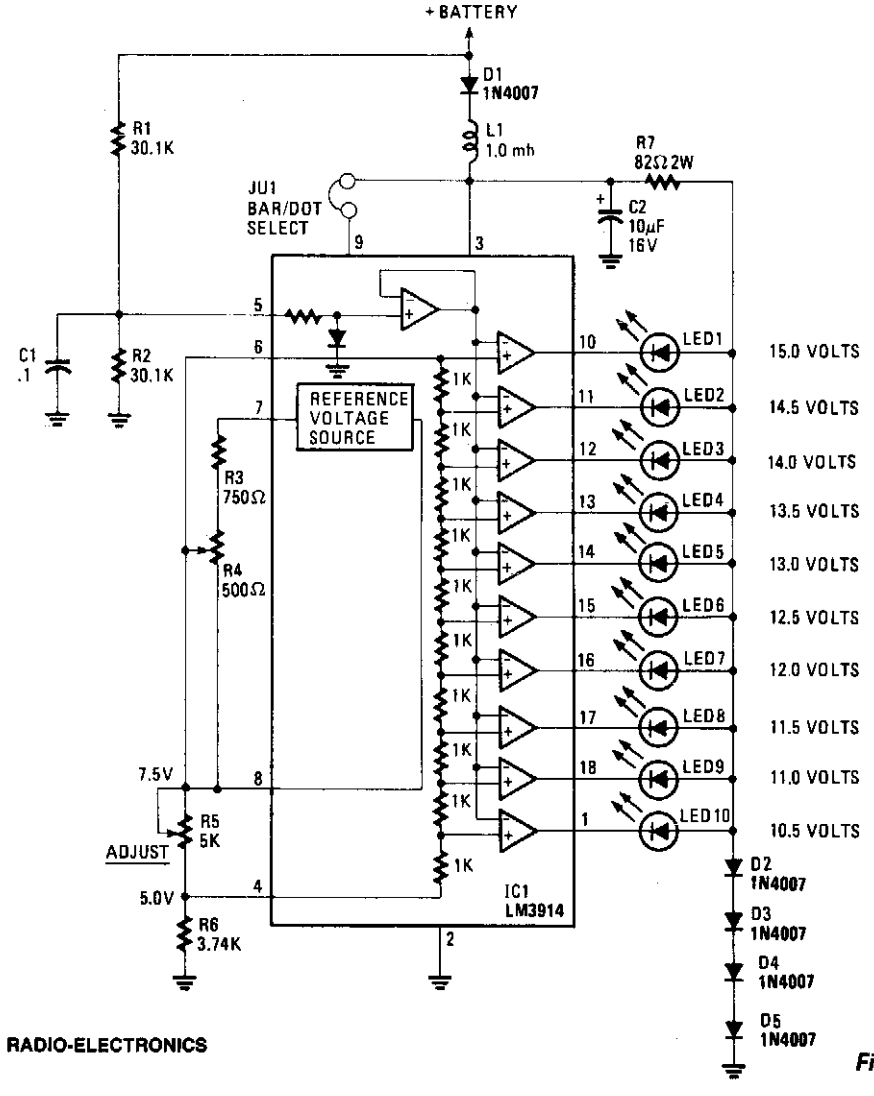
TAB BOOKS INC.

Fig. 7-5

Circuit Notes

Capacitor C1 is permanently connected across the 3-volt supply through 10 megohm resistor R1. The capacitor charges (relatively slowly) to 3 volts. The instant switch SW1 is closed, it connects the charged capacitor (C1) in series with C2 and R2. Capacitor C2 starts to charge, placing a positive-going voltage on the gate of the SCR and causing it to turn on. The two parallel-connected "self-flashing" bulbs I1 and I2 turn on. They flash and turn off the SCR and the circuit is off until car is driven off the switch and C1 can recharge.

BAR-GRAPH VOLTMETER



RADIO-ELECTRONICS

Fig. 7-6

Circuit Notes

This display uses ten LED's to display a voltage range from 10.5 to 15 volts. Each LED represents a 0.5-volt step in voltage. The heart of the circuit is the LM-3914 dot/bar display driver. Trimmer potentiometer R5 is adjusted so that 7.5 volts is applied to the top side of the divider. Resistor R7 and diodes D2 through D5 clamp the voltage applied to the LED's to about 3 volts. A lowpass filter made up of L1 and C2 guards against voltage spikes. Diode D1 is used to protect against reverse voltage in case the voltmeter is hooked up backward.

DELAYED-ACTION WINDSHIELD WIPER CONTROL

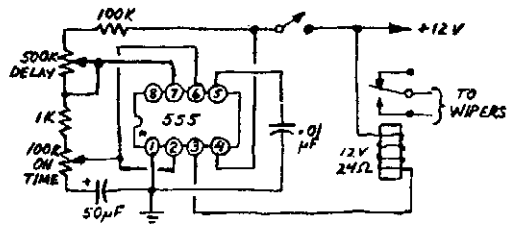
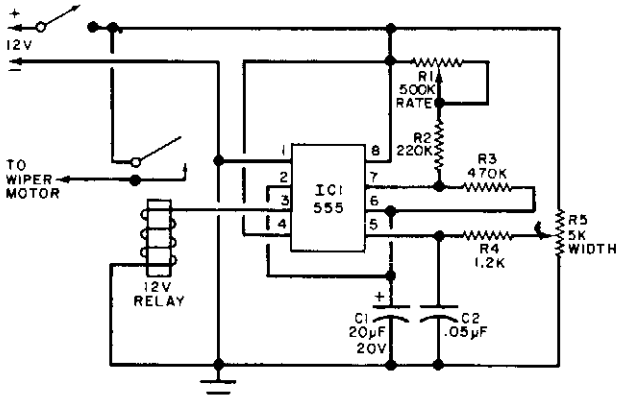


Fig. 7-7

POPULAR ELECTRONICS

SLOW-SWEEP WIPER CONTROL



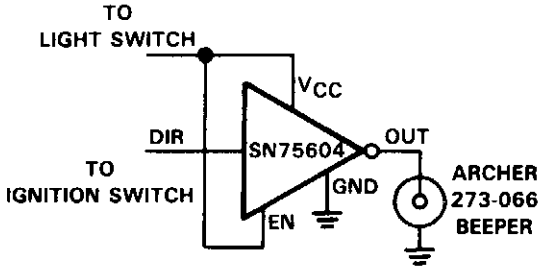
POPULAR ELECTRONICS

Circuit Notes

The relay which applies power to the wiper motor is actuated at periodic intervals by the timer circuit, closing the wiper motor contacts. Potentiometer R1 serves as the pulse rate control and potentiometer R5 as the pulse width control. These two controls should be adjusted for optimum performance after the unit is installed in a car.

Fig. 7-8

AUTOMOTIVE LIGHTS ON WARNING



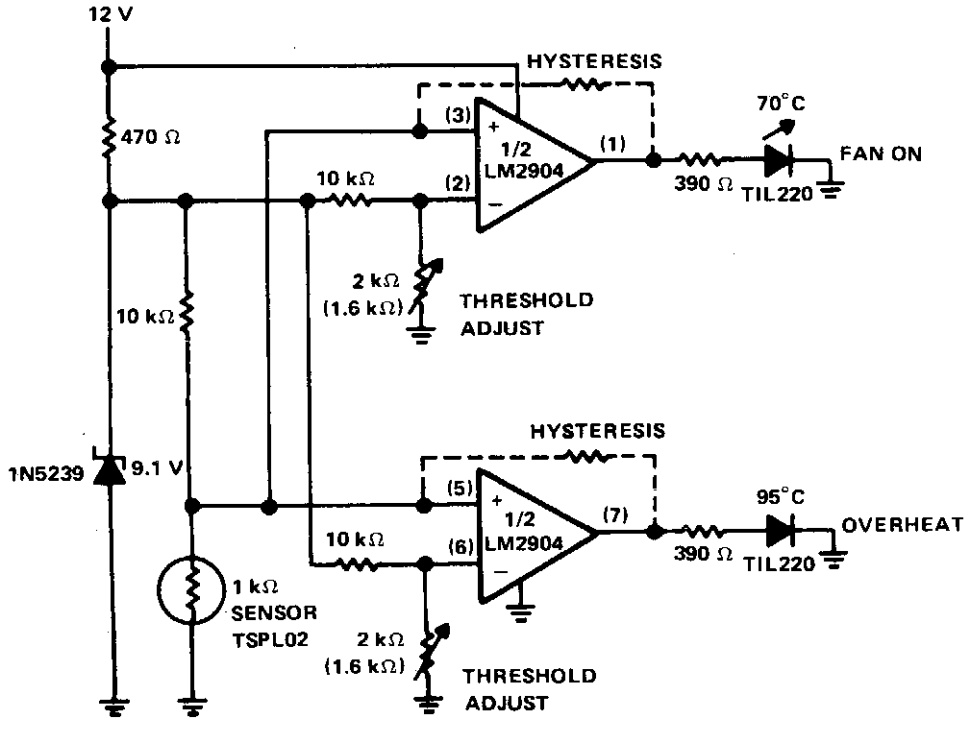
TEXAS INSTRUMENTS

Fig. 7-9

Circuit Notes

The SN75604, with input control logic but requiring only one supply rail, can be used in the "lights on" sensor and alarm driver. The device V_{CC} and enable inputs are connected to a voltage lead from the light switch. The direction control input is connected to a lead from the ignition switch. Only operation of the lights without the ignition will result in the alarm sounding. The beeper used in this application is an Archer 273-066 that will operate from 3 V to 28 V. At a typical 12 V level, it will produce a pulsating tone of about 95 dB at 30 cm. The alarm "on" current is about 12 mA when operating from a 12 V supply.

PTC THERMISTOR AUTOMOTIVE TEMPERATURE INDICATOR



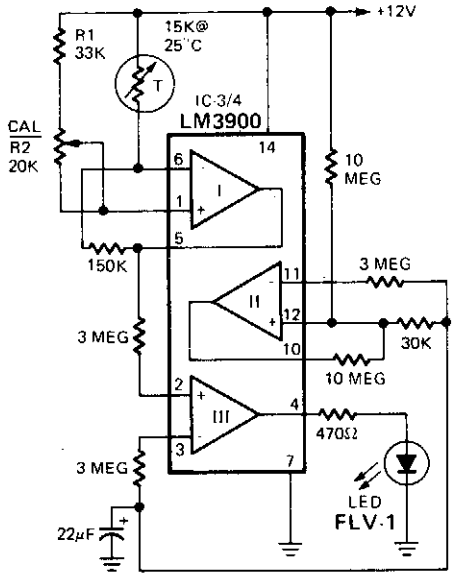
TEXAS INSTRUMENTS

Fig. 7-10

Circuit Notes

The circuit is used to indicate two different water temperature trip points by turning on LEDs when the temperatures are reached. The circuit is constructed around the LM2904 dual operational amplifier powered from the 12 V auto system. The thermistor is in series with a 10 kΩ resistor from ground to the positive 9.1 V point. The top of the thermistor is tied to both non-inverting inputs of the LM2904. The voltage at these inputs will change as the thermistor resistance changes with temperature. Each inverting input on the LM2904 has a reference, or threshold trip point, set by a 10 kΩ resistor and a 2 kΩ potentiometer in series across the 9.1 V regulated voltage. When this threshold is exceeded on the non-inverting input of LM2904, the TIL220 LED lights. The two trip points can be recalibrated or set to trip at different temperatures by adjusting the 2 kΩ potentiometer in each section. In addition to being used as warning lights as shown here, circuits can be added to turn on the fan motor or activate a relay.

ROAD ICE ALARM



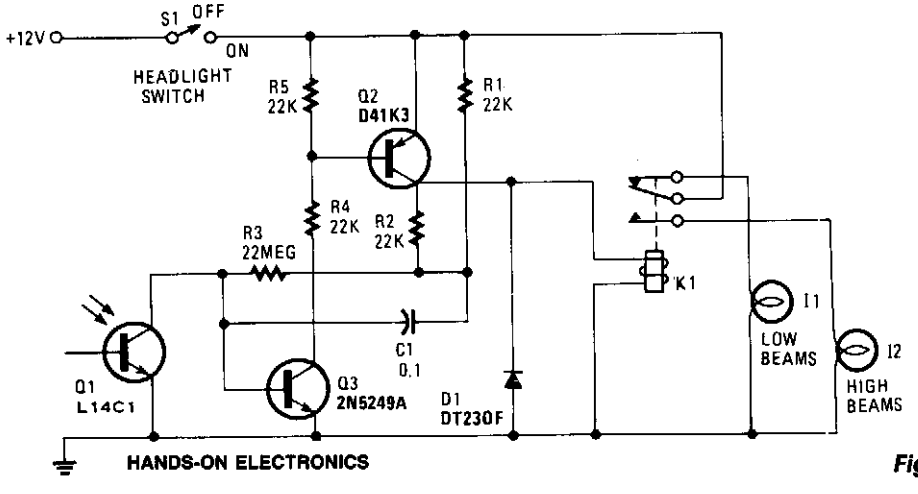
RADIO-ELECTRONICS

Fig. 7-11

Circuit Notes

The circuit uses a thermistor and three sections of a LM3900 quad op amp IC. When the temperature drops to 36°F the LED indicator flashes about once each second. The flashing rate increases as temperature drops to 32°F when the LED remains on. Amplifier I compares the thermistor's resistance to the resistance of the standard network connected to its noninverting input. Its output—fed to the noninverting input of op amp III—varies with temperature. Op amp II is a free-running multivibrator feeding a pulse signal of about 1 Hz to the inverting input of op amp III. This amplifier compares the outputs of op amps I and II and turns on the LED when the multivibrator's output level drops below op amp I. The monitor is calibrated by placing the thermistor in a mixture of crushed ice and water and adjusting the 20 kΩ pot so the LED stays on.

HEADLIGHT DIMMER



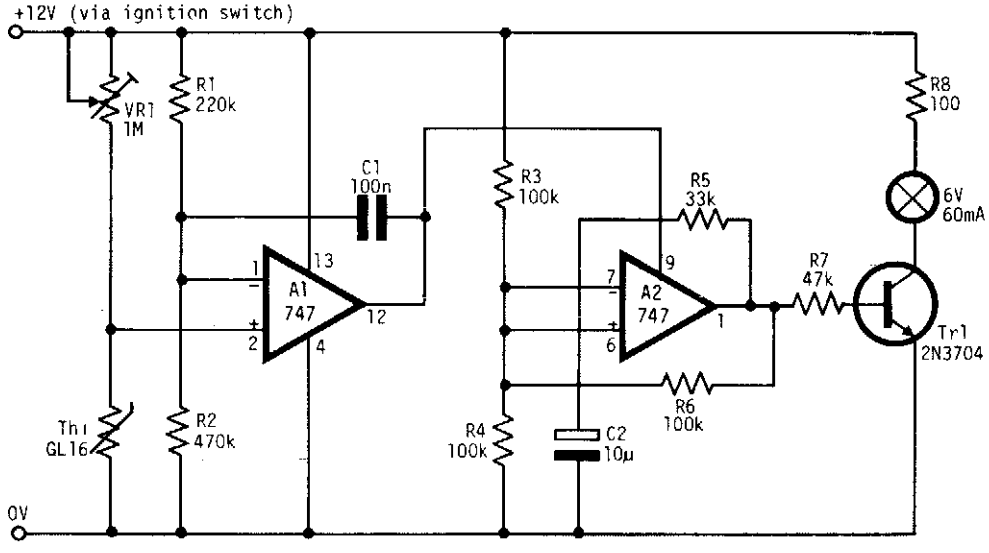
HANDS-ON ELECTRONICS

Fig. 7-12

Circuit Notes

When the lights of an on-coming car are sensed by photo-transistor Q1, things get going. Sensitivity is set by the 22-megohm resistor, R5, to about half a foot-candle. The relay used has a 12-volt, 0.3A coil. The L14C1 is complete with a lens that has a diameter of one inch for a 10° viewing angle.

ICE FORMATION ALARM



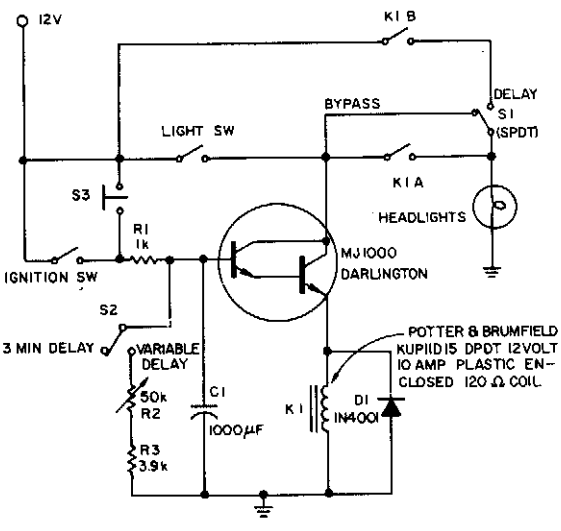
ELECTRONIC ENGINEERING

Fig. 7-13

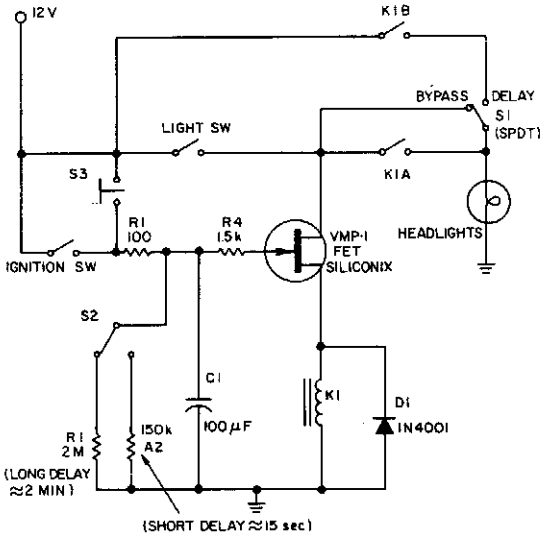
Circuit Notes

The circuit warns car drivers when the air temperature close to the ground approaches 0°C, thereby indicating possible formation of ice on the road surface. Op amp A1 is wired as a voltage level sensor. Op amp A2 is wired as an astable multivibrator which, by means of current buffer Tr1, flashes a filament lamp at about 1 Hz. As air temperature falls, a point is reached when the voltage at pin 2 just rises above the voltage at pin 1. The output of A1 is immediately driven into positive saturation, since it is operated open-loop. This positive output voltage powers A2 through its V + connection on pin 9, starting the oscillator. The thermistor is a glass bead type with a resistance of about 20 MΩ at 20°C. VR1 is adjusted so that the lamp starts flashing when the air temperature is 1 to 2°C.

DELAY CIRCUITS FOR HEADLIGHTS



1. **Automobile headlights may be kept on** up to 3 minutes after you leave the car with this Darlington time-delay circuit.



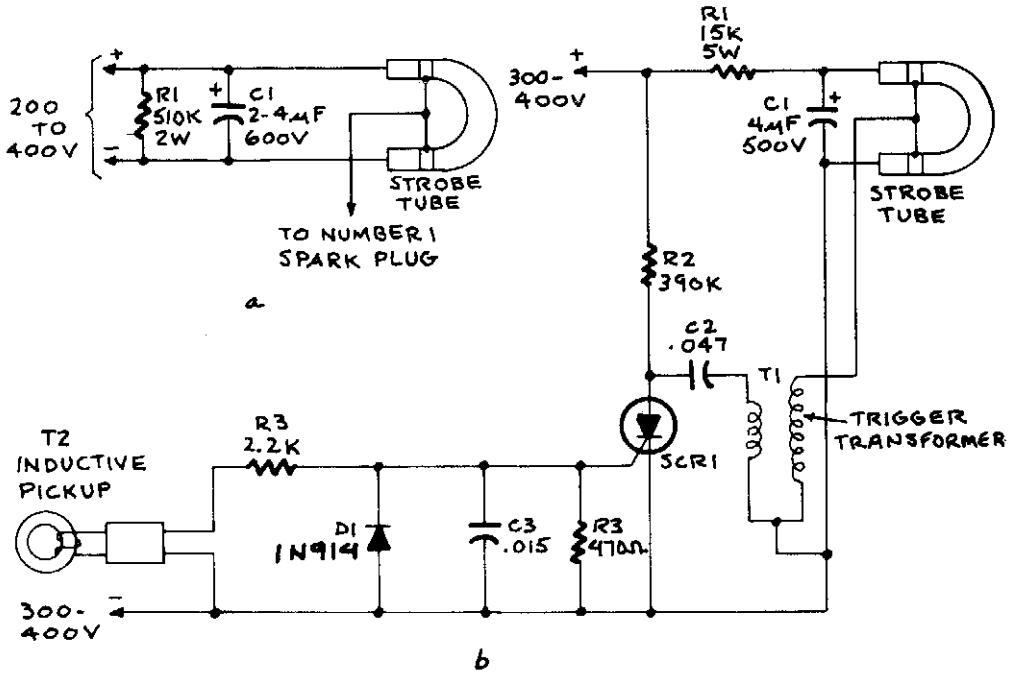
2. **A FET version of the delay circuit** allows the use of a smaller timing capacitor, C_1, for a given delay, and almost instantaneous reset with S_3; the Darlington circuit needs almost 2 s.

ELECTRONIC DESIGN ***Fig. 7-14***

Circuit Notes

This circuit keeps an automobile's headlights on temporarily. It also will turn the lights off, even if you forget to flip the light switch. The circuit's shut-off delay is actuated only after both the ignition and light switches have been on, and only if the ignition switch is turned off first. If the light switch is turned off first, no delay results. Parking and brake-light operation is not affected. The maximum time out can be up to 3 minutes in part 1 and hours with the circuit in part 2, depending on the relay selected and the value of R2. A switch S2 can be used to permit selection of either a short or long delay. Momentary switch S3 can restart circuit timing before the time-out is completed. A bypass switch, S1 removes the delay action.

IGNITION TIMING LIGHT



RADIO-ELECTRONICS

Fig. 7-15

Circuit Notes

Figure *A* shows the circuit of a direct-trigger timing light. The trigger voltage is taken from the car's ignition circuit by a direct connection to a spark plug. A circuit using an inductive pickup is shown in Fig. *B*. A trigger transformer is used to develop the high-voltage pulse for triggering. The triggering circuit consists of T1, C1, SCR1, inductive pickup coil T2, and the waveshaping components in the SCR's gate circuit.

When the spark plug fires, it induces a pulse in pickup coil T2 that triggers the SCR gate. The SCR fires and discharges C2 through the primary of T1. The secondary of T1 feeds a high-voltage pulse to the trigger electrode of the flash tube. That pulse causes the gas—usually neon or xenon—to ionize. The ionized gas provides a low-resistance path for C1 to discharge, thereby creating a brilliant flash of light.

Resistor R1 limits current from the supply as the tube fires. When C1 is fully discharged the strobe tube cuts off and returns to its "high-resistance" state. The current through R2 is not enough to sustain conduction through SCR1, so it cuts off and remains off until it is re-triggered by a gate pulse.

DIGI-TACH

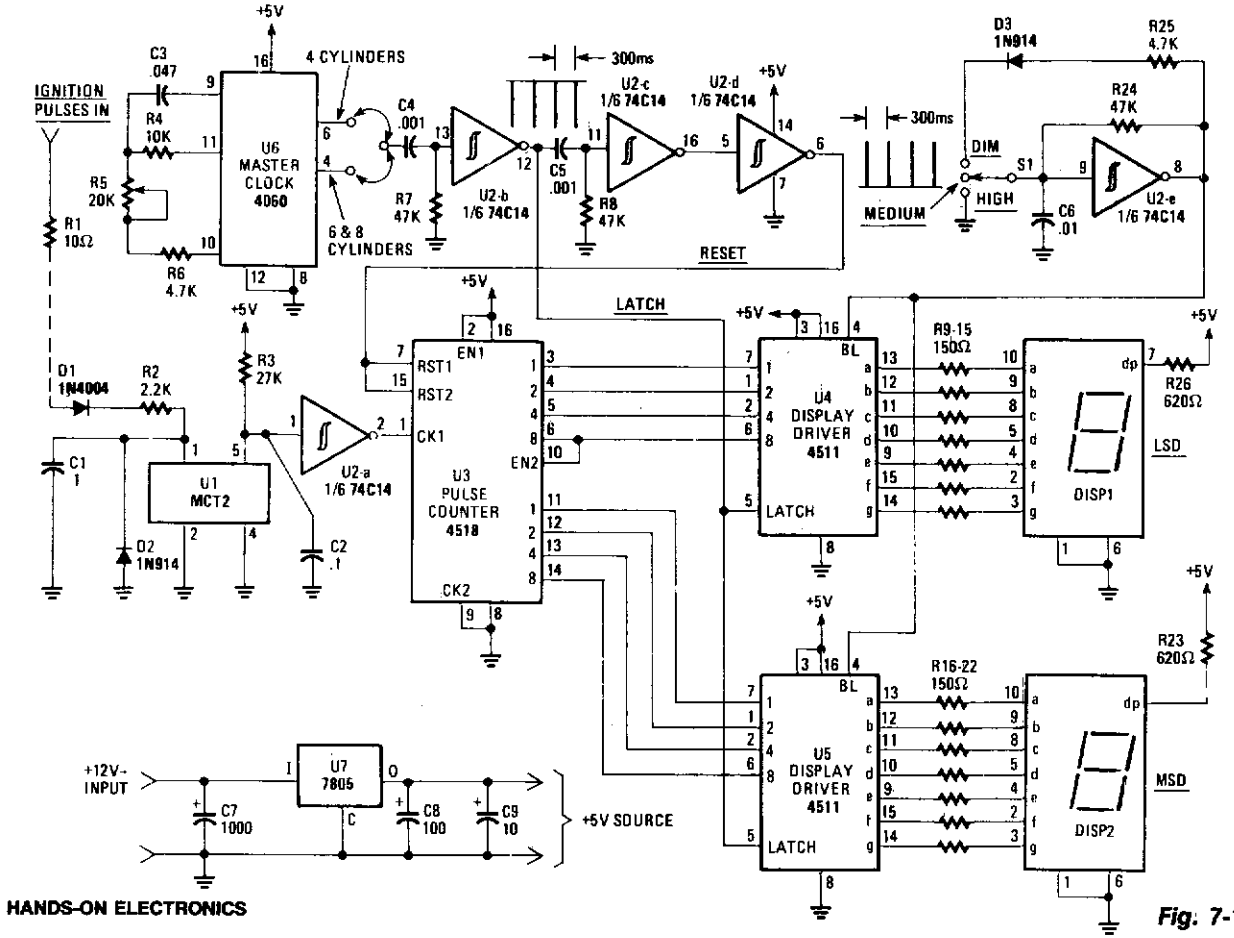
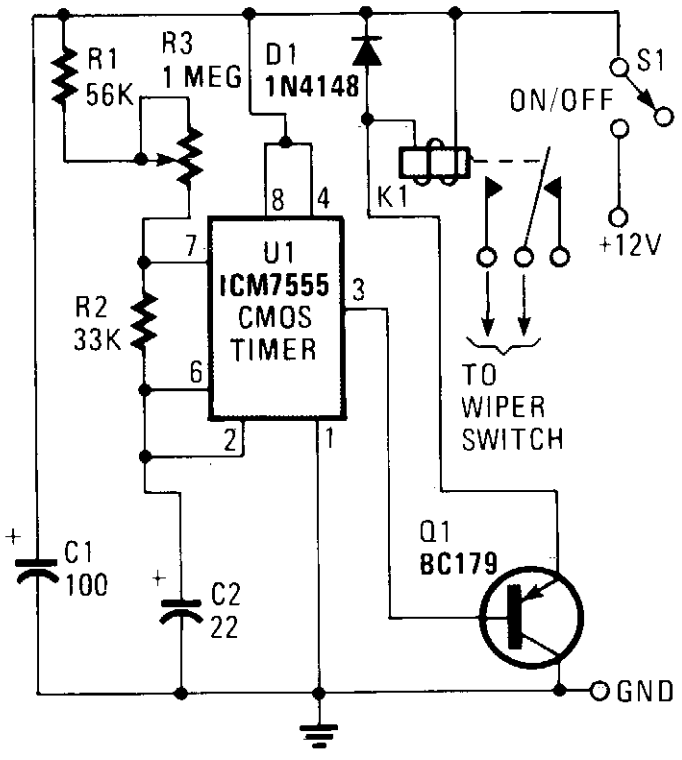


Fig. 7-16

Circuit Notes

The Digi-Tach contains a master-clock circuits (U6), latch and reset pulse generators (U2-b-U2-d), input signal conditioner (U1, U2-a), pulse counter (U3), display and display drivers (DIS1, DIS2, U4, and U5), and a voltage regulator (U7). As an added feature, Digi-Tach contains a dimmer circuit (U2-e).

CAR-WIPER CONTROL

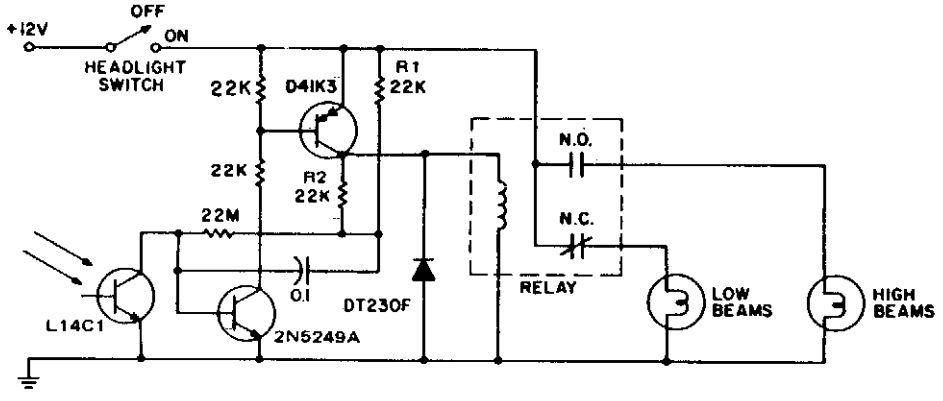


HANDS-ON ELECTRONICS *Fig. 7-17*

Circuit Notes

U1 is configured to operate in the standard astable mode, providing a form of relaxation oscillator. When power is applied, C2 initially charges through R1, R2 and R3 to two-thirds of the supply voltage. At that point, U1 senses that its threshold voltage at pin 6 has been reached, and triggers the timer, causing its output at pin 3 to go high. That high, applied to the base of Q1, keeps the transistor in the off state. Now C2 begins to discharge through R2 to pin 7 of U1. When C2 has discharged to about one-third of the supply voltage, U1 is toggled back to its original state. C2 starts to charge again, as pin 3 of U1 goes low. The low at pin 3 causes Q1—which serves as an emitter-follower buffer stage—to turn on, allowing current to flow through the coil of relay K1. That, in turn, causes K1's contacts to close, applying power to the wipers. The charge time of capacitor C2 is determined by the setting of potentiometer R3. Capacitor C2 should be a tantalum type, and actually, almost any 12-volt coil relay with sufficiently heavy contacts should serve well.

AUTOMATIC HEADLIGHT DIMMER



RELAY: 12V, 0.3A COIL: 20A, FORM C, CONTACTS OR SOLID-STATE SWITCHING OF 16A STEADY-STATE 150A COLD FILAMENT SURGE, RATING.

LENS: MINIMUM 1" DIAMETER, POSITIONED FOR ABOUT 10° VIEW ANGLE.

GENERAL ELECTRIC

Fig. 7-18

Circuit Notes

This circuit switches car headlights to the low beam state when it senses the lights of an on-coming car. The received light is very low level and highly directional, indicating the use of a lens with the detector. A relatively large amount of hysteresis is built into the circuit to prevent "flashing lights." Sensitivity is set by the 22 megohm resistor to about 0.5 ft. candle at the transistor (0.01 at the lens), while hysteresis is determined by the R1, R2 resistor voltage divider, parallel to the D41K3 collector emitter, which drives the 22 megohm resistor; maximum switching rate is limited by the 0.1 μF capacitor to 15/minute.

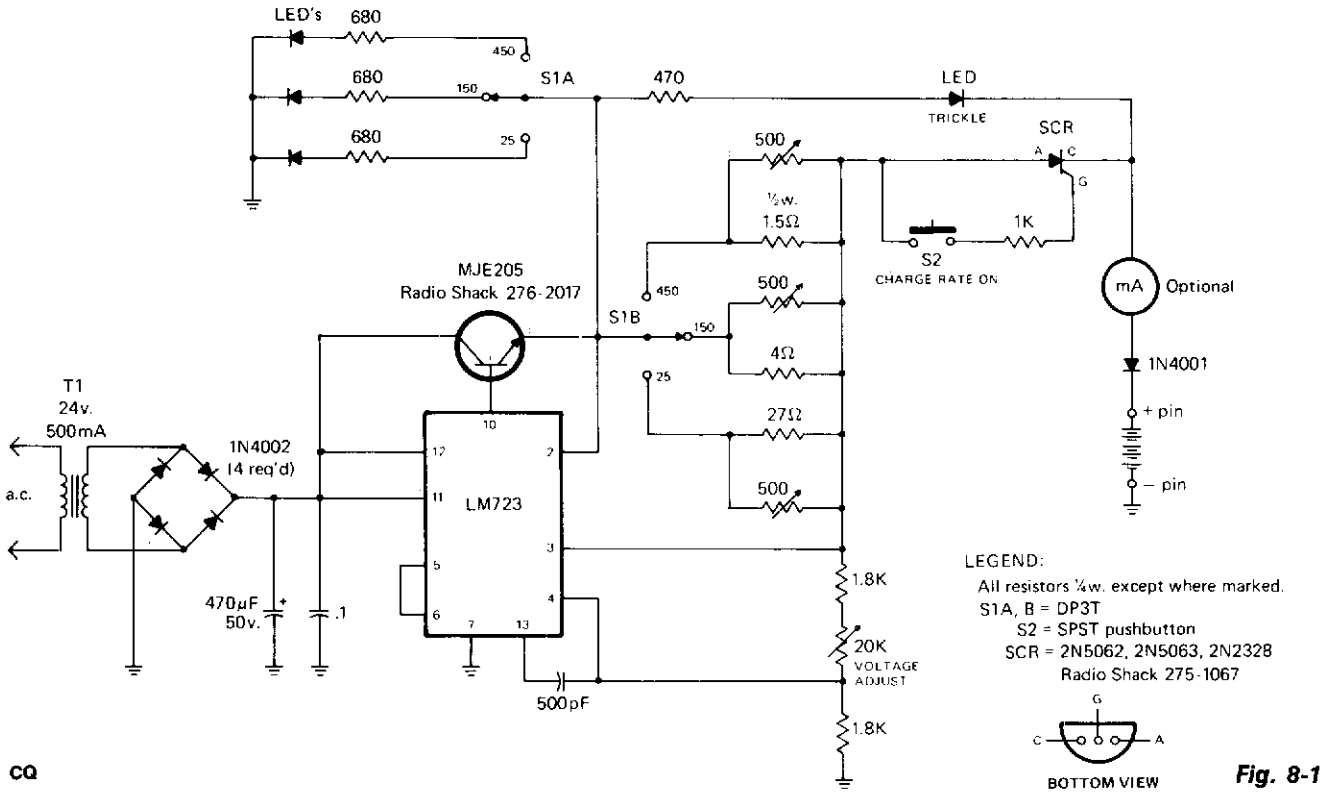
8

Battery Chargers and Zappers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Rapid Battery Charger for ICOM IC-2A
Gel Cell Charger
Ni-Cad Battery Zapper
Lithium Battery Charger
Thermally Controlled Ni-Cad Charger
Ni-Cad Battery Zapper II
Battery Charger
Wind Powered Battery Charger
Battery Charger Operates On Single Solar Cell
Versatile Battery Charger
14-Volt, 4-Amp Battery Charger/Power Supply

RAPID BATTERY CHARGER FOR ICOM IC-2A



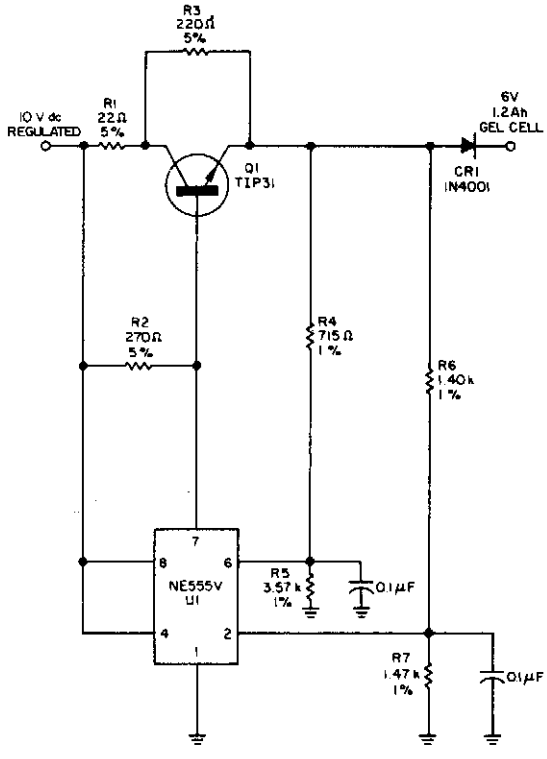
CQ

Fig. 8-1

Circuit Notes

Rectified and filtered voltage from the 24 Vac transformer is applied to the LM723 voltage regulator and the npn pass transistor set up for constant current supply. The 470 ohm resistor limits trickle current until the momentary pushbutton (S2) is depressed, the SCR turns on and current flows through the previously determined resistor network limiting the charging current. The SCR will turn off when the thermal cutout circuit inside the battery pack opens up.

GEL CELL CHARGER



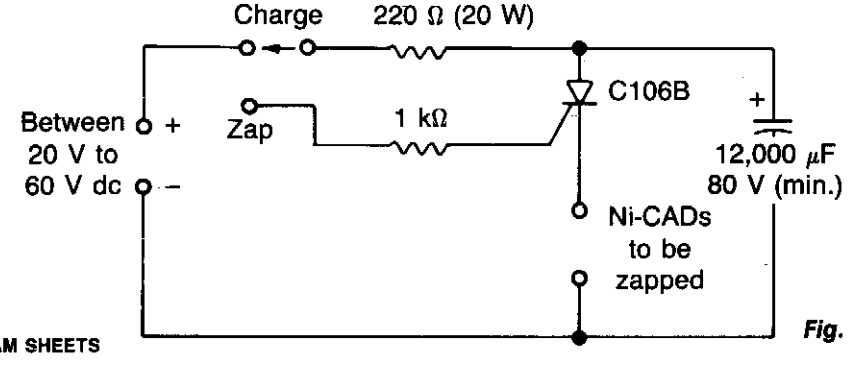
ELECTRONIC DESIGN

Circuit Notes

This circuit detects a full-charge state and automatically switches to a float condition—from 240 mA to 12 mA.

Fig. 8-2

Ni-CAD BATTERY ZAPPER



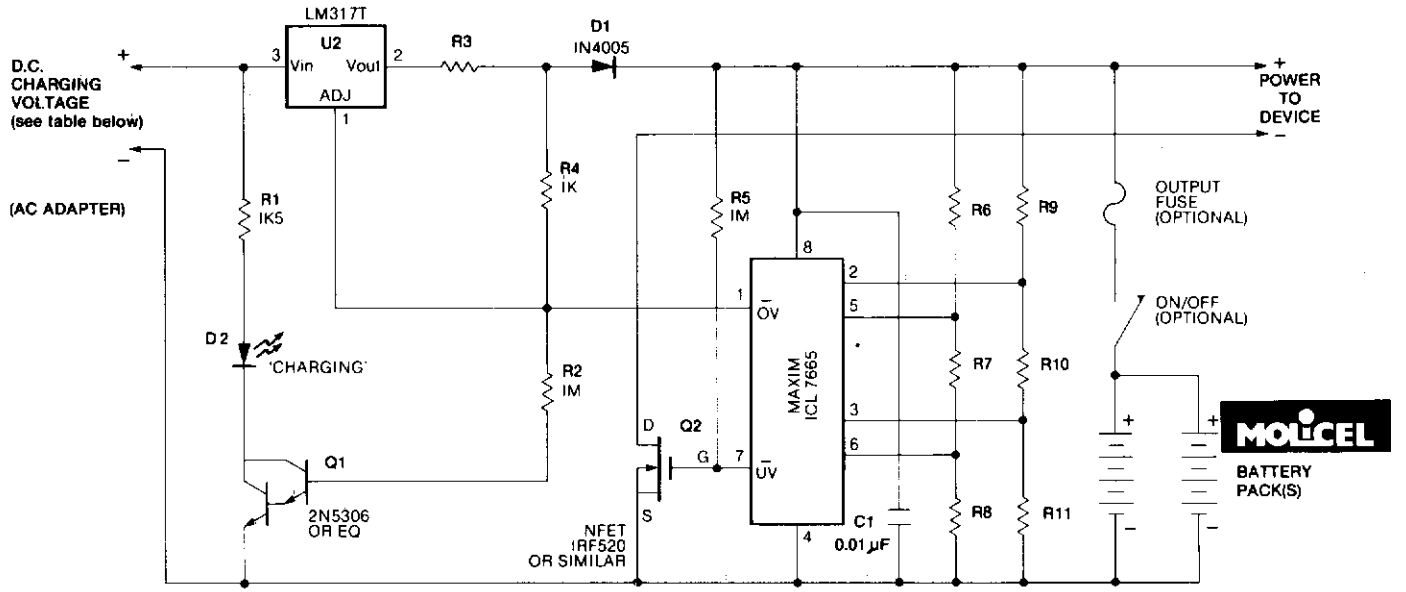
WILLIAM SHEETS

Fig. 8-3

Circuit Notes

The short in a Ni-Cad battery can be "burned off" with this zapper. Use of the SCR keeps heavy discharge current from damaging switch contacts.

LITHIUM BATTERY CHARGER



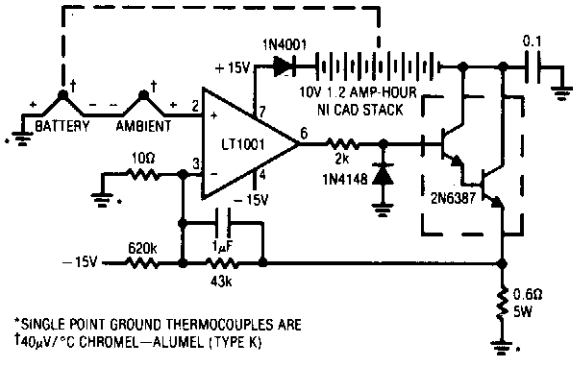
MOLI ENERGY LIMITED

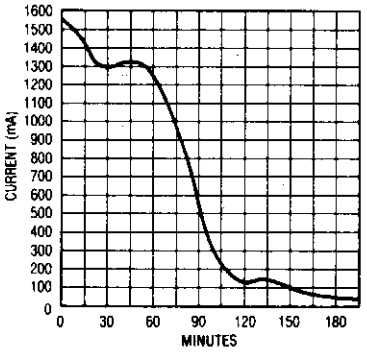
Fig. 8-4

Circuit Notes

Charging is accomplished with a constant current of 60 mA for AA cells to a cutoff voltage of 2.4 V per cell at which point the charge must be terminated. The charging system shown is designed for multi-cell battery packs of 2 to 6 series-connected cells or series/parallel arrangements. It is essential that all cells assembled in the pack be at an identical state-of-charge (voltage) prior to charging. The maximum upper cut-off voltage is 15.6 volts (6 × 2.6 V).

THERMALLY CONTROLLED Ni-CAD CHARGER





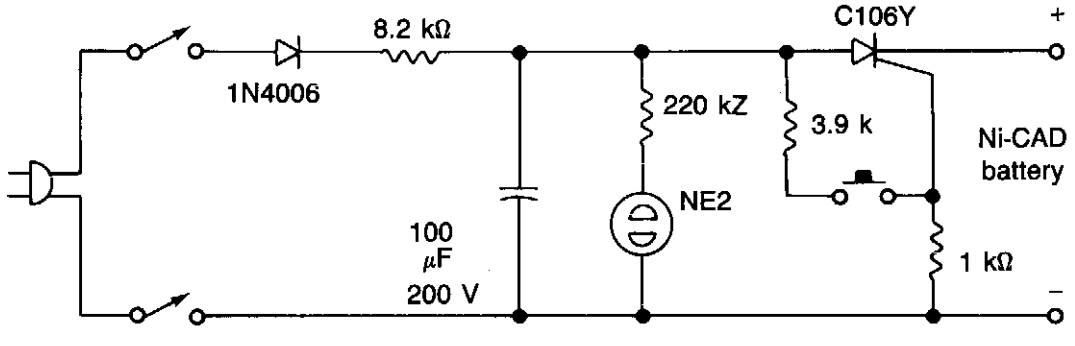
LINEAR TECHNOLOGY CORPORATION

Fig. 8-5

Circuit Notes

One way to charge Ni-Cad batteries rapidly without abuse is to measure cell temperature and taper the charge accordingly. The circuit uses a thermocouple for this function. A second thermocouple nulls out the effects of ambient temperature. The temperature difference between the two thermocouples determines the voltage which appears at the amplifier's positive input. As battery temperature rises, this small negative voltage (1°C difference between the thermocouples equals 40 μV) becomes larger. The amplifier, operating at a gain of 4300, gradually reduces the current through the battery to maintain its inputs at balance. The battery charges at a high rate until heating occurs and the circuit then tapers the charge. The values given in the circuit limit the battery surface temperature rise over ambient to about 5°C.

NI-CAD BATTERY ZAPPER II



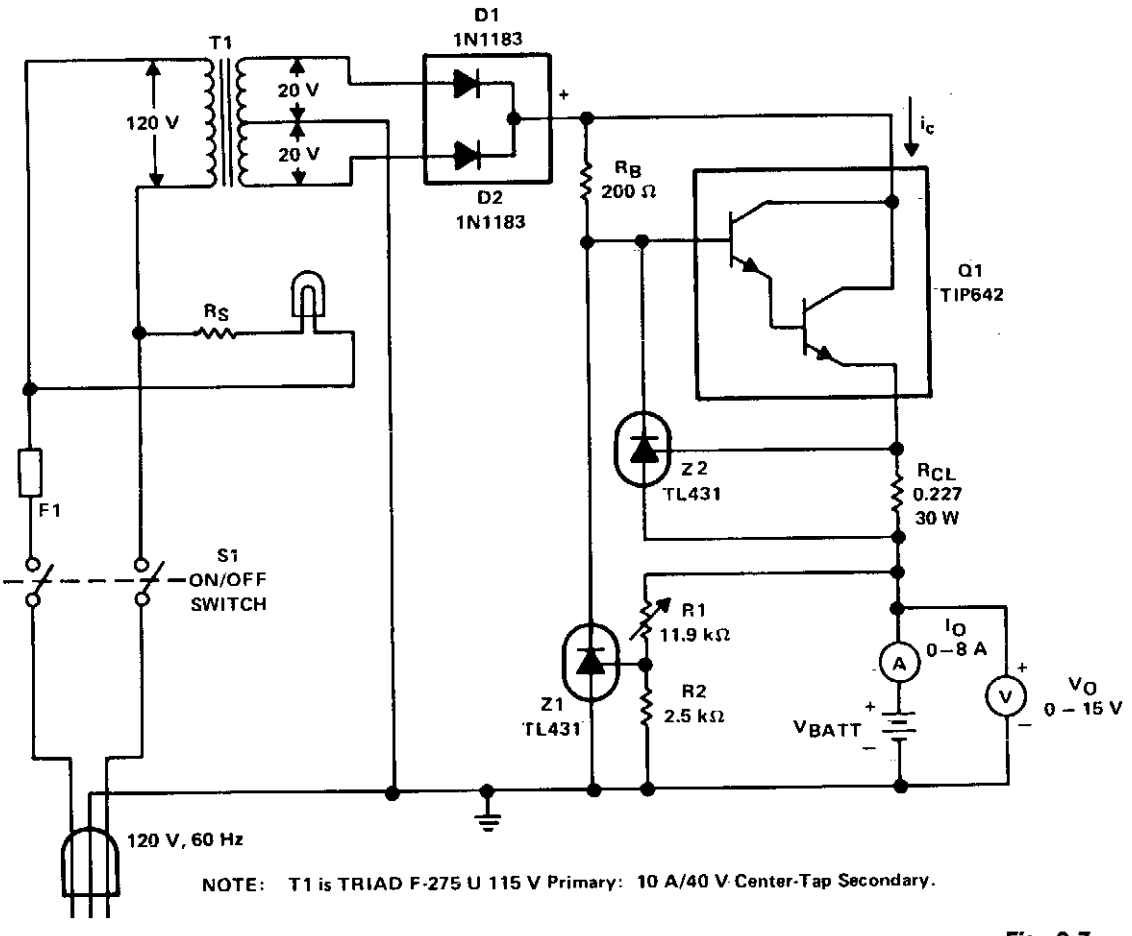
WILLIAM SHEETS

Fig. 8-6

Circuit Notes

This zapper clears internal short in nickel cadmium batteries by burning it away. CAUTION: The negative battery terminal is connected to one side of the ac line. For safety operation use a 1:1 isolation transformer.

BATTERY CHARGER



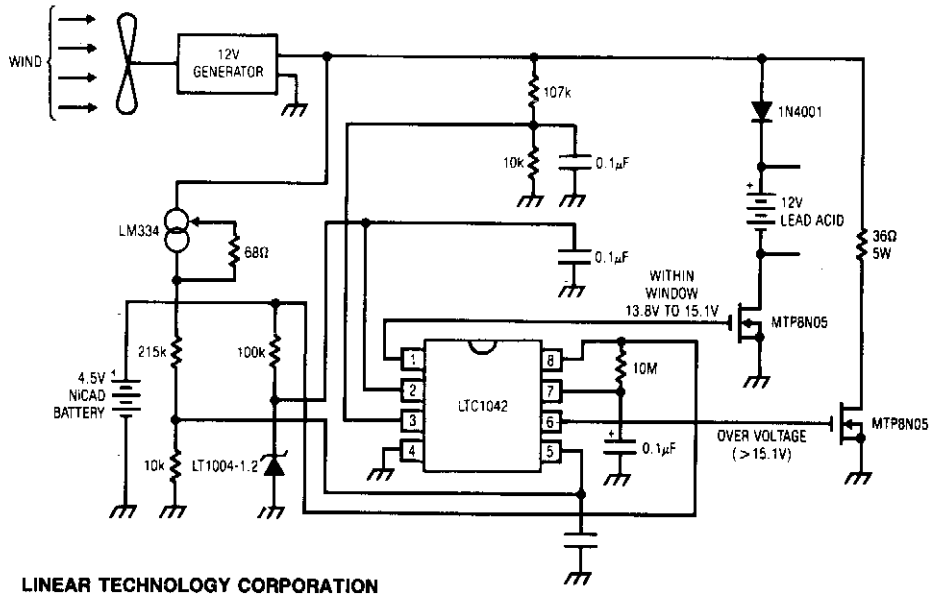
TEXAS INSTRUMENTS

Fig. 8-7

Circuit Notes

The charger is based on a charging voltage of 2.4 V per cell, in accordance with most manufacturers' recommendations. The circuit pulses the battery under charge with 14.4 V (6 cells × 2.4 V per cell) at a rate of 120 Hz. The design provides current limiting to protect the charger's internal components while limiting the charging rate to prevent damaging severely discharged lead-acid batteries. The maximum recommended charging current is normally about one-fourth the ampere-hour rating of the battery. For example, the maximum charging current for an average 44 ampere-hour battery is 11 A. If the impedance of the load requires a charging current greater than the 11 A current limit, the circuit will go into current limiting. The amplitude of the charging pulses is controlled to maintain a maximum peak charging current of 11 A (8 A average).

WIND POWERED BATTERY CHARGER



LINEAR TECHNOLOGY CORPORATION

Fig. 8-8

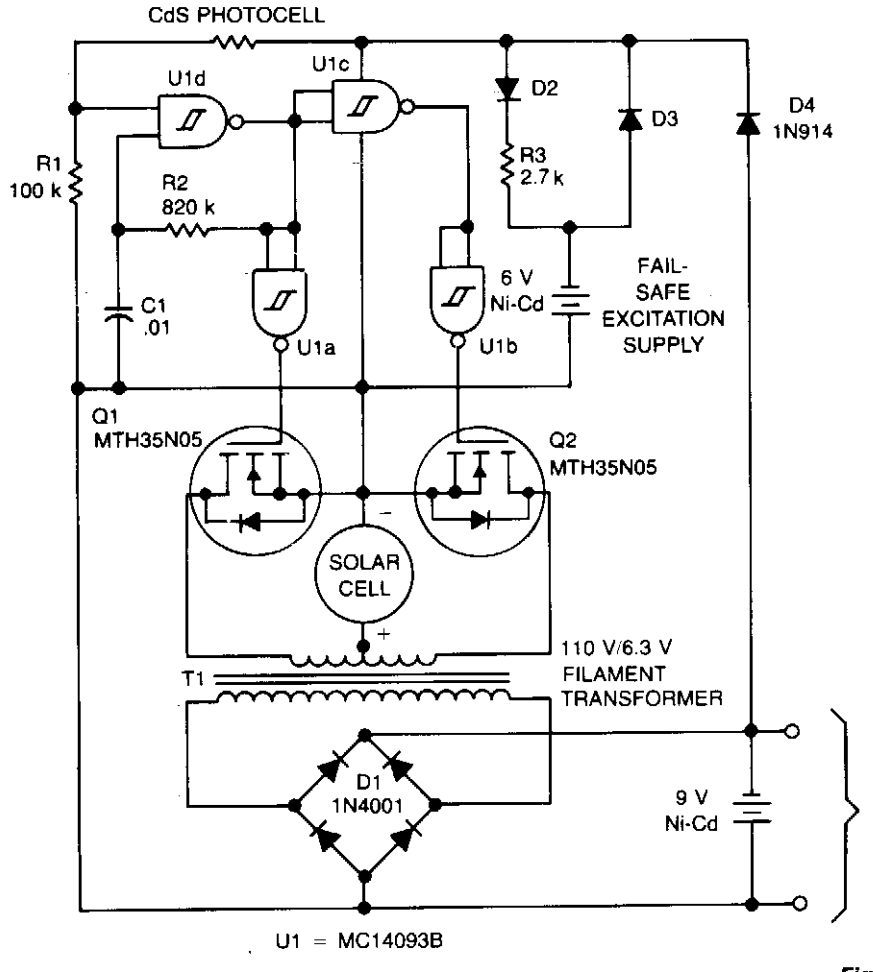
Circuit Notes

The dc motor is used as a generator with the voltage output being proportional to its rpm. The LTC1042 monitors the voltage output and provides the following control functions.

1. If generator voltage output is below 13.8 V, the control circuit is active and the Ni-Cad battery is charging through the LM334 current source. The lead acid battery is not being charged.
2. If the generator voltage output is between 13.8 V and 15.1 V, the 12 V lead acid battery is being charged at about 1 amp/hour rate (limited by the power FET).
3. If generator voltage exceeds 15.1 V (a condition caused by excessive wind speed or 12 V battery being fully charged) then a fixed load is connected limiting the generator rpm to prevent damage.

This charger can be used as a remote source of power where wind energy is plentiful such as on sailboats or remote radio repeater sites. Unlike solar powered panels, this system will function in bad weather and at night.

BATTERY CHARGER OPERATES ON SINGLE SOLAR CELL



MOTOROLA

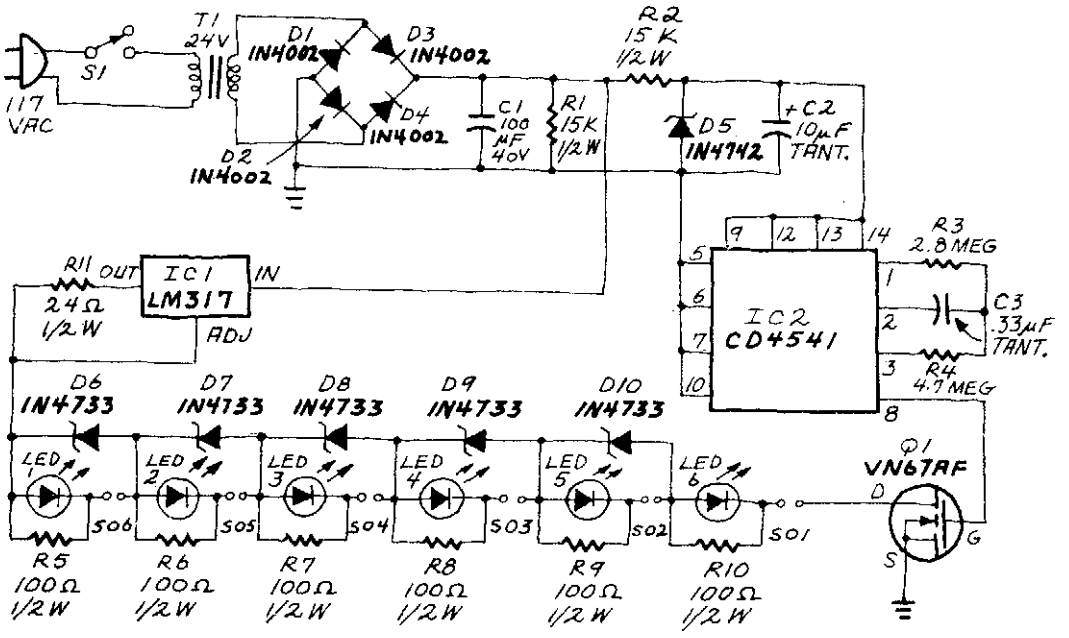
Fig. 8-9

Circuit Notes

The circuit charges a 9-V battery at about 30 mA per input ampere at 0.4 V. U1, a quad Schmitt trigger, operate as an astable multivibrator to drive push-pull TMOS devices Q1 and Q2. Power for U1 is derived from the 9-V battery via D4; power for Q1 and Q2 is supplied by the solar cell. The multivibrator frequency, determined by R2-C1, is set to 180 Hz for maximum efficiency from a 6.3-V filament transformer, T1. The secondary of the transformer is applied to a full wave bridge rectifier, D1, which is connected to the batteries being charged. The small Ni-Cad battery is a fail-safe excitation supply to allow the system to recover if the 9-V battery becomes fully discharged.

A CdS photocell shuts off the oscillator in darkness to preserve the fail-safe battery during shipping and storage, or prolonged darkness.

VERSATILE BATTERY CHARGER



RADIO ELECTRONICS

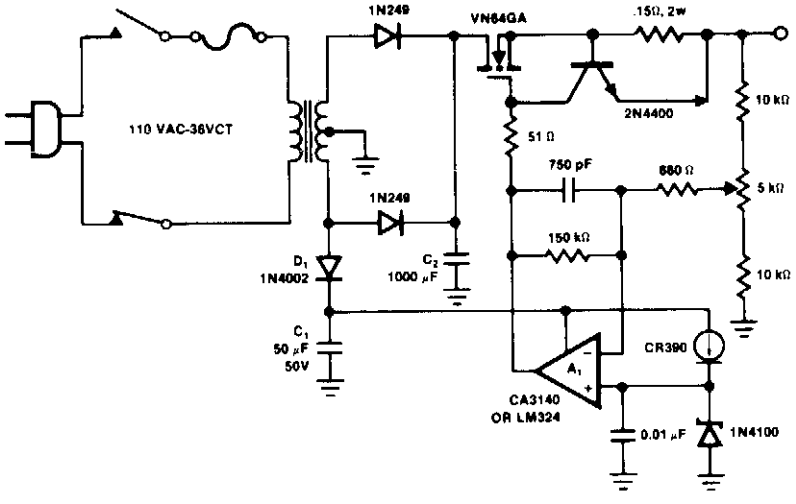
Fig. 8-10

Circuit Notes

An LM317 voltage regulator is configured as a constant-current source. It is used to supply the 50 mA charging current to S01-S06, an array of AA-cell battery holders. Each of the battery holders is wired in series with an LED and its associated shunt resistor. When the battery holder contains a battery, the LED glows during charging. Each battery holder/LED combination is paralleled by a 5.1-volt Zener diode. If the battery holder is empty, the Zener conducts the current around the holder.

A timing circuit prevents overcharging. When power is applied to the circuit, timing is initiated by IC2, a CD4541 oscillator/programmable timer. The output of IC2 is fed to Q1. When that output is high, the transistor is on, and the charging circuit is completed. When the output is low, the transistor is off, and the path to ground is interrupted.

14-VOLT, 4-AMP BATTERY CHARGER/POWER SUPPLY



SILICONIX, INC.

Fig. 8-11

Circuit Notes

Operation amplifier A1 directly drives the VN64GA with the error signal to control the output voltage. Peak rectifier D1, C1 supplies error amplifier A1 and the reference zener. This extra drive voltage must exceed its source voltage by several volts for the VN64GA to pass full load current. The output voltage is pulsating dc which is quite satisfactory for battery charging. To convert the system to a regulated dc supply, capacitor C2 is increased and another electrolytic capacitor is added across the load. The response time is very fast, being determined by the op-amp. The 2N4400 current limiter circuit prevents the output current from exceeding 4.5 A. However, maintaining a shorted condition for more than a second will cause the VN64GA to exceed its temperature ratings. A generous heat sink, on the order of 1°C/W, must be used.

9

Battery Monitors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Dynamic, Constant Current Load for Fuel Cell/Battery Testing
Voltage Detector Relay for Battery Charger
Battery Status Indicator
Low-Battery Indicator
A Lithium Battery's State-of-Charge Indicator
Step-Up Switching Regulator for 6-V Battery
Battery Voltage Monitor
Battery Monitor

DYNAMIC, CONSTANT CURRENT LOAD FOR FUEL CELL/BATTERY TESTING

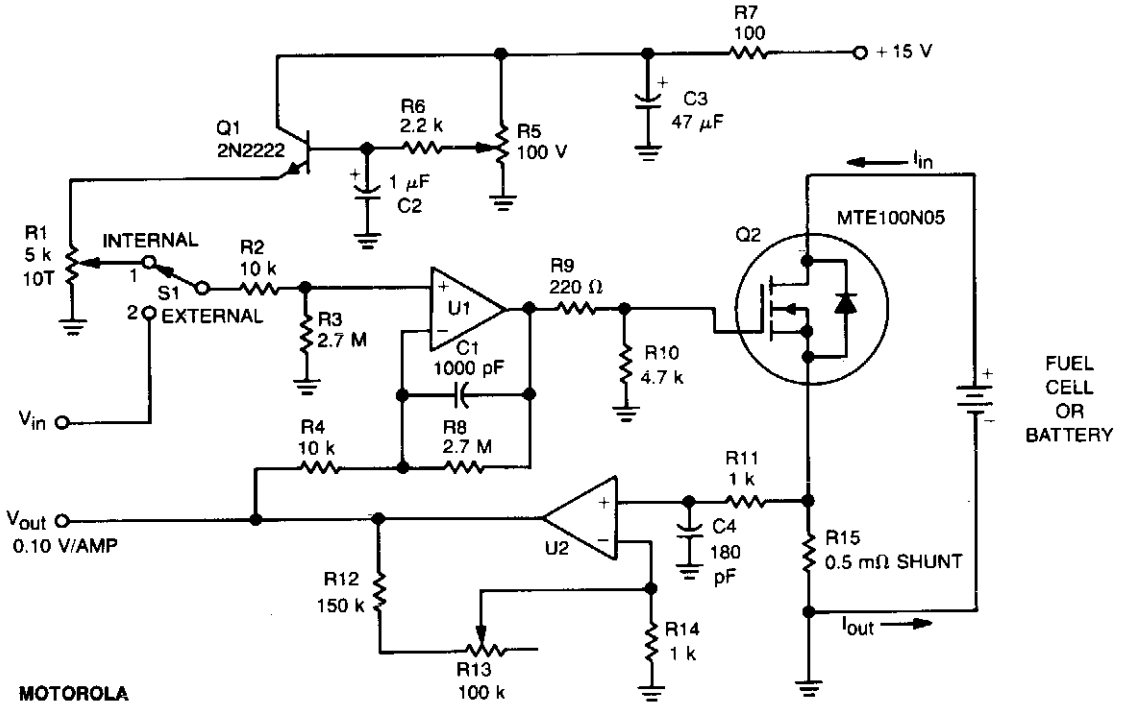


Fig. 9-1

Circuit Notes

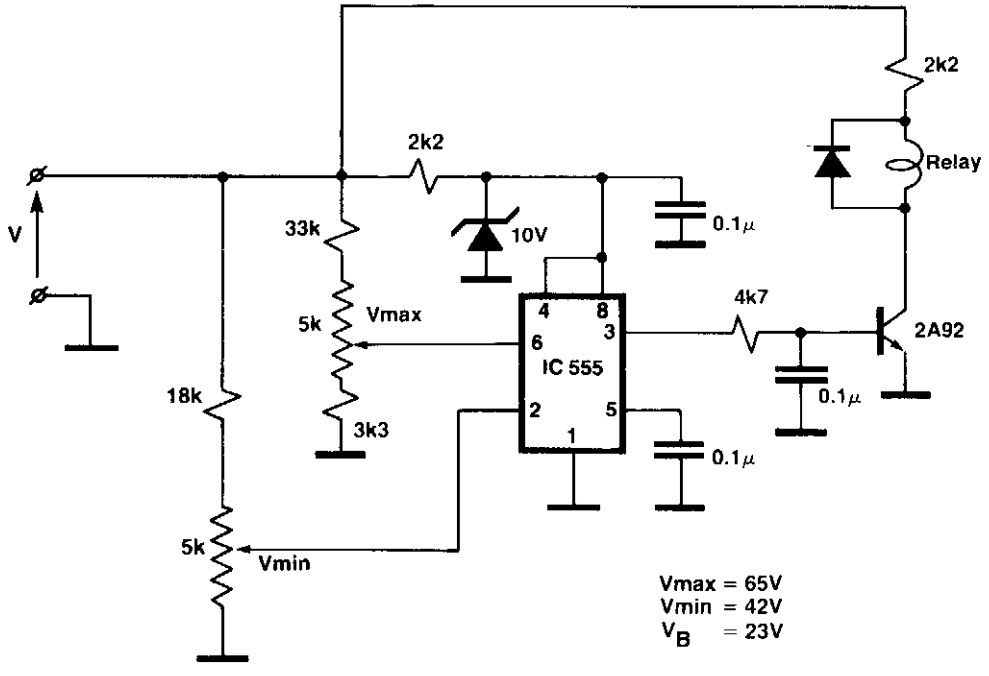
This circuit was designed for testing fuel cells, but it could also be used for testing batteries under a constant current load. It provides a dynamic, constant current load, eliminating the need to manually adjust the load to maintain a constant load.

For fuel cell application, the load must be able to absorb 20-40 A, and since a single cell develops only 0.5 to 1.0 V, bipolar power devices (such as a Darlington) are impractical. Therefore, this dynamic load was designed with a TMOS Power FET (Q2).

With switch S1 in position 1, emitter follower Q1 and R1 establish the current level for the load. In position 2, an external voltage can be applied to control the current level.

Operational amplifier U1 drives TMOS device Q1, which sets the load current seen by the fuel cell or battery. The voltage drop across R15, which is related to the load current, is then applied to U2, whose output is fed back to U1. Thus, if the voltage across R15 would tend to change, feedback to the minus input of U1 causes that voltage (and the load current) to remain constant. Adjustment of R13 controls the volts/amp of feedback. The V_{OUT} point is used to monitor the system.

VOLTAGE DETECTOR RELAY FOR BATTERY CHARGER



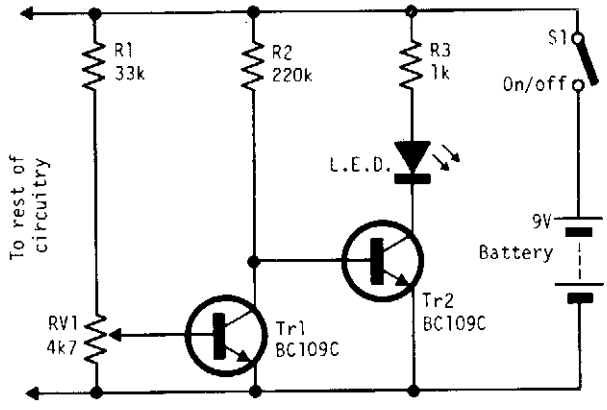
ELECTRONIC ENGINEERING

Fig. 9-2

Circuit Notes

While the battery is being charged, its voltage is measured at V. If the measured voltage is lower than the minimum the relay will be energized, that will connect the charger circuit. When the battery voltage runs over the maximum set point, the relay is deenergized and it will be held that way until the voltage decreases below the minimum when it will be connected again. The voltage is lower than a threshold V_B (low breaking voltage) the relay will be assumed that such a low voltage is due to one or several damaged battery components. Of course V_B is much lower than the minimum set point.

BATTERY STATUS INDICATOR



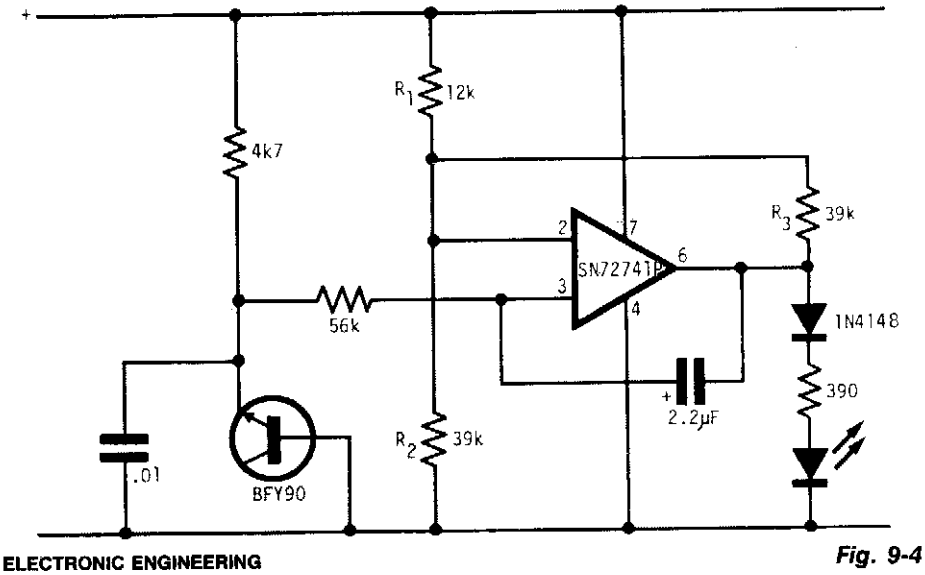
ELECTRONIC ENGINEERING

Fig. 9-3

Circuit Notes

Continually monitors battery voltage during use and consumes only about 250 μA (until the end point is reached). Near the end point Tr1 turns off, allowing Tr2 to illuminate the LED to increase current drain further leading to a distinct turn off point.

LOW-BATTERY INDICATOR



ELECTRONIC ENGINEERING

Fig. 9-4

Circuit Notes

Under good battery conditions the LED is off. As the battery voltage falls, the LED begins to flash until, in the low battery condition, the LED lights continuously. Designed for a 9-volt battery, with the values shown the LED flashes from 7.5 to 6.5 volts.

A LITHIUM BATTERY'S STATE-OF-CHARGE INDICATOR

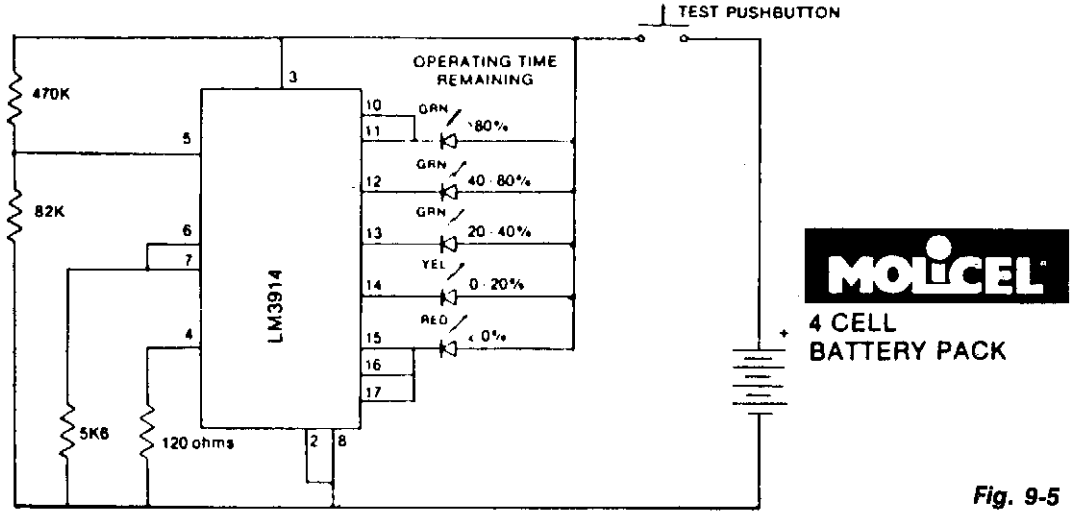


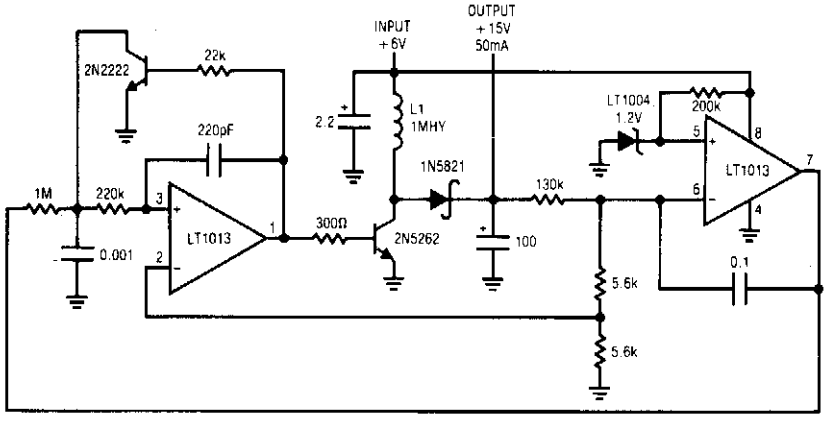
Fig. 9-5

MOLI ENERGY LIMITED

Circuit Notes

State-of-Charge indication of a sloping-voltage discharge can be used as a state-of-charge indicator. A typical voltage comparator circuit that gives a visual indication of state-of-charge is shown. Components identified are for a 4-cell input voltage of 9.6 to 5.2 volts.

STEP-UP SWITCHING REGULATOR FOR 6-V BATTERY

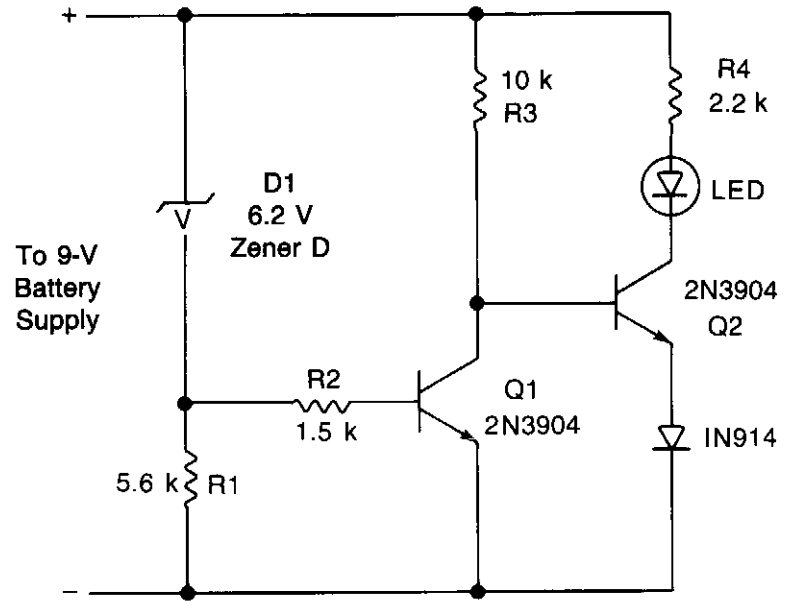


L1 = AIE–VERNITRON 24-104
78% EFFICIENCY

LINEAR TECHNOLOGY CORP.

Fig. 9-6

BATTERY VOLTAGE MONITOR



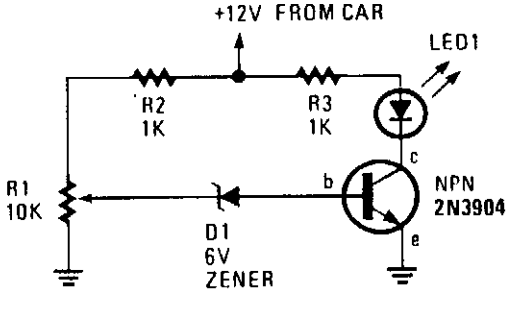
WILLIAM SHEETS

Fig. 9-7

Circuit Notes

This circuit gives an early warning of the discharge of batteries. Zener diode D1 is chosen for the voltage below which an indication is required (9 V). Should the supply drop to below 7 V, D1 will cease conducting causing Q1 to shut off. Its collector voltage will now increase causing Q2 to start conducting via LED1 and its limiting resistor R4.

BATTERY MONITOR



TAB BOOKS, INC.

Fig. 9-8

Circuit Notes

The circuit is quick and easy to put together and install, and tells you when battery voltage falls below the set limit as established by R1 (a 10,000-ohm potentiometer). It can indicate, via LED1, that the battery may be defective or in need of change if operating the starter causes the battery voltage to drop below the present limit.

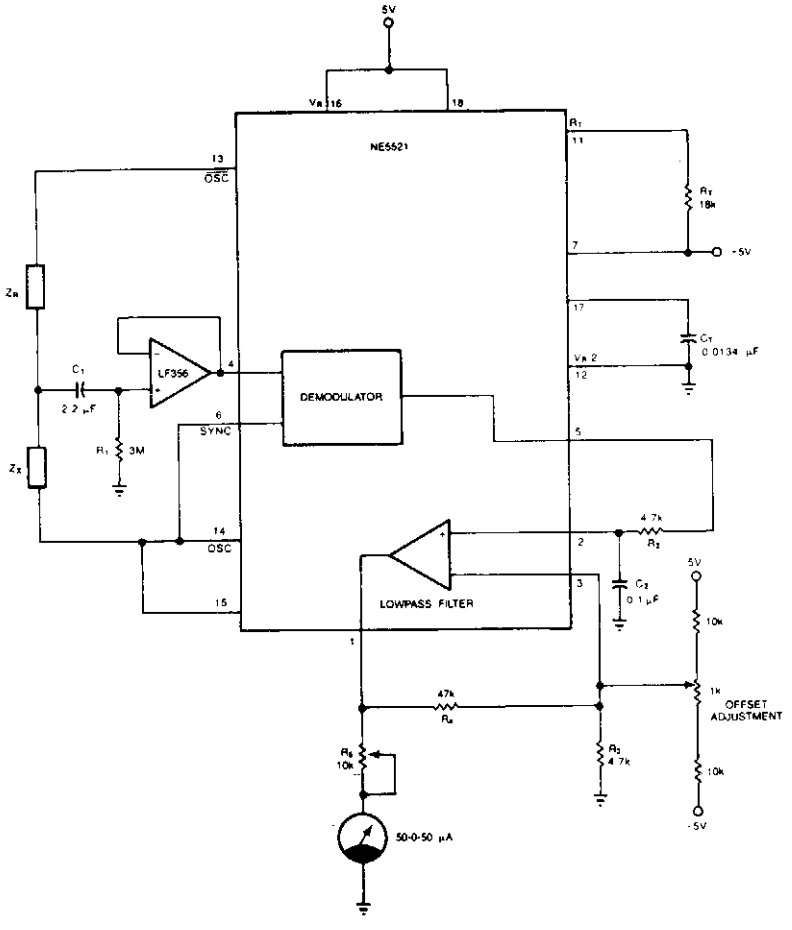
10

Bridge Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Ac Bridge
Bridge-Balance Indicator
Bridge Circuit
Typical Two Op Amp Bridge-Type Differential Amplifier
Low-Power Common Source Amplifier
Amplifier for Bridge Transducers
Strain Gage Bridge Signal Conditioner

AC BRIDGE



SIGNETICS

Fig. 10-1

Circuit Notes

The circuit provides a simple and cost-effective solution to matching resistors and capacitors. Impedances Z_R and Z_X form a half-bridge, while OSC and $\overline{OSC}$ excite the bridge differentially. The external op amp is a FET input amplifier (LF356) with very low input bias current on the order of 30 pA (typical). C1 allows ac coupling by blocking the dc common mode voltage from the bridge, while R1 biases the output of LF356 to 0 V at dc. Use of FET input op amp insures that dc offset due to bias current through R1 is negligible. Ac output of the demodulator is filtered via the uncommitted amp to provide dc voltage for the meter. The 10 k potentiometer, R5, limits the current into the meter to a safe level. Calibration begins by placing equal impedances at Z_R and Z_X, and the system offset is nulled by the offset adjust circuit so that Pin 1 is at 0 V. Next, known values are placed at Z_X and the meter deviations are calibrated. The bridge is now ready to measure an unknown impedance at Z_X with $\pm 0.05\%$ accuracy or better.

BRIDGE-BALANCE INDICATOR

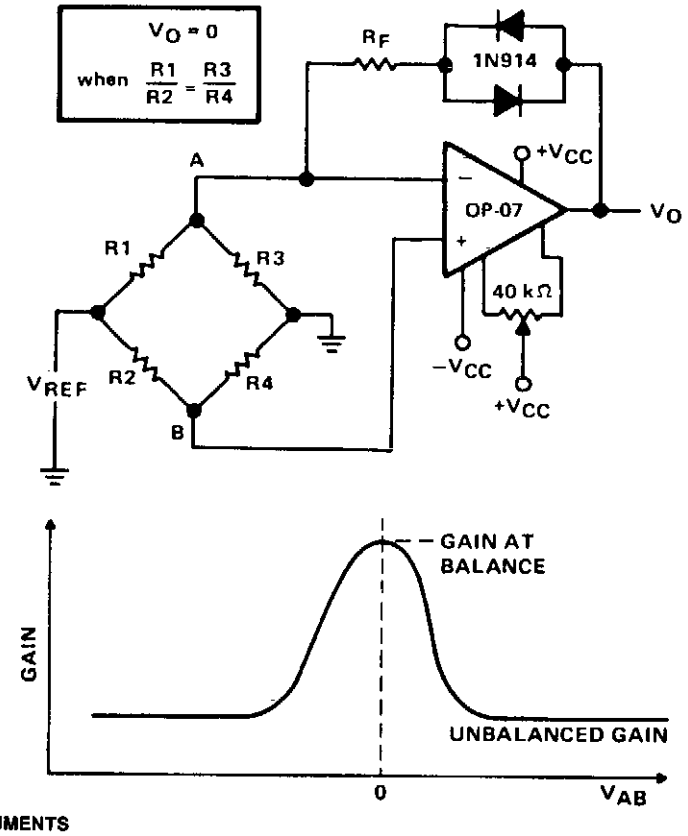


Fig. 10-2

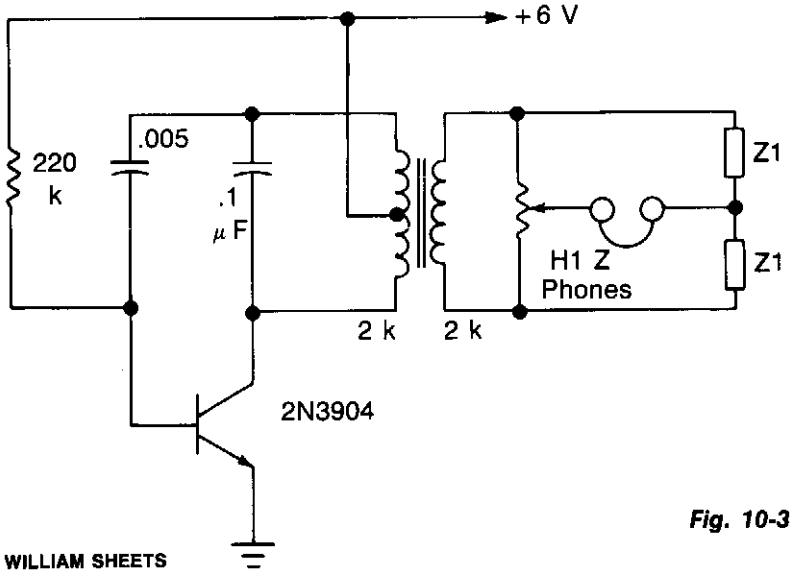
TEXAS INSTRUMENTS

Circuit Notes

Indicator provides an accurate comparison of two voltages by indicating their degree of balance (or imbalance). Detecting small variations near the null point is difficult with the basic Wheatstone bridge alone. Amplification of voltage differences near the null point will improve circuit accuracy and ease of use.

The 1N914 diodes in the feedback loop result in high sensitivity near the point of balance (R1/R2 = R3/R4). When the bridge is unbalanced the amplifier's closed-loop gain is approximately R_F/r, where r is the parallel equivalent of R1 and R3. The resulting gain equation is $G = R_F(1/R1 + 1/R3)$. During an unbalanced condition the voltage at point A is different from that at point B. This difference voltage (V_{AB}), amplified by the gain factor G, appears as an output voltage, As the bridge approaches a balanced condition (R1/R2 = R3/R4), V_{AB} approaches zero. As V_{AB} approaches zero the 1N914 diodes in the feedback loop lose their forward bias and their resistance increases, causing the total feedback resistance to increase. This increases circuit gain and accuracy in detecting a balanced condition. The figure shows the effect of approaching balance on circuit gain. The visual indicator used at the output of the OP-07 could be a sensitive voltmeter or oscilloscope.

BRIDGE CIRCUIT



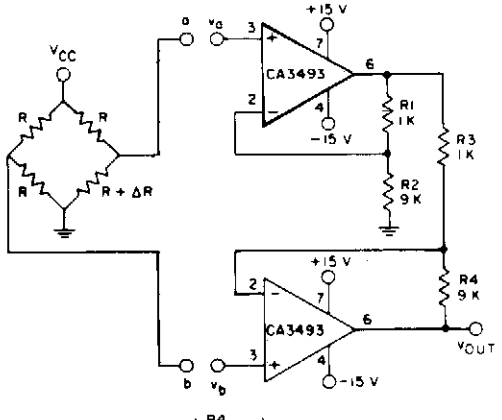
WILLIAM SHEETS

Fig. 10-3

Circuit Notes

The transistor is connected as an audio oscillator, using an audio transformer in the collector. The secondary goes to a linear pot. The ratio between the two parts of the pot from the slider is proportional to the values of Z1 and Z2 when no signal is heard in the phones.

TYPICAL TWO OP AMP BRIDGE-TYPE DIFFERENTIAL AMPLIFIER



$V_{OUT} = V_b - V_a \left(\frac{R4}{R3} + 1\right)$

$A = \frac{V_{OUT}}{V_b - V_a} = \left(\frac{R4}{R3} + 1\right)$

FOR VALUES ABOVE $V_{OUT} = (V_b - V_a)(10)$

ALL RESISTANCE VALUES ARE IN OHMS

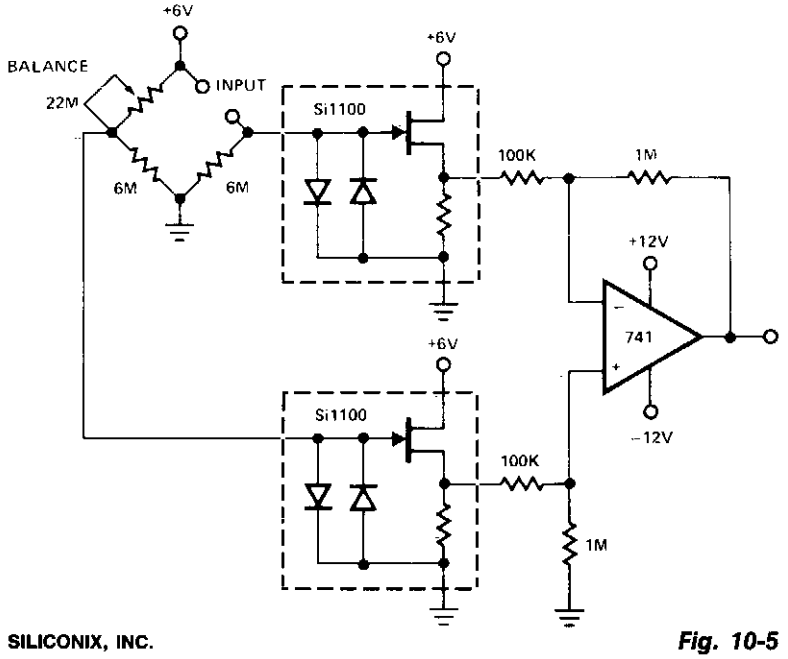
GENERAL ELECTRIC/RCA

Circuit Notes

Using a CA3493 BiMOS op amp to provide high input impedance and good common-mode rejection ratio (depends primarily on matching of resistor networks).

Fig. 10-4

LOW-POWER COMMON SOURCE AMPLIFIER



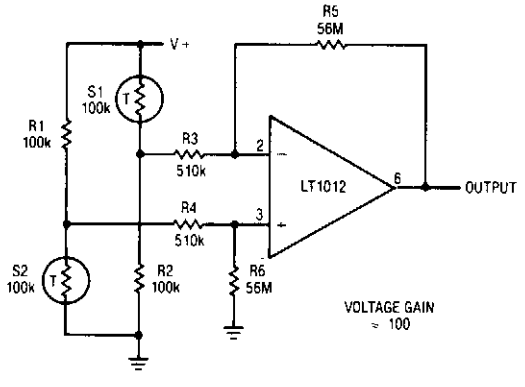
SILICONIX, INC.

Fig. 10-5

Circuit Notes

A circuit that will operate in the 10- to 20- microamp range at a 12-volt supply voltage. The diode protection is available in this configuration. The circuit voltage gain will be between 10 and 20, with extremely low power consumption (approximately 250 μW). This is very desirable for remote or battery operation where minimum maintenance is important.

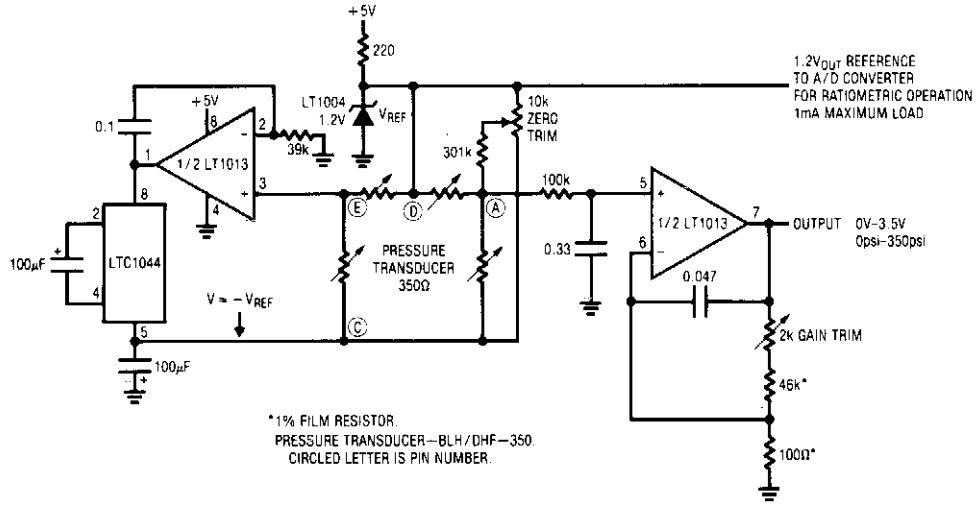
AMPLIFIER FOR BRIDGE TRANSDUCERS



LINEAR TECHNOLOGY CORP.

Fig. 10-6

STRAIN GAUGE BRIDGE SIGNAL CONDITIONER



LINEAR TECHNOLOGY CORP.

Fig. 10-7

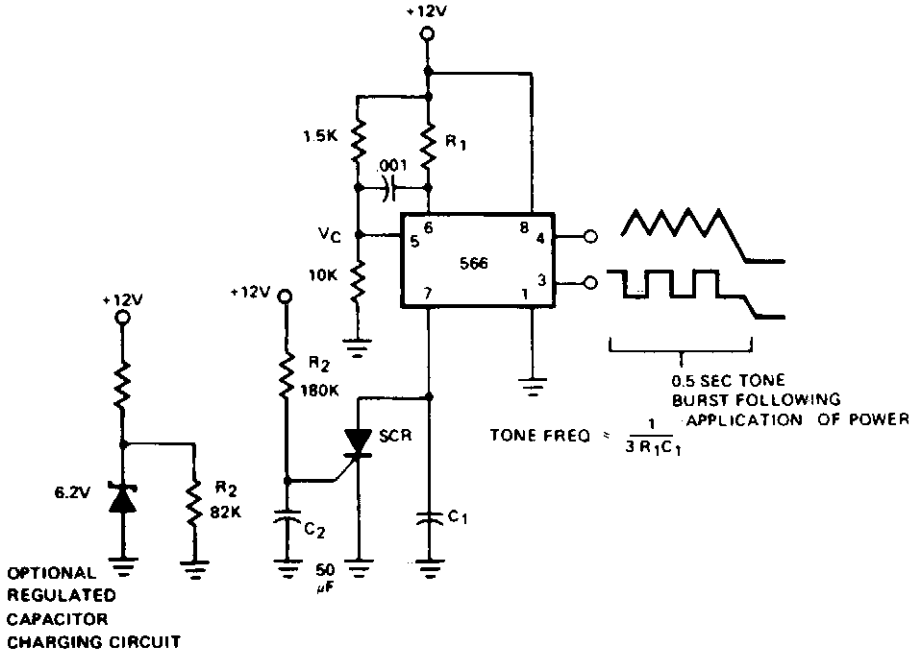
11

Burst Generators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Single-Tone Burst Generator
Square Waveform Multiburst Generator
Single-Timer IC Provides Square-Wave Tone Bursts
Strobe-Tone Burst Generator
Tone Burst Generator

SINGLE-TONE BURST GENERATOR



SIGNETICS

Fig. 11-1

Circuit Notes

The tone burst generator supplies a tone for one-half second after the power supply is activated; its intended use is a communications network alert signal. Cessation of the tone is accomplished at the SCR, which shunts the timing capacitor C1 charge current when activated. The SCR is gated on when C2 charges up to the gate voltage which occurs in 0.5 seconds. Since only 70 μA are available for triggering, the SCR must be sensitive enough to trigger at this level. The triggering current can be increased, of course, by reducing R2 (and increasing C2 to keep the same time constant). If the tone duration must be constant under widely varying supply voltage conditions, the optional Zener diode regulator circuit can be added, along with the new value for R_2 $R_2' = 82\ k\Omega$. If the SCR is replaced by an npn transistor, the tone can be switched on and off at will at the transistor base terminal.

SQUARE WAVEFORM MULTIBURST GENERATOR

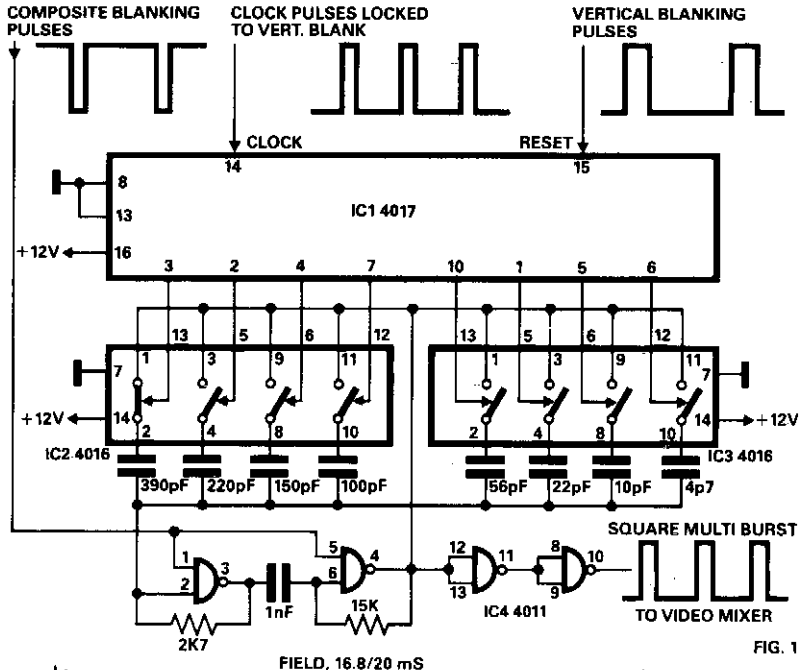
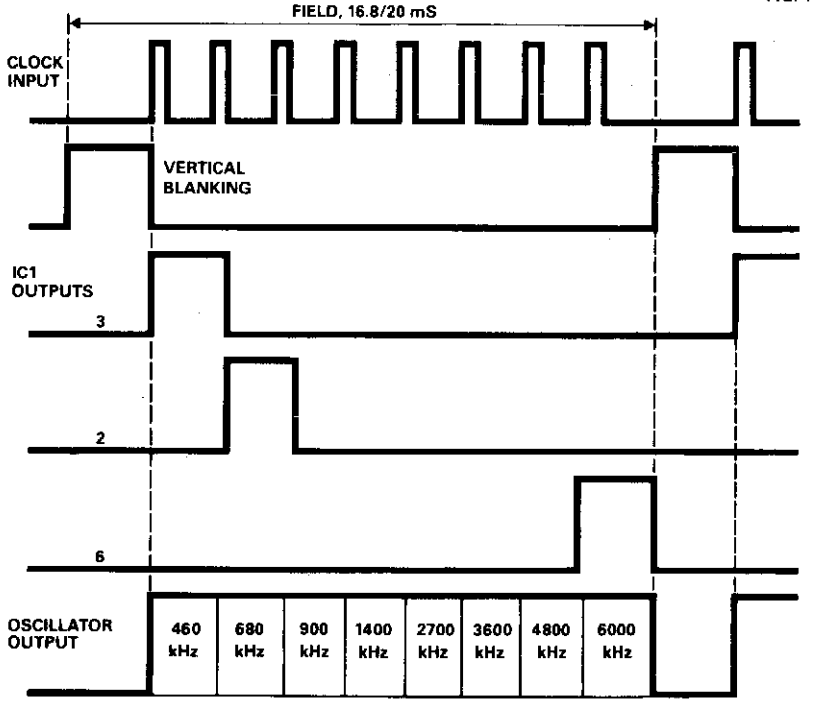


FIG. 1



ELECTRONIC ENGINEERING

Fig. 11-2

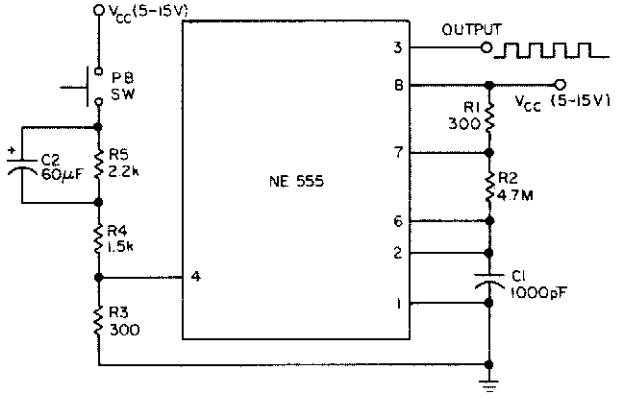
SQUARE WAVEFORM MULTIBURST GENERATOR, Continued.

Circuit Notes

The generator described here is intended for multiburst signal square waveform generation and can be used as a device for characterizing the response of TV monitor amplifiers as shown. The circuit is an RC oscillator with NAND gates (IC4-4011), with its capacitor C changed periodically by means of bilateral switches (IC2, IC3-4016). The control inputs of bilateral switches are driven by the outputs of a counter/decoder (IC1-4017) the operation of which is determined by generated clock pulses, so that they occur eight times at half-picture (field). These pulses are locked to vertical blank pulses.

Horizontal synchronization is achieved by means of composite blanking pulses (negative polarization) applied to pins 1 and 5 of IC4. The oscillator frequency changes in the following discrete steps: 460 kHz, 680 kHz, 900 kHz, 1400 kHz, 2700 kHz, 3600 kHz, for the time of one frame. The video signal is fed on a mixer where it is superimposed with a composite sync signal.

SINGLE-TIMER IC PROVIDES SQUARE-WAVE TONE BURSTS



ELECTRONIC DESIGN

Fig. 11-3

Circuit Notes

The tone-burst generator gives a 50-ms burst of 1.5 kHz square waves with each operation of the pushbutton and can source or sink 200 mA.

STROBE-TONE BURST GENERATOR

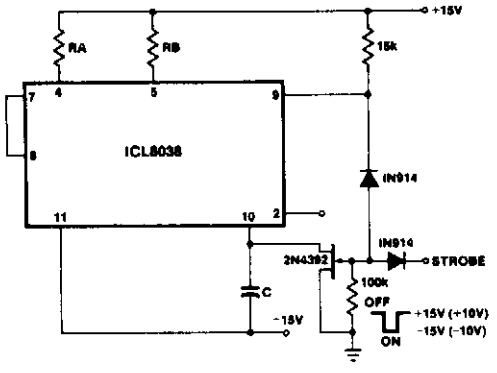
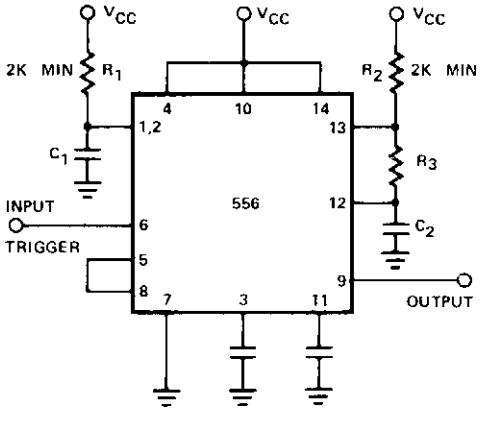


Fig. 11-4

INTERSIL

TONE BURST GENERATOR



Circuit Notes

The dual timer makes an excellent tone burst generator. The first half is connected as a one shot and the second half as an oscillator. The pulse established by the one shot turns on the oscillator allowing a burst of pulses to be generated.

SIGNETICS

Fig. 11-5

12

Capacitance Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Capacitance-to-Voltage Meter
Accurate Digital Capacitance Meter

CAPACITANCE-TO-VOLTAGE METER

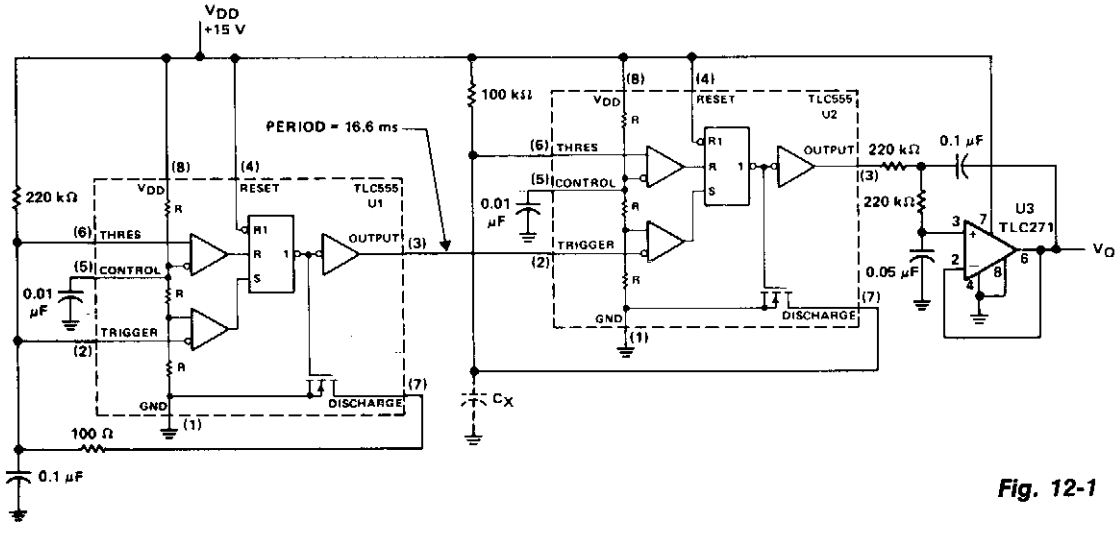


Fig. 12-1

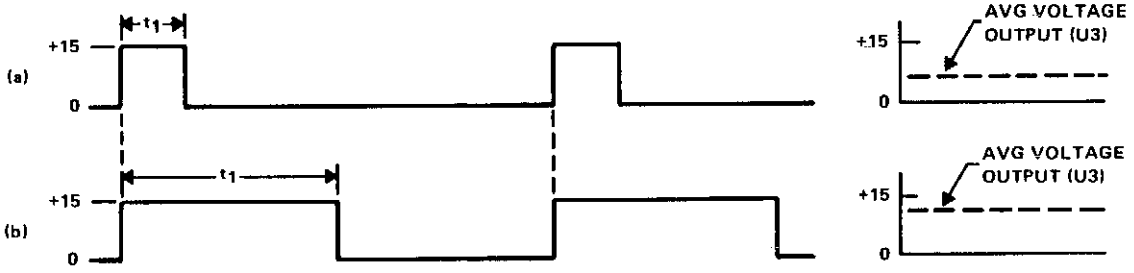
TEXAS INSTRUMENTS

Circuit Notes

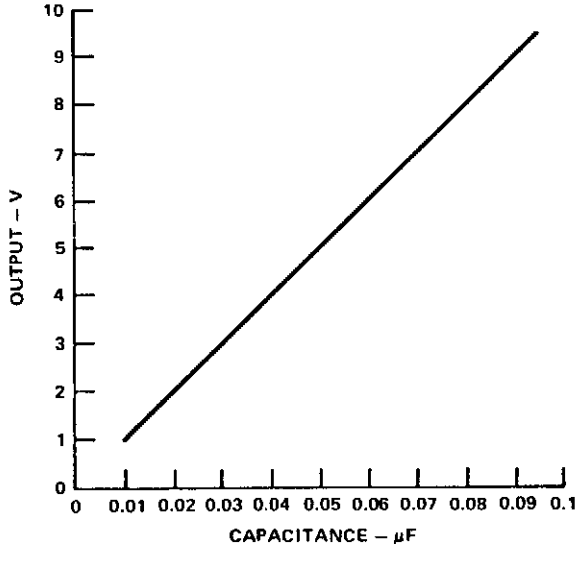
Timer U1 operates as a free-running oscillator at 60 Hz, providing trigger pulses to timer U2 which operates in the monostable mode. Resistor R1 is fixed and capacitor Cx is the capacitor being measured. While the output of U2 is 60 Hz, the duty cycle depends on the value of Cx. U3 is a combination low-pass filter and unity-gain follower whose dc voltage output is the time-averaged amplitude of the output pulses of U2, as shown in the timing diagram.

The diagram shows when the value of Cx is small the duty cycle is relatively low. The output pulses are narrow and produce a lower average dc voltage level at the output of U3. As the capacitance value of Cx increases, the duty cycle increases making the output pulses at U2 wider and the average dc level output at U3 increases. The graph illustrates capacitance values of 0.01 μF to 0.1 μF plotted against the output voltage of U3. Notice the excellent linearity and direct one-to-one scale calibration of the meter. If this does not occur the 100 k ohm resistor, R1, can be replaced with a potentiometer which can be adjusted to the proper value for the meter being used.

CAPACITANCE-TO-VOLTAGE METER (CONT'D)



U2 Duty Cycle Change



TEXAS INSTRUMENTS

Fig. 12-1

ACCURATE DIGITAL CAPACITANCE METER

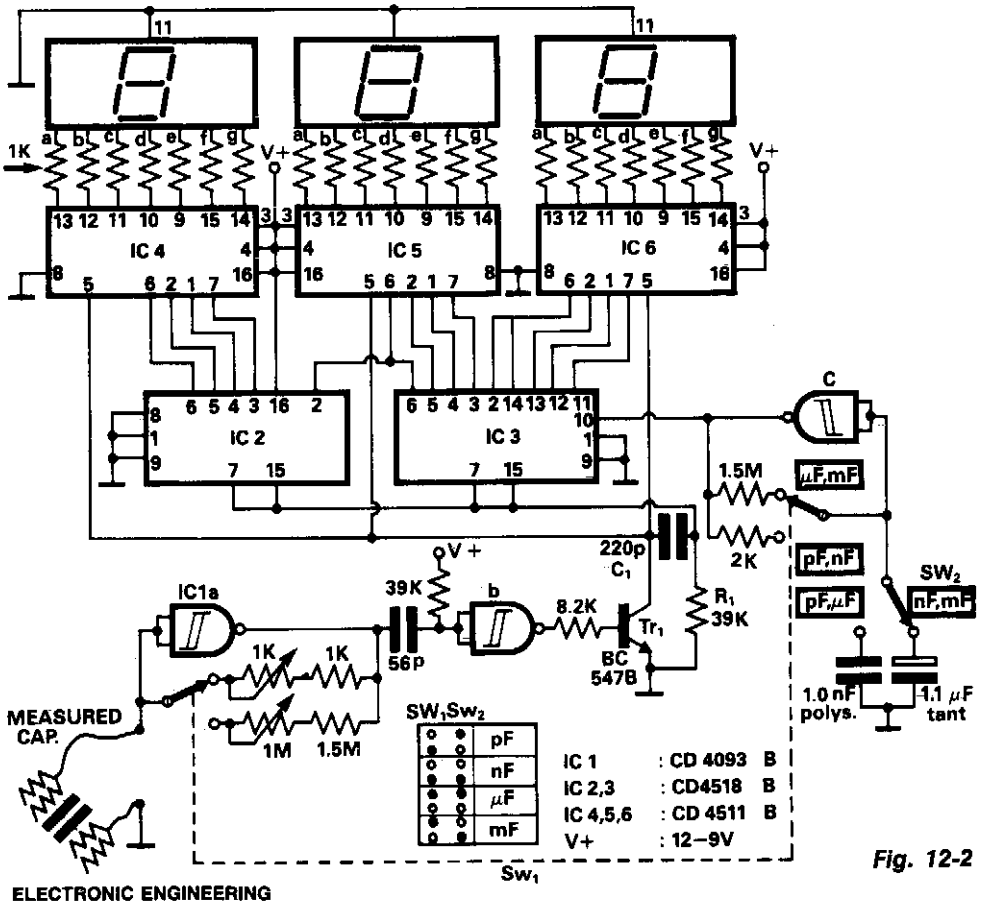


Fig. 12-2

ELECTRONIC ENGINEERING

Circuit Notes

The principle of operation is counting the pulse number derived from a constant frequency oscillator during a fixed time interval produced by another lower frequency oscillator. This oscillator uses the capacitor being measured as the timing. The capacitance measurement is proportional during pulse counting during a fixed time interval. The astable oscillator formed by IC1c produces a pulse train of constant frequency. Gate IC1a also forms an oscillator whose oscillation period is given approximately by the equation: $T = 0.7\ RC$.

Period T is linearly dependent on the capacitance C. This period is used as the time interval for one measurement. The differentiator network following the oscillator creates the negative spikes shaped in narrow pulses by IC1b NAND Schmitt Trigger. The differentiator formed by R1 and C1 produces a negative spike which resets the counters. The display shows the number of high frequency oscillator pulses entering the counter during the measurement period.

13

Circuit Protection Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Overvoltage Protector
High Speed Electronic Circuit Breaker
12 ns Circuit Breaker
Low Voltage Power Disconnector
Automatic Power-Down Protection Circuit
Line Dropout Detector
Electronic Crowbar

OVERVOLTAGE PROTECTOR

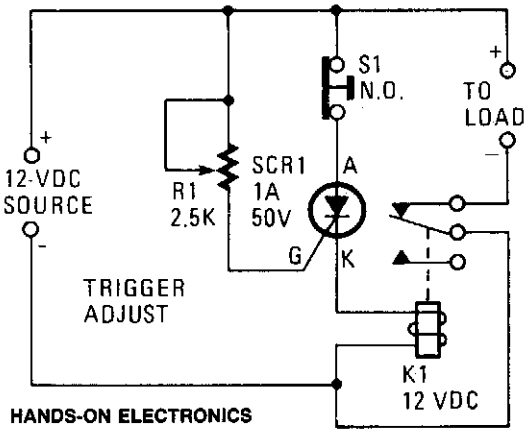


Fig. 13-1

Circuit Notes

A silicon-controlled rectifier is installed in parallel with the 12-V line and connected to a normally-closed 12-V relay, K1. The SCR's gate circuit is used to sample the applied voltage. As long as the applied voltage stays below a given value, SCR1 remains off and K1's contacts remain closed, thereby supplying power to the load. When the source voltage rises above 12 V, sufficient current is applied to the gate of SCR1 to trigger it into conduction. The trigger point of SCR1 is dependent on the setting of R1. Once SCR1 is triggered (activating the relay), K1's contacts open, halting current flow to the load.

HIGH SPEED ELECTRONIC CIRCUIT BREAKER

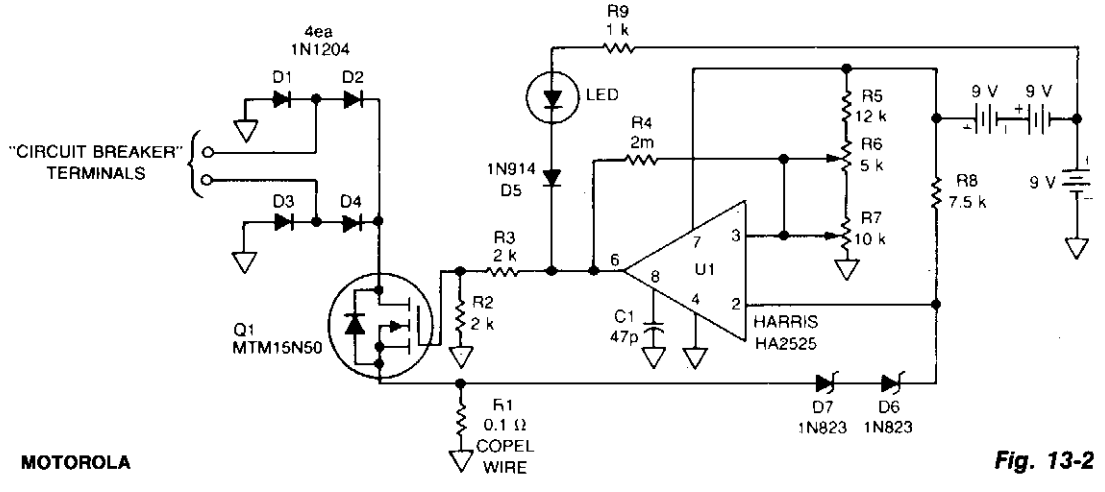


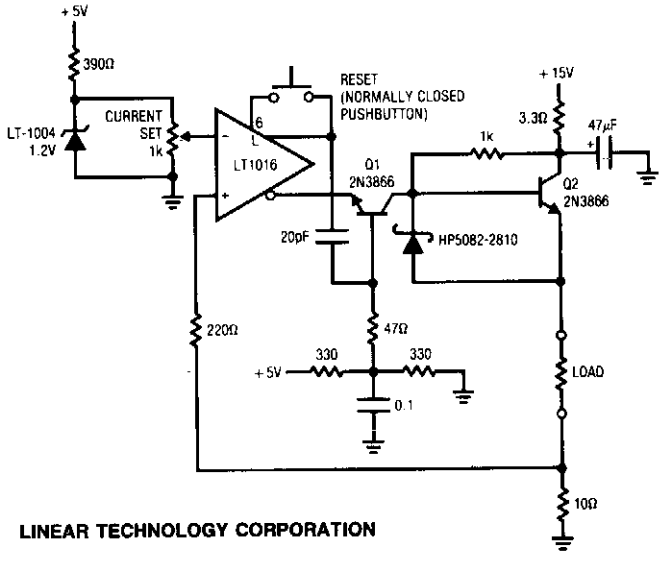
Fig. 13-2

Circuit Notes

This 115 Vac, electronic circuit breaker uses the low drive power, low on resistance and fast turn off of the TMOS MTM15N50. The trip point is adjustable, LED fault indication is provided and battery power provides complete circuit isolation.

The two "circuit breaker" terminals are across one leg of a full wave diode bridge consisting of D1–D4. Normally, Q1 is turned ON so that the circuit breaker looks like a very low resistance. One input to comparator U1 is a fraction of the internal battery voltage and the other input is the drop across zeners D6 and D7 and the voltage drop across R1. If excessive current is drawn, the voltage drop across R1 increases beyond the comparator threshold (determined by the setting of R6), U1 output goes low, Q1 turns OFF, and the circuit breaker "opens." When this occurs, the LED fault indicator is illuminated.

12 ns CIRCUIT BREAKER



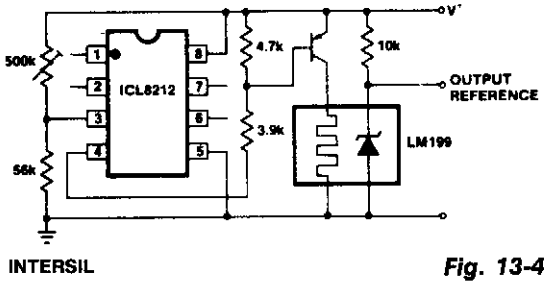
LINEAR TECHNOLOGY CORPORATION

Fig. 13-3

Circuit Notes

This circuit will turn off current in a load 12 ns after it exceeds a preset value. Under normal conditions the voltage across the 10 ohm shunt is smaller than the potential at the LT1016's negative input. This keeps Q1 off and Q2 receives bias, driving the load. When an overload occurs the current through the 10 ohm sense resistor begins to increase. When this current exceeds the preset value, the LT1016's outputs reverse. This provides ideal turn-on drive for Q1 and it cuts off Q2 in 5 ns. The delay from the onset of excessive load current to complete shutdown is just 13 ns. Once the circuit has triggered, the LT1016 is held in its latched state by feedback from the non-inverting output. When the load fault has been cleared the pushbutton can be used to reset the circuit.

LOW VOLTAGE POWER DISCONNECTOR



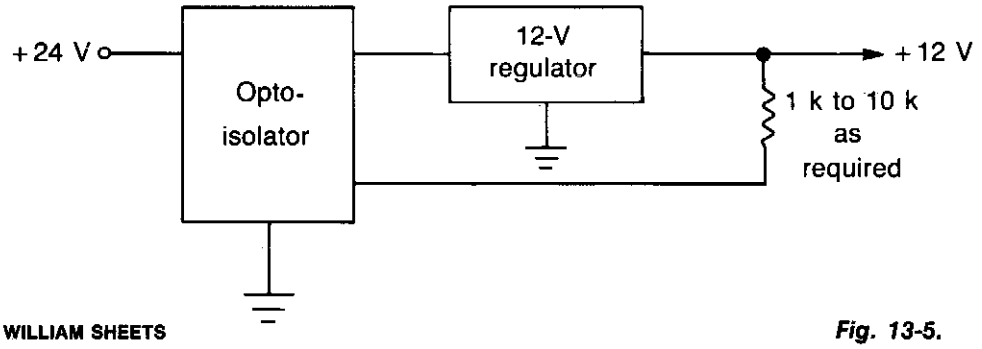
INTERSIL

Fig. 13-4

Circuit Notes

There are some classes of circuits that require the power supply to be disconnected if the power supply voltage falls below a certain value. As an example, the National LM199 precision reference has an on chip heater which malfunctions with supply voltages below 9 volts causing an excessive device temperature. The ICL8212 can be used to detect a power supply voltage of 9 volts and turn the power supply off to the LM199 heater section below that voltage.

AUTOMATIC POWER-DOWN PROTECTION CIRCUIT



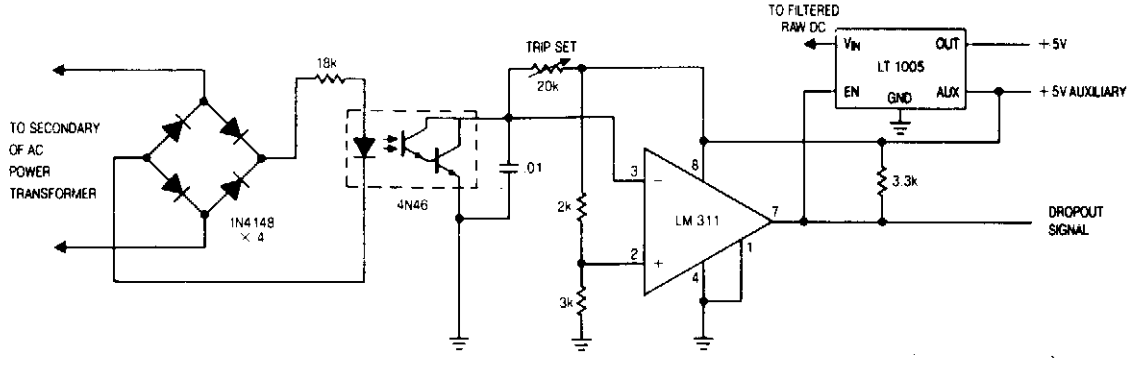
WILLIAM SHEETS

Fig. 13-5.

Circuit Notes

This circuit is faster than a fuse and automatically resets itself when a short is removed. The normal regulated dc input line is opened and the phototransistor of the opto isolator is connected in series with the source and regulator. Between the output of the regulator and ground is a LED and an associated current-limiting resistor, placed physically close to the surface of the photosensitive device. As long as the regulator is delivering its rated output, the LED glows and causes the photo device to have a low resistance. Full current is thus allowed to flow. If a short circuit occurs on the output side of the regulator, the LED goes dark, the resistance of the photo device increases, and the regulator shuts off. When the short is removed, the LED glows, and the regulator resumes operation.

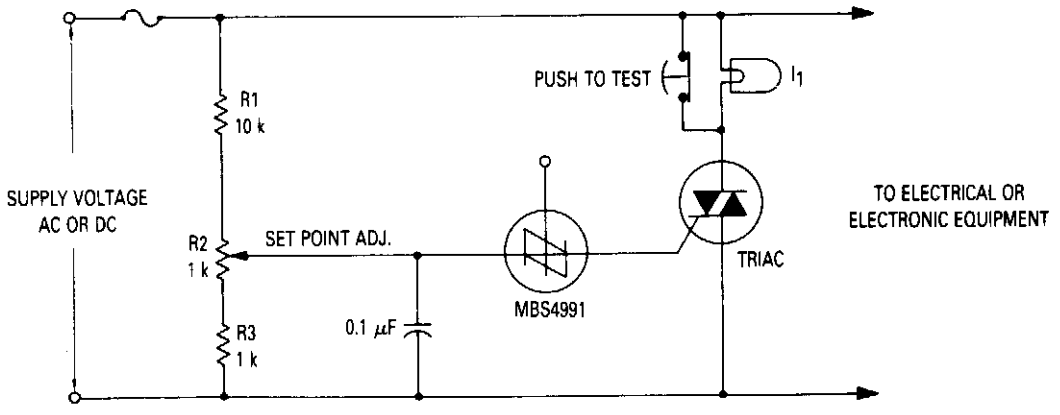
LINE DROPOUT DETECTOR



LINEAR TECHNOLOGY CORP.

Fig. 13-6

ELECTRONIC CROWBAR



MOTOROLA

Fig. 13-7

Circuit Notes

Where it is desirable to shut down equipment rather than allow it to operate on excessive supply voltage, an electronic "crowbar" circuit can be employed to quickly place a short-circuit across the power lines, thereby dropping the voltage across the protected device to near zero and blowing a fuse. Since the TRIAC and SBS are both bilateral devices, the circuit is equally useful on ac or dc supply lines. With the values shown for R1, R2 and R3, the crowbar operating point can be adjusted over the range of 60 to 120 volts dc or 42 to 84 volts ac. The resistor values can be changed to cover a different range of supply voltages. The voltage rating of the TRIAC must be greater than the highest operating point as set by R2. I_1 is a low power incandescent lamp with a voltage rating equal to the supply voltage. It may be used to check the set point and operation of the unit by opening the test switch and adjusting the input or set point to fire the SBS.

14

Clock Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Three Phase Clock From a Reference Clock
60 Hz Clock Pulse Generator

THREE PHASE CLOCK FROM A REFERENCE CLOCK

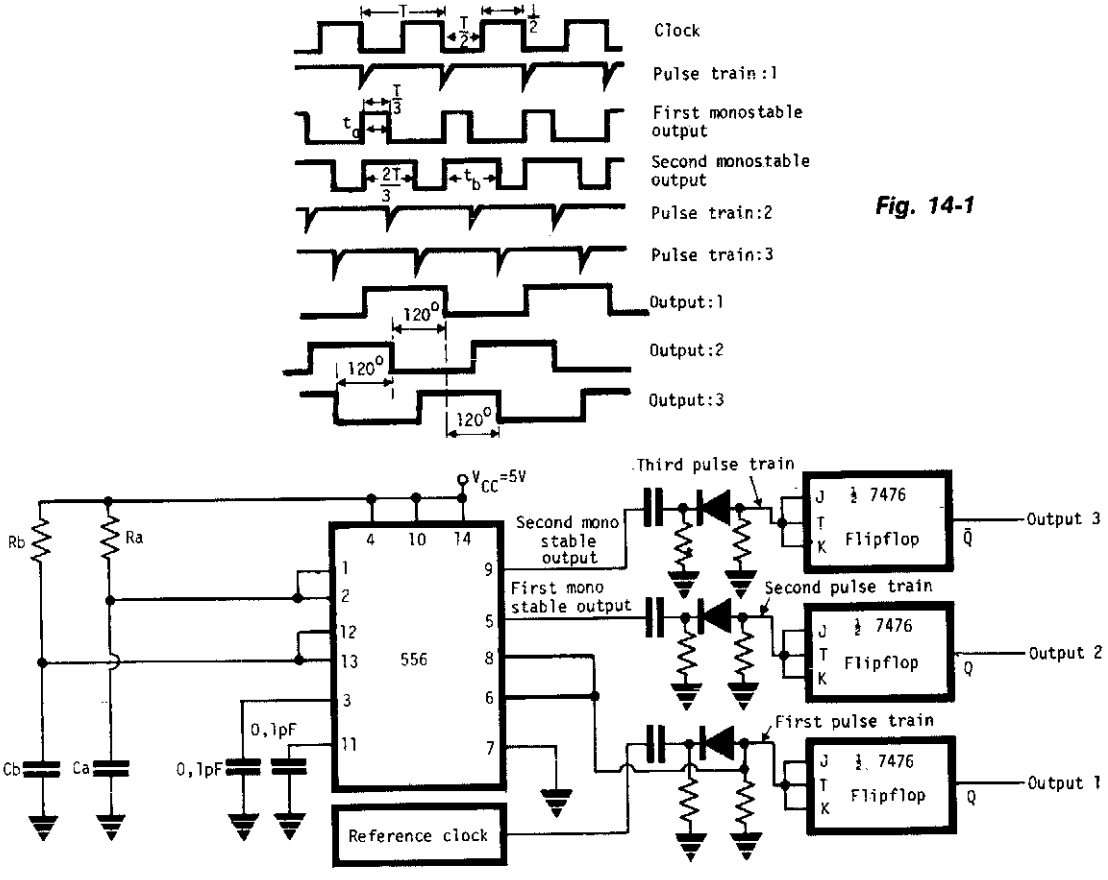


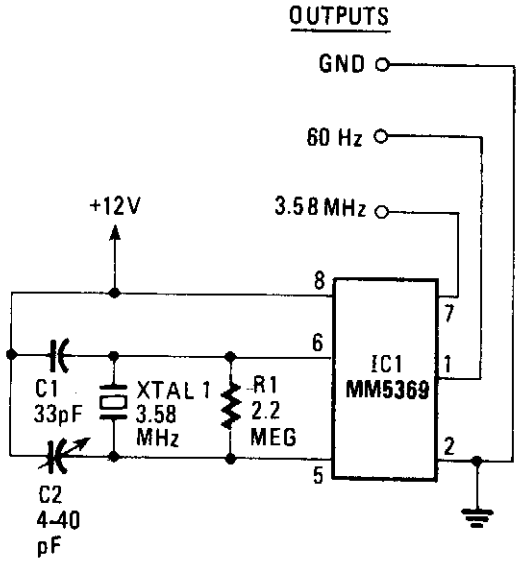
Fig. 14-1

ELECTRONIC ENGINEERING

Circuit Notes

The circuit provides three square wave outputs with 120° of phase difference between each other. Reference clock frequency is twice that of the required frequency. This can be obtained from a crystal oscillator with a chain of dividers or by using LM 555 in 50% duty cycle astable mode. If 1/T is the frequency of the reference clock, the dual timer 556 is connected to give two mono-stable output pulses of duration T/3 and 2T/3. The first timer R and C value are adjusted so that t_a = 1.1RaCa = T/3 and the second timer R and C values so that t_b = 1.1RbCb = 2T/3. For triggering the two monostables a negative pulse train (1st) is derived from the reference clock with a differentiator and a clipper combination as shown. The three pulse trains trigger three JK flip flops giving three phase square wave outputs.

60 Hz CLOCK PULSE GENERATOR



HANDS-ON ELECTRONICS

Fig. 14-2

Circuit Notes

The circuit provides a clean, stable square wave and it will operate on anywhere from 6 to 15 volts. The IC and color-burst crystal are the kind used in TV receivers. The 3.58 MHz output makes a handy marker signal for shortwave bands.

15

Comparators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low-Power Comparator with Less than 10 μV Hysteresis
Voltage Monitor/Comparator
Limit Comparator
Double-Ended Limit Comparator
Low-Cost Comparator and Display
Window Comparator
Comparator Detects Power Supply Overvoltages, Catches Glitches
High-Low Level Comparator with One Op Amp
High-Input-Impedance Window Comparator
Frequency Comparator
Demonstration Comparator Circuit
LED Frequency Comparator
TTL-Compatible Schmitt Trigger

LOW-POWER COMPARATOR WITH LESS THAN 10 μV HYSTERESIS

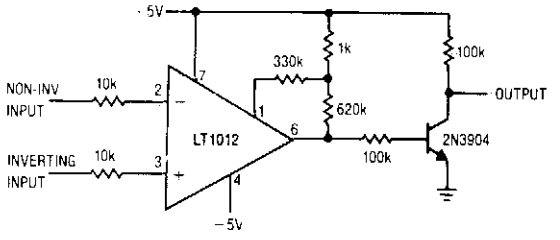
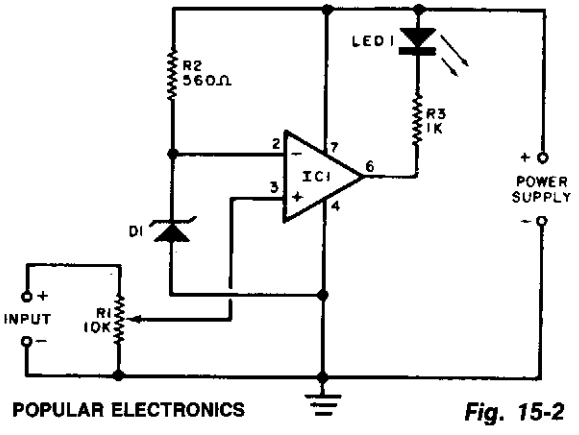


Fig. 15-1

LINEAR TECHNOLOGY CORP.

VOLTAGE MONITOR/COMPARATOR



POPULAR ELECTRONICS

Fig. 15-2

Circuit Notes

A portion of the monitored voltage (determined by R1's adjustment) is compared to a fixed voltage obtained from a zener reference network, R2-D1. As long as the monitored voltage remains at or above its present monitor point (determined by R1's setting), the output indicator, LED1, remains dark. If the voltage drops below this level, the LED goes on. D1 is a 3.3-V zener. A 12 Vdc power supply is suitable for monitoring input voltages of up to 12 volts.

LIMIT COMPARATOR

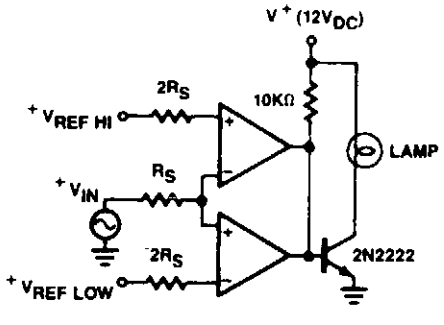


Fig. 15-3

SIGNETICS

DOUBLE-ENDED LIMIT COMPARATOR

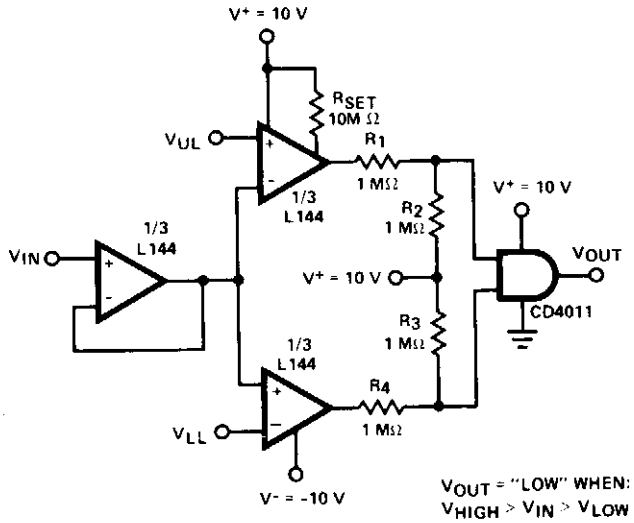


Fig. 15-4

SILICONIX P_D = 290 μW

LOW-COST COMPARATOR AND DISPLAY

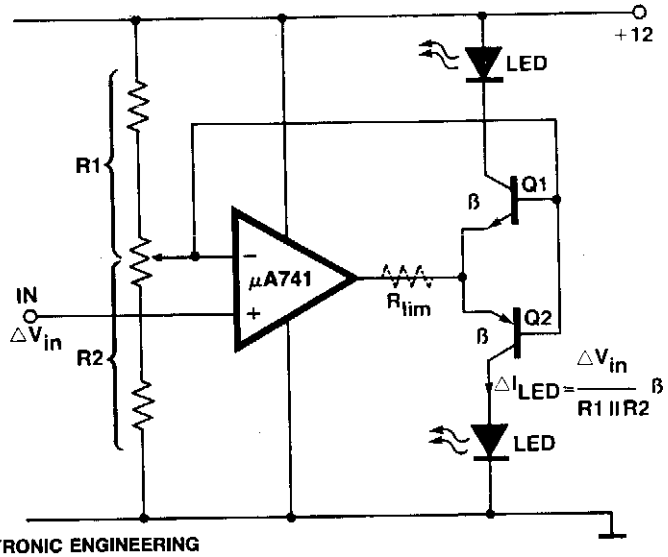


Fig. 15-5

ELECTRONIC ENGINEERING

Circuit Notes

An op amp is used as a comparator and a sink for LED current. The output voltage of the amplifier changes about 1.4 V depending on the direction of the current. Only one transistor is on at any time. Maximum LED current is limited to 25 mA by overcurrent protection of the uA741. If LEDs are not capable of carrying such a current or an alternative op amp is used and an additional resistor R_{lim} is necessary.

WINDOW COMPARATOR

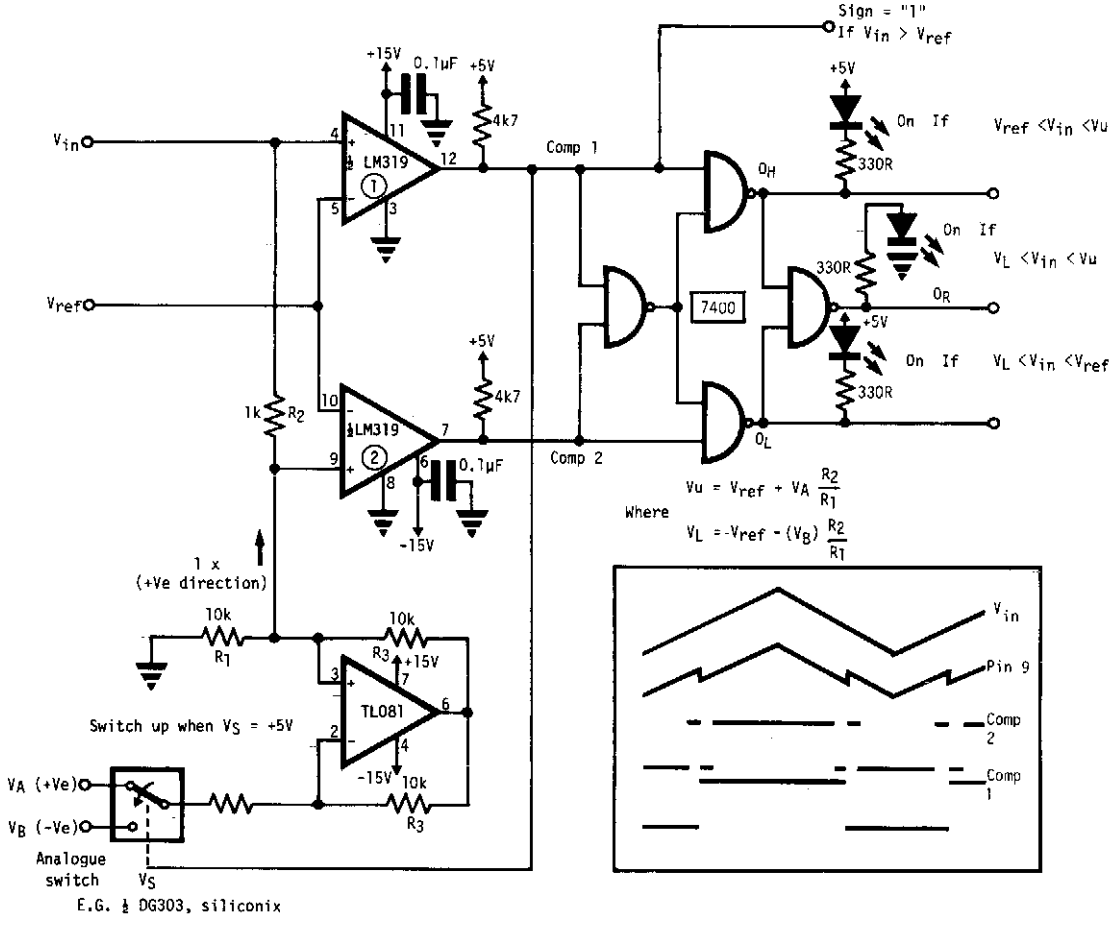


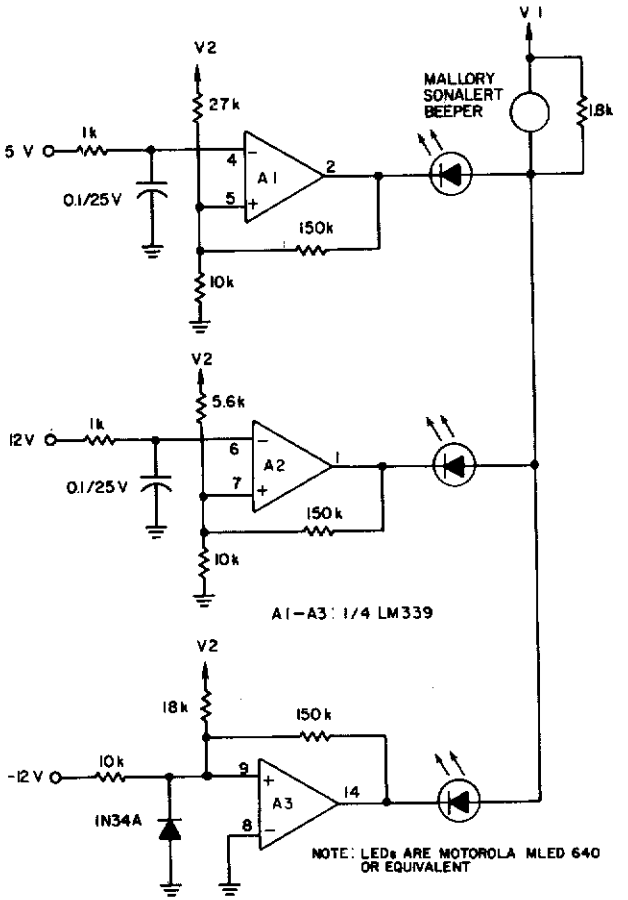
Fig. 15-6

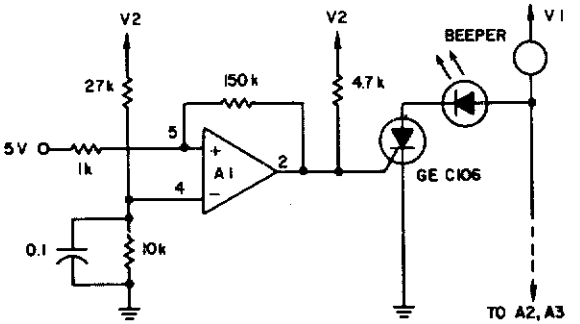
ELECTRONIC ENGINEERING

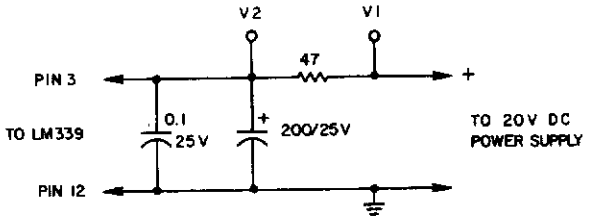
Circuit Notes

This circuit provides independently adjustable upper and lower threshold settings, and has sign, in window range, in upper window, and in lower window digital outputs.

COMPARATOR DETECTS POWER SUPPLY OVERVOLTAGES, CATCHES GLITCHES







ALL RESISTORS 1/4 W 5%

Fig. 15-7.

ELECTRONIC DESIGN

Circuit Notes

(A) To maintain an alarm condition when an overvoltage transient disappears, add an SCR to the comparator circuits. For SCR operation, voltages to the comparator inputs are inverted. (B) The triple-voltage monitoring circuit detects transient power-supply overvoltages. If excessive voltage momentarily appears at the 5, 12 and −12 V inputs, the LED for that circuit lights and the beeper sounds for as long as the overvoltage lasts.

HIGH-LOW LEVEL COMPARATOR WITH ONE OP AMP

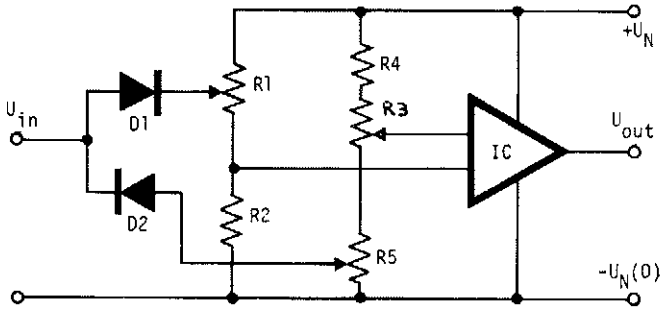


Fig. 15-8

ELECTRONIC ENGINEERING

Circuit Notes

The voltage to be compared is fed through diode D1 and D2 to the voltage dividers R1 and R5 where the low and high limits are present. When the voltage level of an input signal exceeds the high threshold limit set with potentiometer R1, the diode D1 becomes forward biased and the increased voltage on the inputs of the op amp drives it into positive saturation. Similarly, a decrease of the input voltage at the op amp inputs turns the op amp to positive saturation. Potentiometer R3 is used for zeroing the op amp in the off state.

HIGH-INPUT-IMPEDANCE WINDOW COMPARATOR

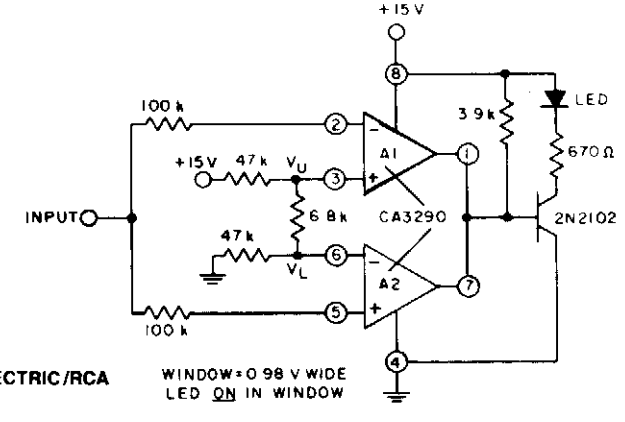


Fig. 15-9

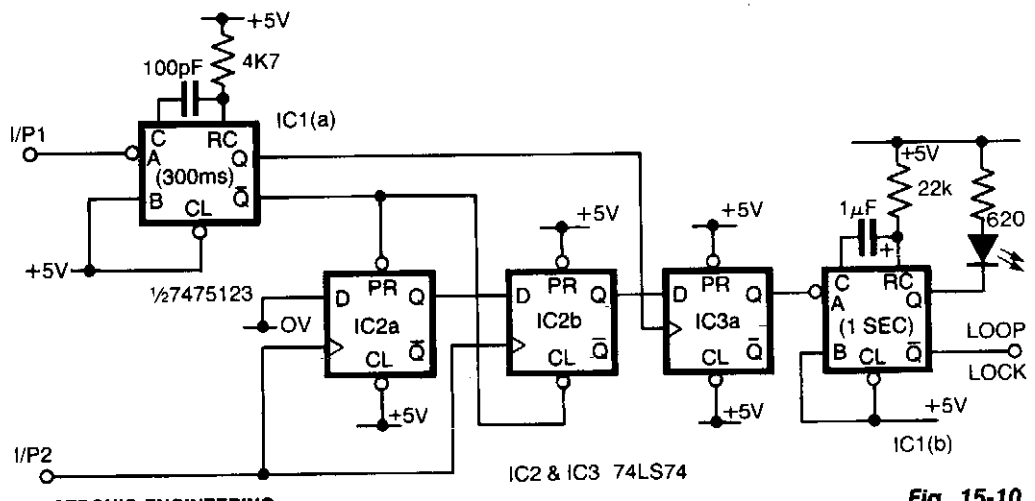
GENERAL ELECTRIC/RCA

WINDOW=0.98 V WIDE
LED ON IN WINDOW

Circuit Notes

The circuit uses both halves of the CA3290 BiMOS dual voltage comparator. The LED will be turned "ON" whenever the input signal is above the lower limit (V_L) but below the upper limit (V_U).

FREQUENCY COMPARATOR



ELECTRONIC ENGINEERING

Fig. 15-10

Circuit Notes

Input 1 is used as a gating period, during which a single rising edge on input 2 will cause a logic 1 output-any other number, indicating non-identical frequencies causes a logic 0 output.

IC1a converts input 1 to a narrow pulse which initializes IC2 which forms a two-stage shift register clocked by input 2. On the first edge of input 2 a logic 1 appears on the output of IC2b and for all subsequent inputs a logic 0 is present. At the end of the gating period this output is latched by IC3 forming the lock output. As this is only valid for one input period a monostable is added to the output to enable, for example, visual monitoring of the output. Either output from IC3 can be used depending on which state is most important. As connected the failure state is indicated.

DEMONSTRATION COMPARATOR CIRCUIT

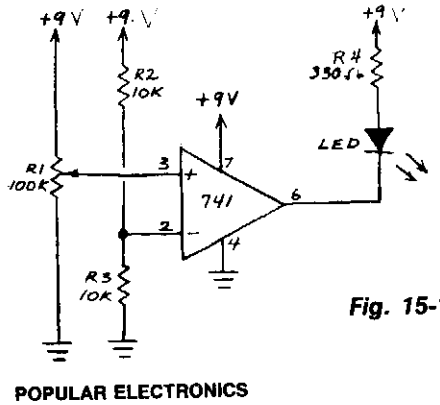


Fig. 15-11

POPULAR ELECTRONICS

Circuit Notes

This circuit is an op amp without a feedback resistor. R2 and R3 junction point sets the reference voltage. When the input voltage set by R1 is below the reference voltage the LED glows. If voltage is above reference, the LED goes off.

LED FREQUENCY COMPARATOR

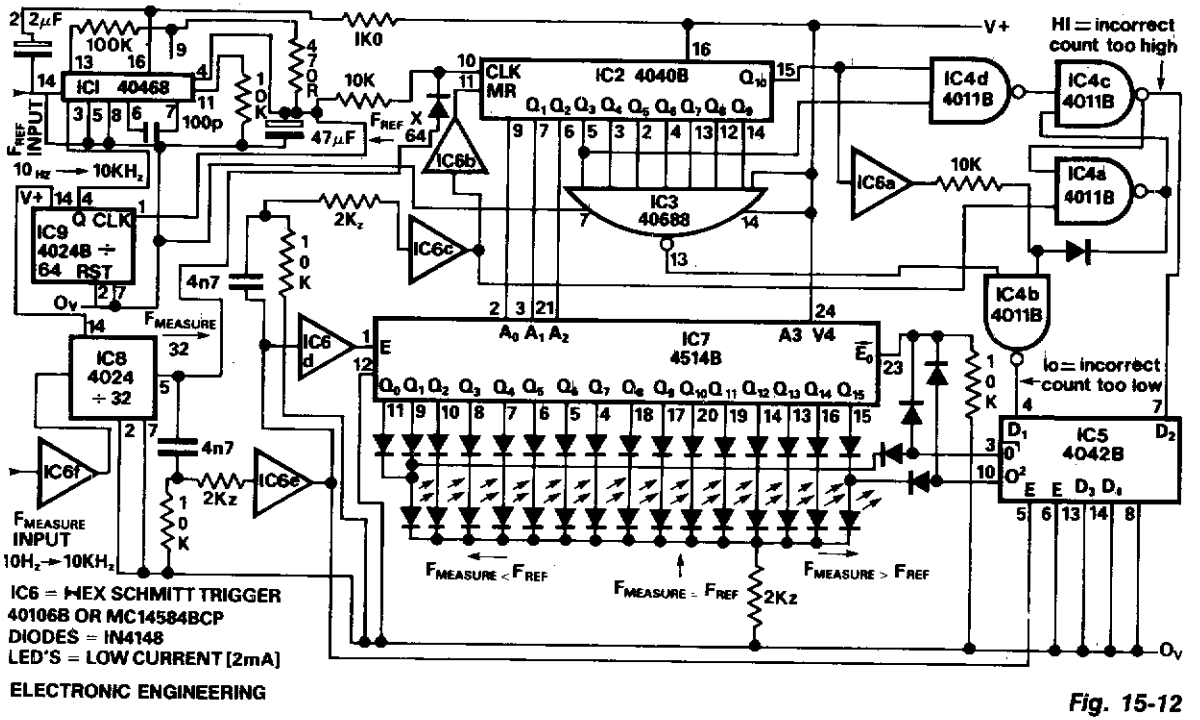


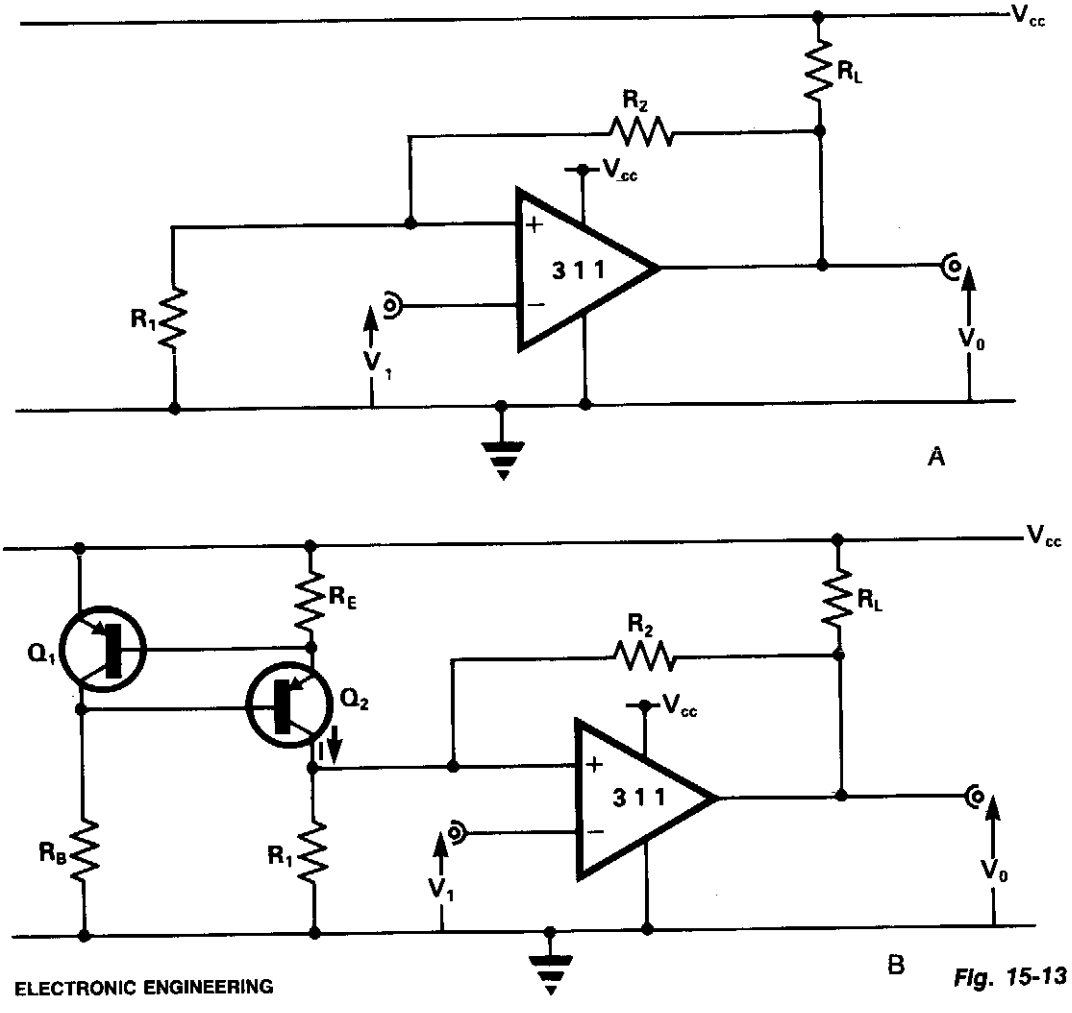
Fig. 15-12

Circuit Notes

The circuit provides unambiguous LED + or − bar readout with steps of 0.1%. The reference frequency is multiplied by the PLLIC1 and divider IC9 to output 64 × F (ref) and this is then gated by dividing F (measure) by 32 in IC8 thus is F (ref) = (measure) then IC2 counts 1024 pulses. Should the count be more than 1031 than the latch IC4c/IC4a is set to indicate count too high (F (measure) F (ref)) and if the count is less than 1017 then IC3/IC4b indicate count too low (F (measure) F (ref). These signals are latched by IC5 at the end of each period by the latch signal from IC6e.

When the two frequencies are within + or − 0.6% the LSB's of the counter IC2 are decoded and latched by IC7 and displayed on LED's IC6c resets the counter after latching the data.

TTL-COMPATIBLE SCHMITT TRIGGER



ELECTRONIC ENGINEERING

Fig. 15-13

Circuit Notes

The comparator has an output pull-up resistor R_L and is connected up to operate as a Schmitt trigger using the single rail supply V_{CC}. The feedback resistors R1 and R2 give upper and lower threshold levels V_{T+}, VVV_{TW}, respectively. V_{T+} is easily set by suitable resistor selection but there is little independent choice of V_{T-} because V_{T-} cannot exceed $V_{CE(SAT)}$. In Fig. 15-13B current-source, comprising the transistors R_E, R_B produces a current I ~: (V_{EB}/R_E), V_{EB} (~ 0.65 V) being the emitter-base voltage of Q1 and Q2. Fig. 15-13C shows the results of a practical test using the circuit of Fig. 15-13B, and the following operating and component data:

V_{CC} = 5 V; R_L = 1 K ohm; R_1 = R_2 = 10 K ohm;
R_B = 3.6 K ohm; R_E = 1 K ohm + 10 K ohm pot;
Q_1 = ZTX500; Q_2 = ZTX500.

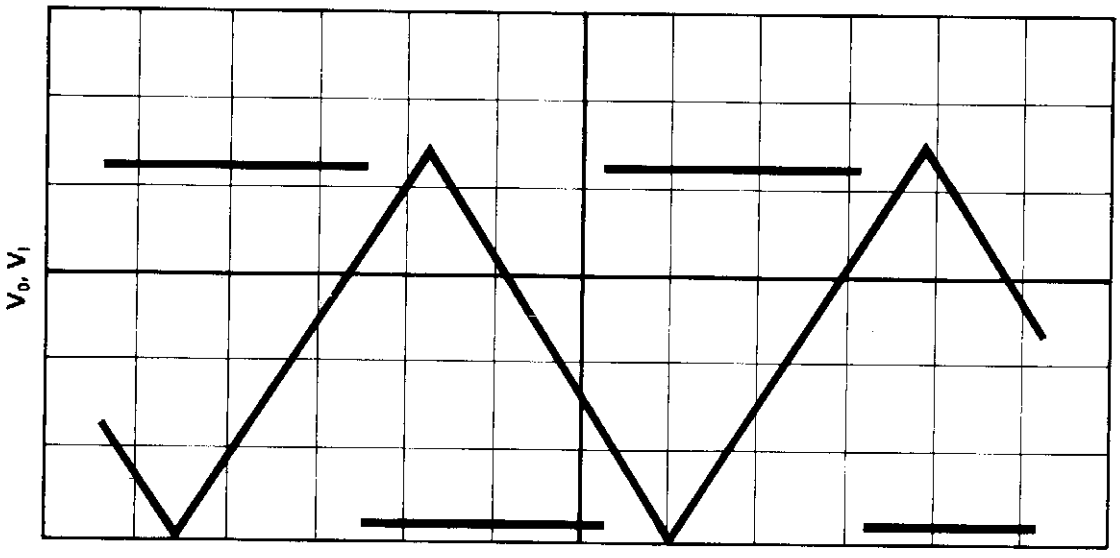


Fig. 15-13 Continued

C

16

Computer Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

8-Bit μP Bus Interface
V_{PP} Generator for Eproms
Eight Channel Mux/Demux System
Microprocessor Selected Pulse Width Control
8048/IM80C48 Microcomputer with 8-Character 16-Segment ASCII Triplex Liquid Crystal Display
CMOS Data Acquisition System
High Speed Data Acquisition System
Buffered Breakout Box
Z80 Clock
Data Separator for Floppy Disks

8-BIT μP BUS INTERFACE

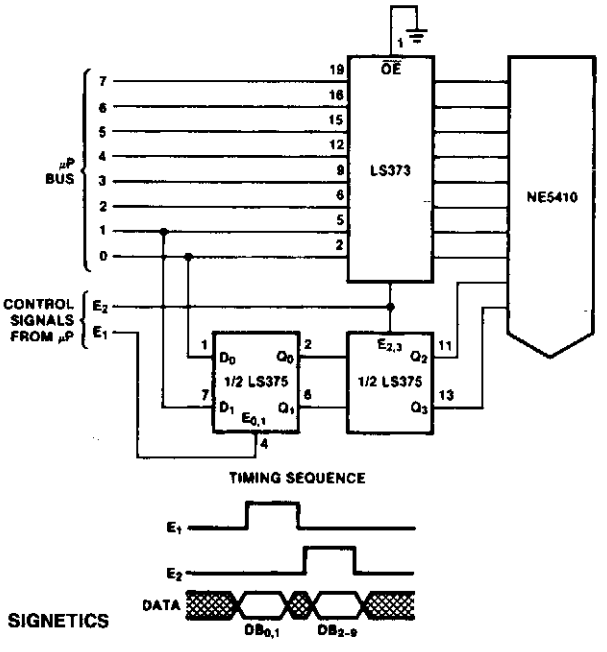
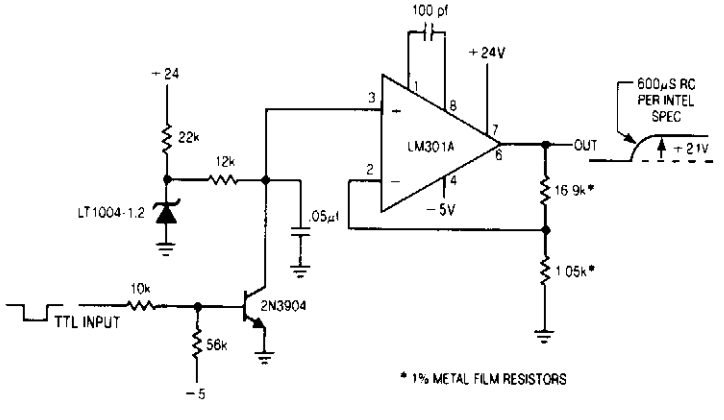


Fig. 16-1

Circuit Notes

With this double latch technique, valid data will be latched to the DAC until updated with the E_2 pulse. Timing will depend on the processor used.

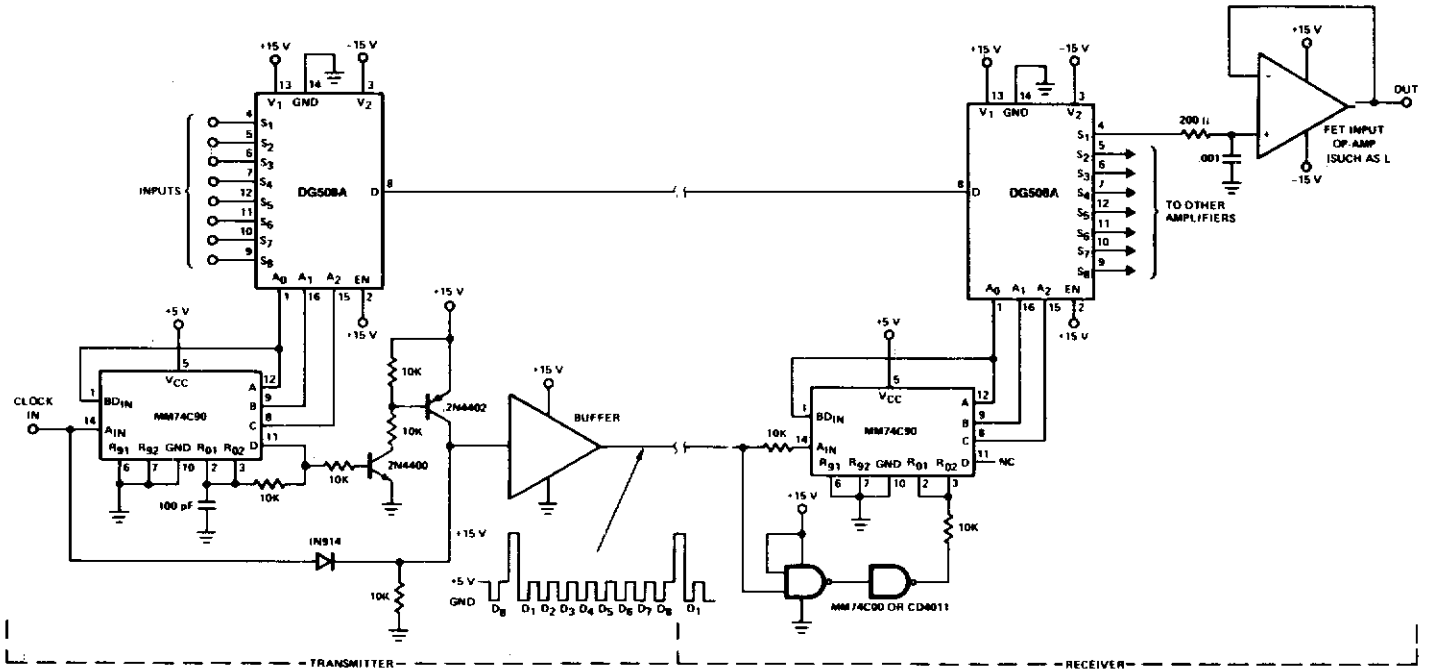
V_{PP} GENERATOR FOR EPROMS



LINEAR TECHNOLOGY CORP.

Fig. 16-2

EIGHT-CHANNEL MUX/DEMUX SYSTEM



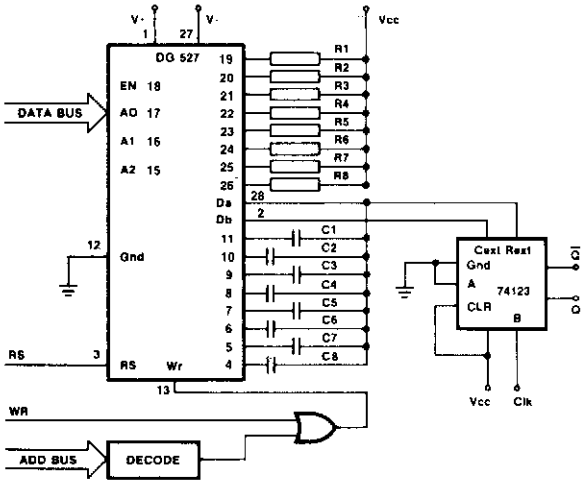
SILICONIX

Fig. 16-3

Circuit Notes

A number of signals may be sent between two points simultaneously by making a slight modification in the receiver circuit. A second DG508A is used as a demultiplexer, allowing all 8 channels to be monitored continuously.

MICROPROCESSOR SELECTED PULSE WIDTH CONTROL



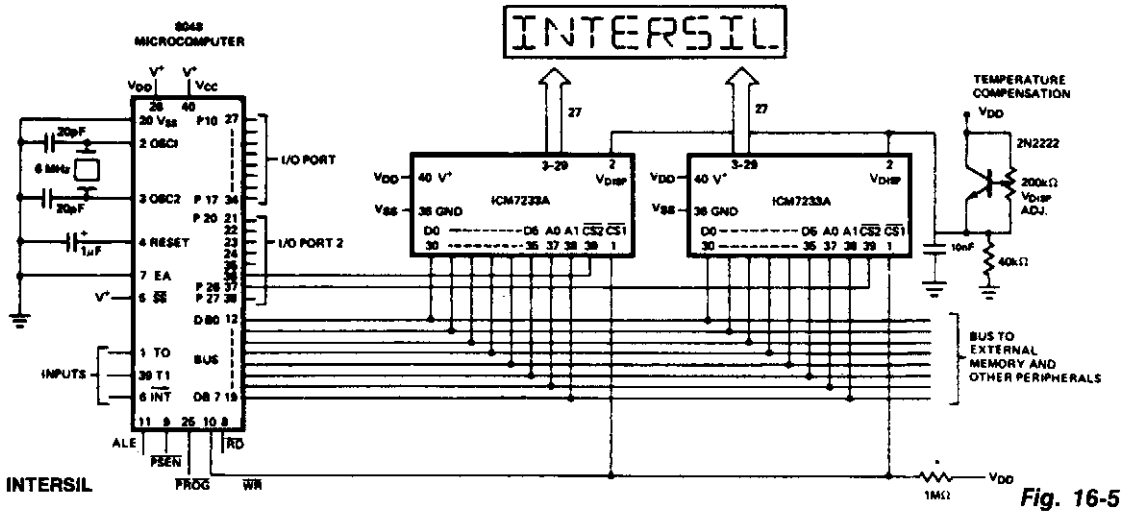
SILICONIX

Fig. 16-4

Circuit Notes

Differential multiplexers are generally used in process control applications to eliminate errors due to common mode signals. In this circuit however, advantage is taken of the dual multiplexing capability of the switch. This is achieved by using the multiplexer to select pairs of RC networks to control the pulse width of the multivibrator. This can be a particularly useful feature in process control applications where there is a requirement for a variable width sample "window" for different control signals.

8048/IM80C48 MICROCOMPUTER WITH 8-CHARACTER 16-SEGMENT ASCII TRIPLEX LIQUID CRYSTAL DISPLAY



INTERSIL

Fig. 16-5

Circuit Notes

The two bit character address is merged with the data and written to the display driver under the control of the WR line. Port lines are used to either select the target driver, or deselect all of them for other bus operations.

CMOS DATA ACQUISITION SYSTEM

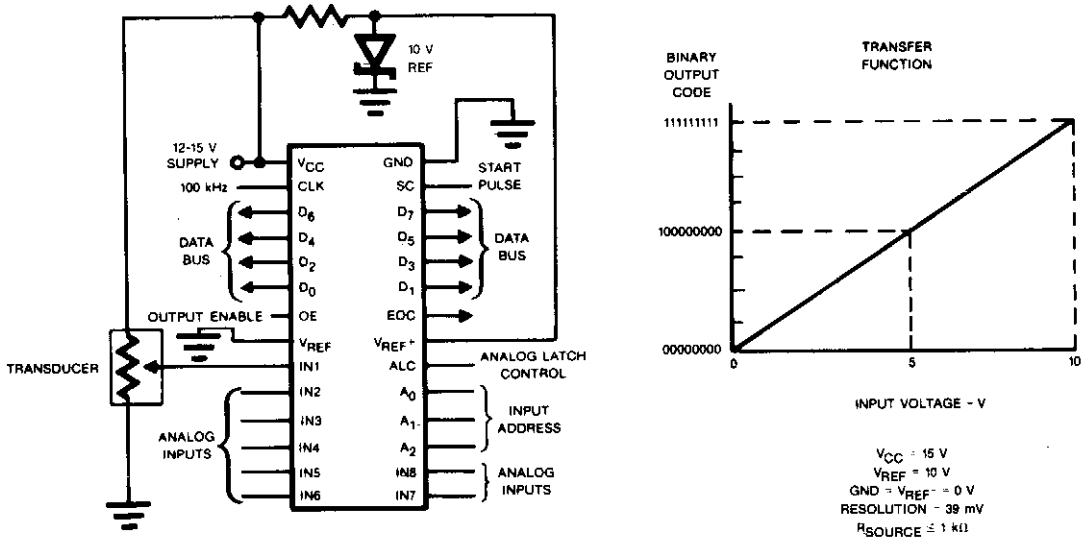


Fig. 16-6

SILICONIX

Circuit Notes

Charge redistribution to achieve A/D conversion. In typical applications, as a ratiometric conversion system for a microprocessor, V_{REF-} will be connected to ground and V_{REF+} will be connected to V_{CC}. The output will then be a simple proportional ratio between analog input voltage and V_{CC}. The general relationship is:

$$\frac{D_{OUT}}{2^8} = \frac{V_{IN}}{V_{REF+} - V_{REF-}}$$

Where D_{OUT} = Digital Output
V_{IN} = Analog Input
V_{REF} = Positive Reference Potential
V_{REF} = Negative Reference Potential

HIGH SPEED DATA ACQUISITION SYSTEM

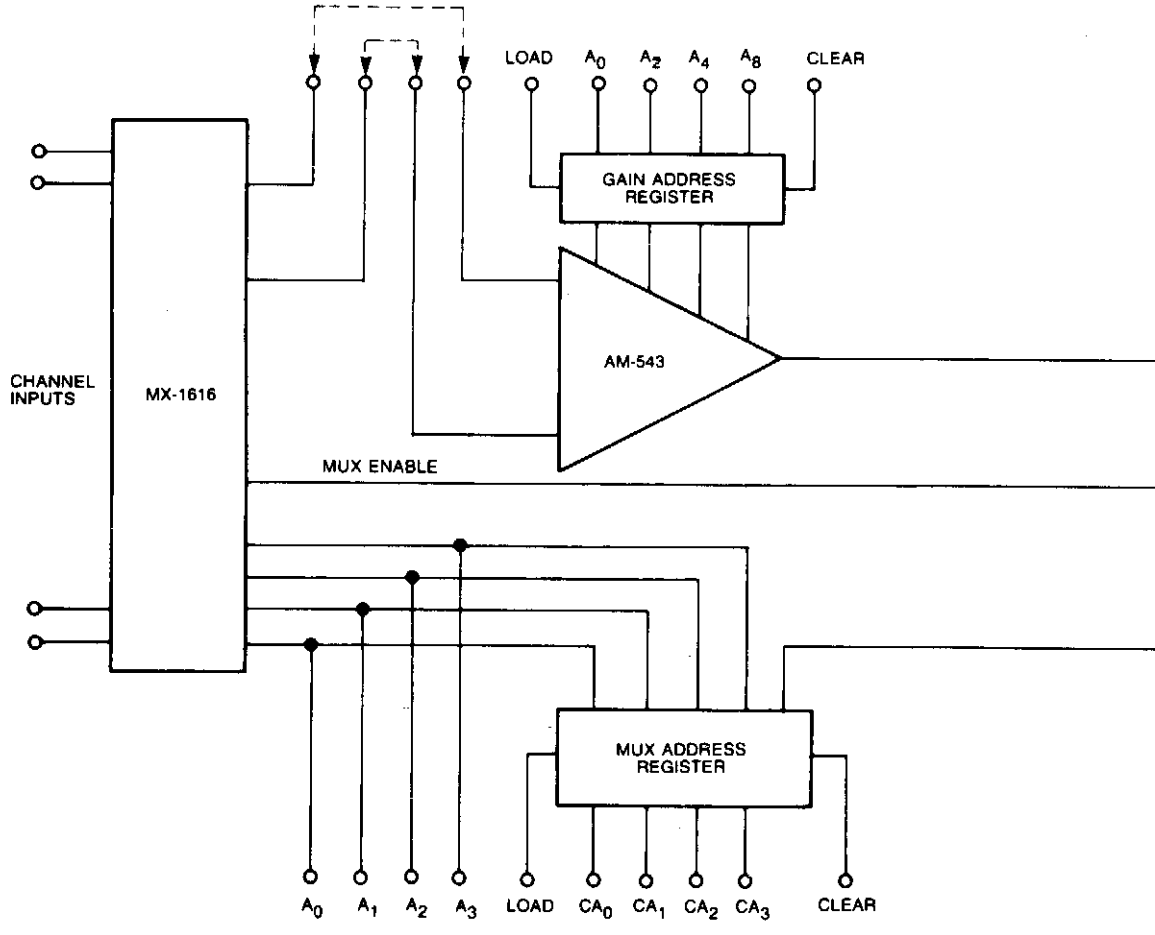
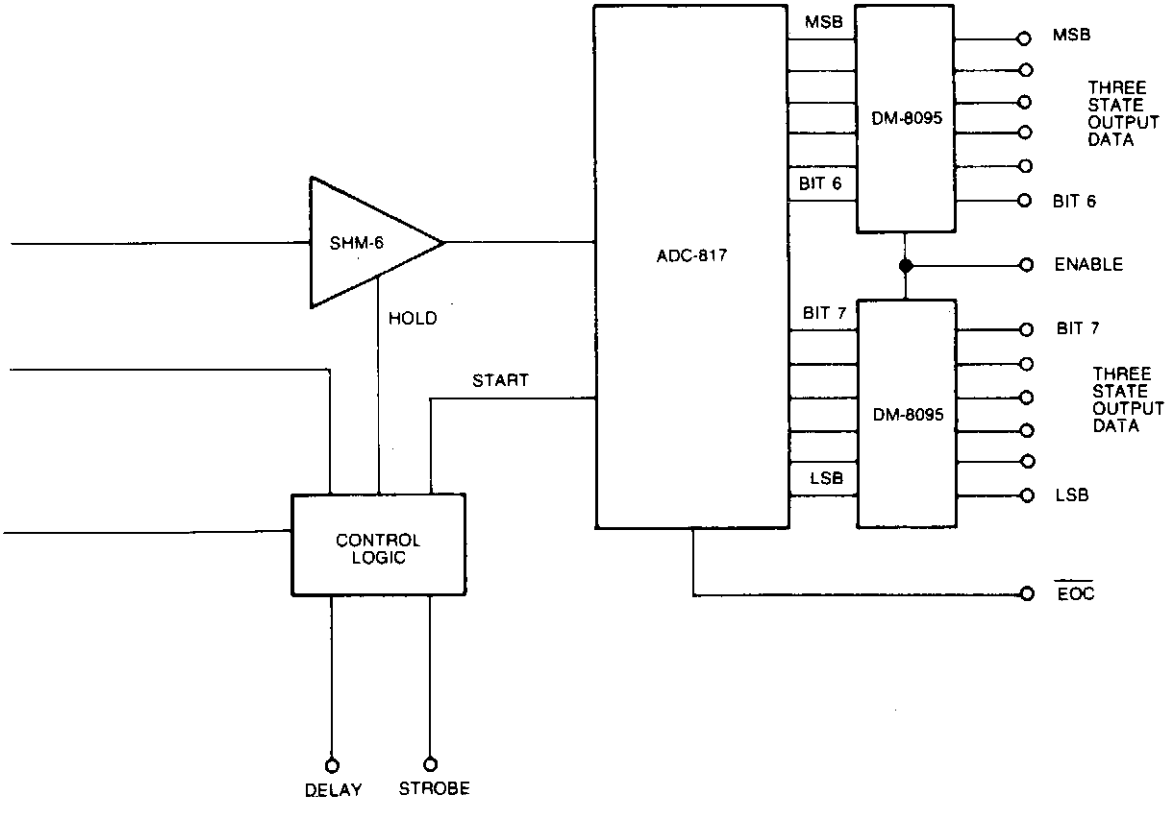


Fig. 16-7

Circuit Notes

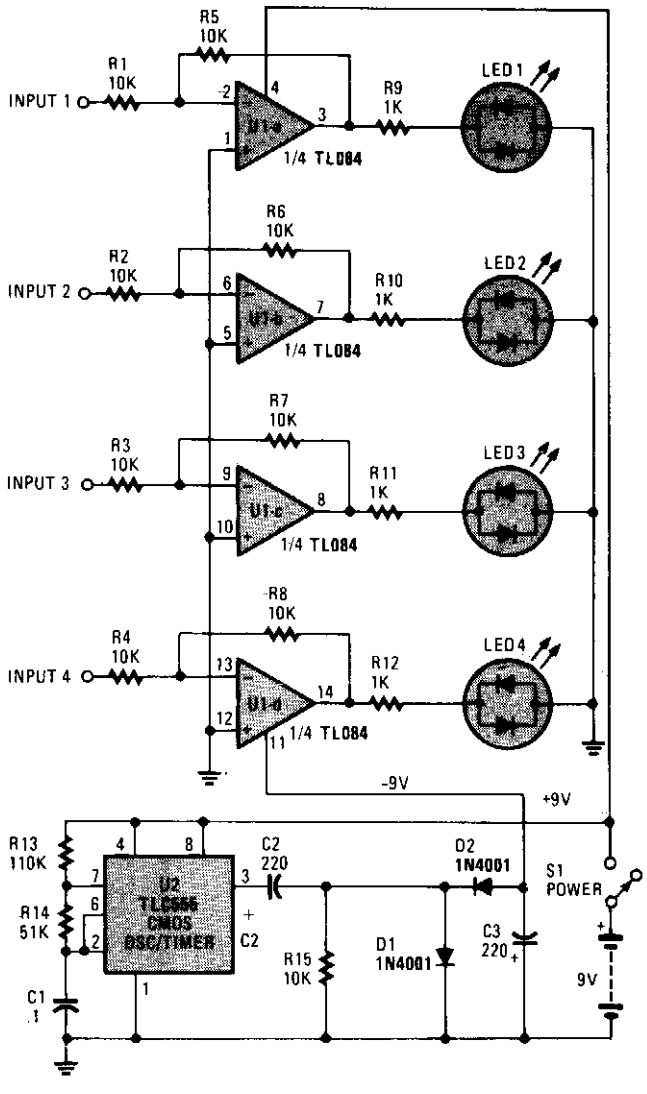
This diagram shows a high-speed data acquisition system with 8 differential inputs and 12-bit resolution using the AM-543. If the control logic is timed so that the Sample-Hold-ADC section is converting one analog value while the mux-amplifier section is allowed to settle to the next input value, throughout rates greater than 156 KHz can be achieved. The AM-543 is used with Datel's ADV-817, a 12-bit hybrid A/D with a 2 μsec conversion rate, the SHM-6, a 0.01%, 1 μsec hybrid Sample-Hold, and the MX-1616,



a low cost, high-speed monolithic analog multiplexer. The system works as follows:

The μP selects a channel and initiates a conversion at G=1 and then looks at the MSB of the conversion result. If the MSB = 1, the μP will store the value. If the MSB = O, the μP will select G = 2. The μP will repeat the cycle of gain incrementing, comparison, and analog-to-digital conversion until the MSB = 1. The μP will then test for an output of all 1's, as this is the full-scale output of the A/D. If the output is all 1's, the μP will decrement the gain by 1 step and perform the final conversion.

BUFFERED BREAKOUT BOX



HANDS-ON ELECTRONICS

Fig. 16-8

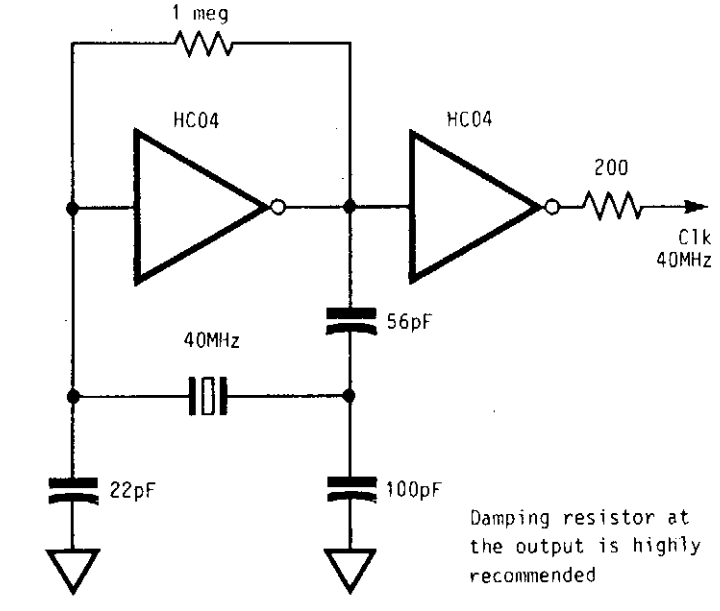
Circuit Notes

The monitoring circuit consists of four tri-color LEDs driven by an equal number of op amps configured as gain-of-one inverting amplifiers. Each LED is wired in the circuit so that it glows red when the input to the op amp is high, and green when the input is low. The LED remains off when the input is disconnected from a circuit, when it's at ground potential, and when it's connected to a 3-state output that's in the high-impedance state. Each input has an impedance of 10,000 ohms preventing the circuit

BUFFERED BREAKOUT BOX, Continued.

from loading communication lines. The op amp requires both positive and negative supply voltages to properly drive the LEDs. Both voltages are supplied by a single, nine-volt battery. The battery supplies the positive source directly. The negative source is supplied via a CMOS 555 oscillator/timer that's configured as an astable oscillator, which is used to drive a standard diode/capacitor voltage doubler. When the 555 is connected to the monitoring circuit, the output voltage is not 18 volts (2 × 9), but a little under nine volts, due to loading. The circuit draws about 16 mA with all LEDs off; with all four on, it draws between 20 and 30 mA, depending on how many LEDs are high, and how many are low. The use of CMOS op amps reduces quiescent current drain considerably.

Z80 CLOCK



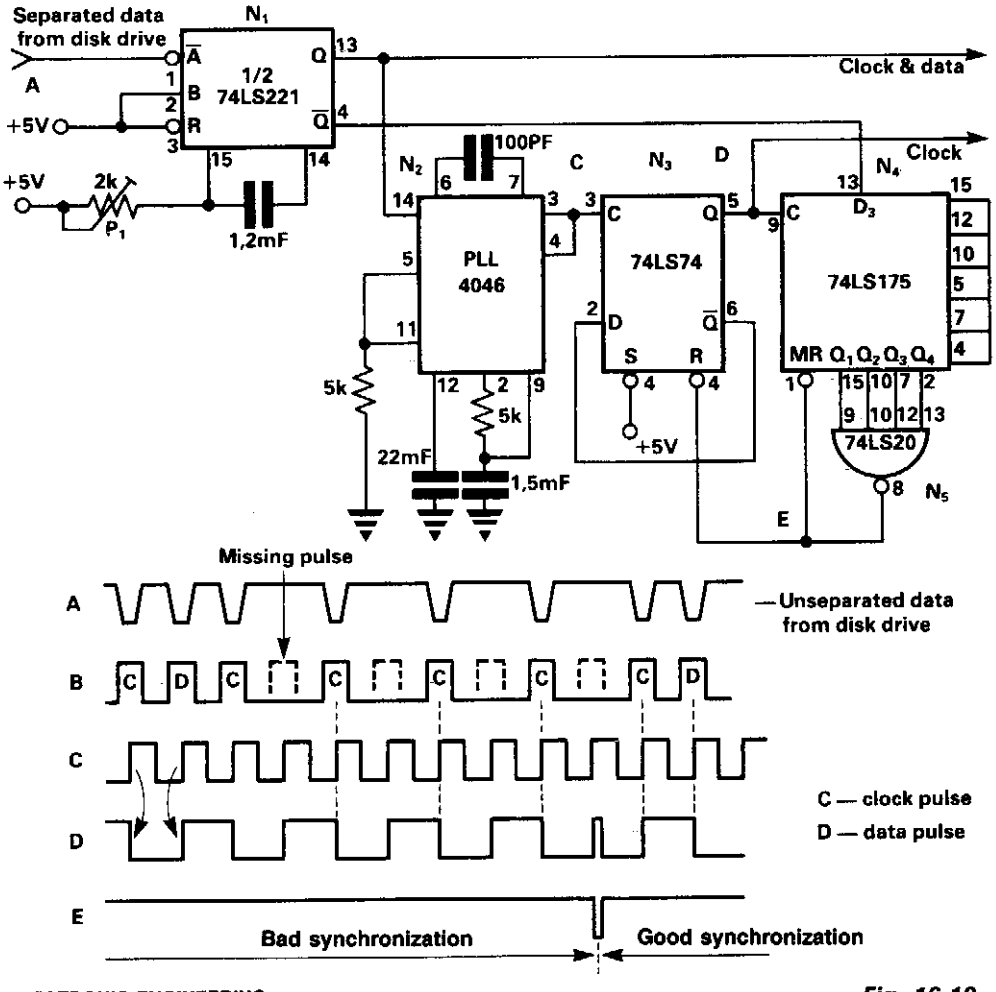
ELECTRONIC ENGINEERING

Fig. 16-9

Circuit Notes

The circuit will operate reliably from below 1 MHz to above 400 MHz. With V_{CC} = 5 V the output of the second inverter essentially attains a full swing from 0 V to 5 V. Such large logic output levels and broad frequency range capabilities make this oscillator quite suitable for driving MOS components such as CPU, controller chip, peripheral devices, as well as other TTL products. A damping resistor in series between the clock output of the oscillator and the input of the device being driven will remove the undesirable undershoot and ringing caused by the high speed CMOS part.

DATA SEPARATOR FOR FLOPPY DISKS



ELECTRONIC ENGINEERING

Fig. 16-10

Circuit Notes

The data separator is intended for use with 8" flexible diskettes with IBM 3870 soft sectored format. The circuit delivers data and clock (B) and clock pulses (D). These two signals must be in such a sequence that the negative edge of the clock pulse is at the middle of a data cell.

Unseparated data (A) from the floppy unit is shaped with one shot N1. Trimmer P1 should be adjusted so that pulses (B) are 1 μs wide. This signal synchronizes PLL N2 with a free running frequency adjusted to 500 kHz. The output of the PLL is 90° out of phase with its input. D-type flip-flop N3 is connected as a divider by two and changes state at each positive edge of (C). N4, connected as a shift register, looks for four consecutive missing pulses. When this happens, the circuit is resynchronized with (E) so that the negative edge of (D) is in the middle of a data cell.

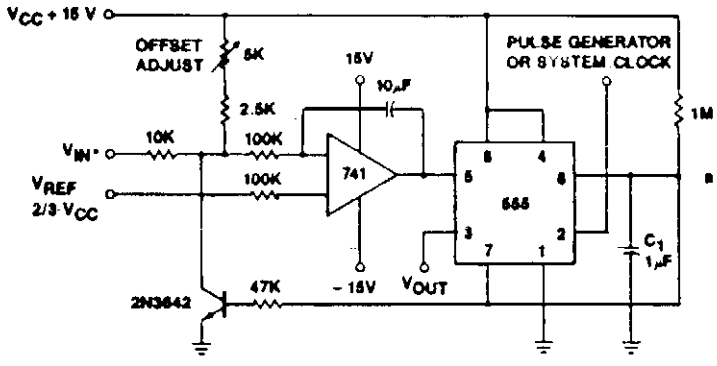
17

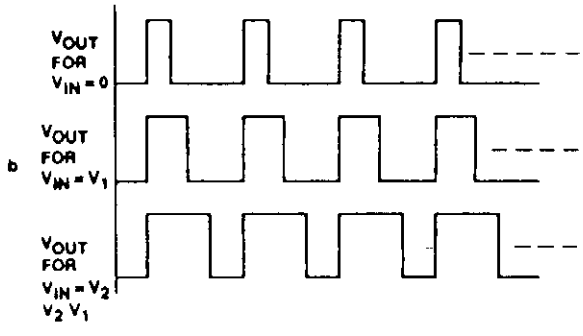
Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Voltage-to-Pulse Duration Converter
Voltage-to-Current Converters
TTL-to-MOS Logic Converter
TTL Square Wave-to-Triangle Converter
A Regulated DC-to-DC Converter
Capacitance to Pulse Width Converter
Current-to-Voltage Converter with Grounded Bias and Sensor
Triangle-to-Sine Converters
Precision Peak-to-Peak AC-DC Converter
Photodiode Current-to-Voltage Converter
Self Oscillating Flyback Converter
RMS-to-DC Converter
100 MHz Converter
Precision Voltage-to-Frequency Converter
Bipolar DC-DC Converter Requires No Inductor

VOLTAGE-TO-PULSE DURATION CONVERTER





NOTES:
All resistor values in ohms
*V_{IN} is limited to 2 diode drops within ground or below V_{CC}

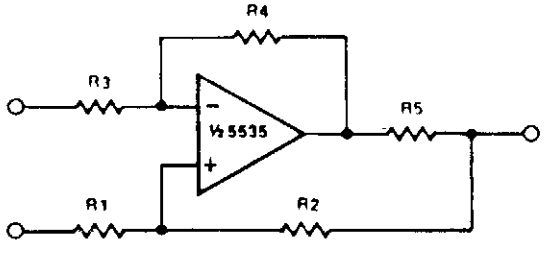
SIGNETICS

Fig. 17-1.

Circuit Notes

Voltage levels can be converted to pulse durations by combining an op amp and a timer IC. Accuracies to better than 1% can be obtained with this circuit (a), and the output signals (b) still retain the original frequency, independent of the input voltage.

VOLTAGE-TO-CURRENT CONVERTERS



NOTES:

$$\frac{R_2}{R_1} = \frac{R_4}{R_3}$$

$$I_{OUT} = \frac{V_{IN} \cdot R_2}{R_5 \cdot R_1}$$

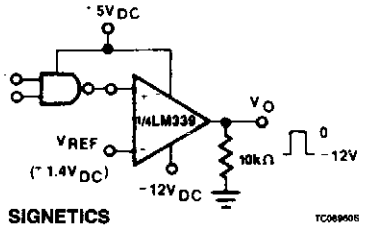
SIGNETICS

Circuit Notes

A simple voltage-to-current converter is shown in the figure. The current out is I_{OUT} or V_{IN}/R. For negative currents, a pnp can be used and, for better accuracy, a Darlington pair can be substituted for the transistor. With careful design, this circuit can be used to control currents of many amps. Unity gain compensation is necessary.

Fig. 17-2

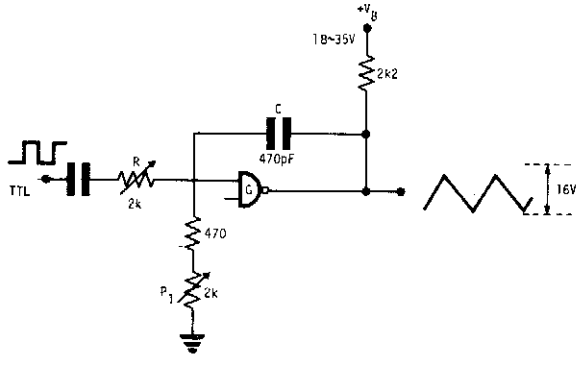
TTL-TO-MOS LOGIC CONVERTER



SIGNETICS

Fig. 17-3

TTL SQUARE WAVE-TO-TRIANGLE CONVERTER



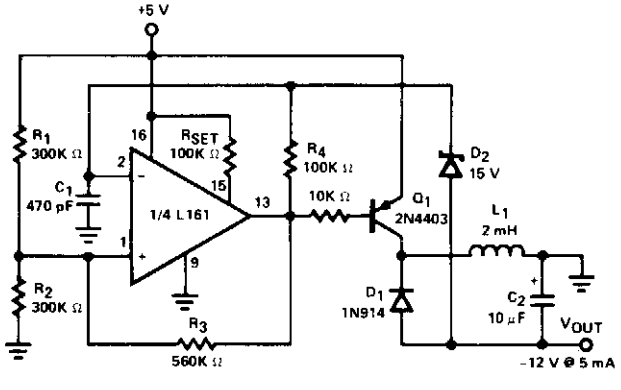
ELECTRONIC ENGINEERING

Fig. 17-4

Circuit Notes

This fixed frequency triangular waveform generator driven by a TTL square wave generates typically 16-V p-p triangles at frequencies up to several MHz. It uses only one NAND open collector gate, or one open collector inverter as a fast integrator with gain. Careful successive adjustments of R and P1 are needed. When correct adjustments are reached, output amplitude and linearity are largely independent of the value of V_B, from a minimum of 18 V up to 35 V. The value of C shown is for 100 kHz; at higher frequencies, it must be reduced in proportion.

A REGULATED DC-TO-DC CONVERTER



SILICONIX

Fig. 17-5

CAPACITANCE TO PULSE WIDTH CONVERTER

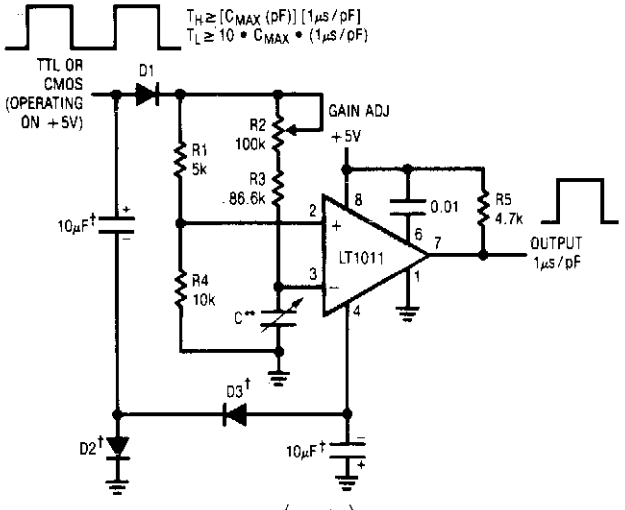


Fig. 17-6

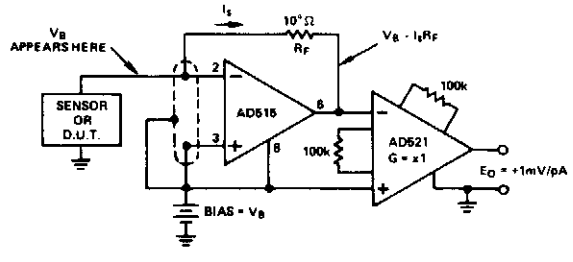
*$PW = (R2 + R3)(C)\left(\frac{R1 + R4}{R1}\right)$, INPUT CAPACITANCE OF LT1011 IS ≈6pF. THIS IS AN OFFSET TERM.

†THESE COMPONENTS MAY BE ELIMINATED IF NEGATIVE SUPPLY IS AVAILABLE (−1V TO −15V).

**TYPICAL 2 SECTIONS OF 365pF VARIABLE CAPACITOR WHEN USED AS SHAFT ANGLE INDICATION.

LINEAR TECHNOLOGY CORP.

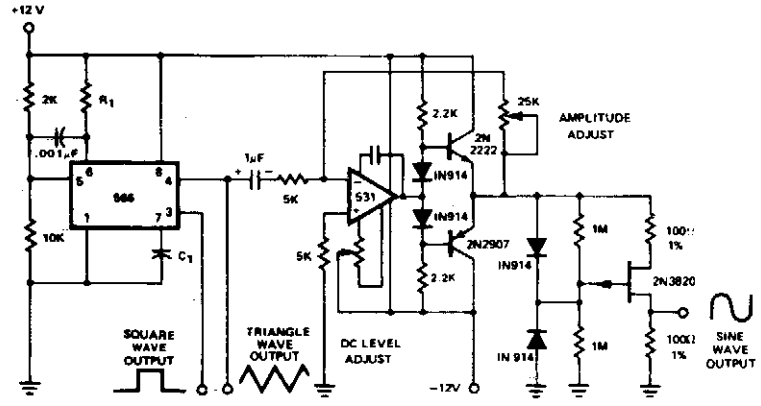
CURRENT-TO-VOLTAGE CONVERTER WITH GROUNDED BIAS AND SENSOR

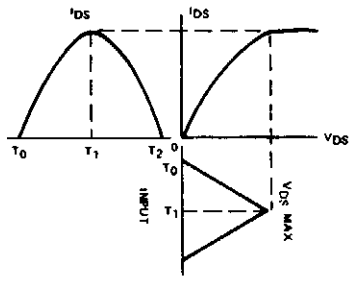


ANALOG DEVICES, INC.

Fig. 17-7

TRIANGLE-TO-SINE CONVERTERS





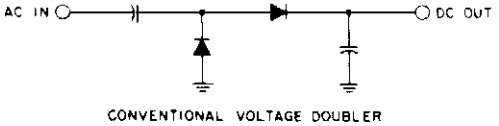
SIGNETICS

Fig. 17-8

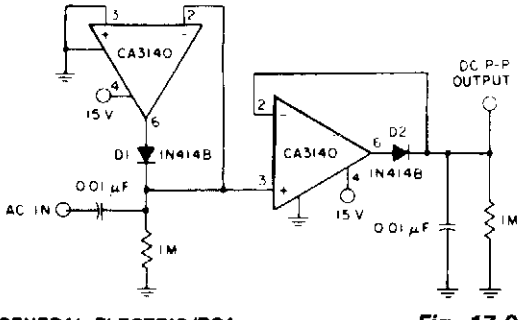
Circuit Notes

Conversion of triangle wave shapes to sinusoids is usually accomplished by diode-resistor shaping networks, which accurately reconstruct the sine wave segment by segment. Two simpler and less costly methods may be used to shape the triangle waveform of the 566 into a sinusoid with less than 2% distortion. The non-linear $I_{DS}V_{DS}$ transfer characteristic of a P-channel junction FET is used to shape the triangle waveform. The amplitude of the triangle waveform is critical and must be carefully adjusted to achieve a low distortion sinusoidal output. Naturally, where additional waveform accuracy is needed, the diode-resistor shaping scheme can be applied to the 566 with excellent results since it has very good output amplitude stability when operated from a regulated supply.

PRECISION PEAK-TO-PEAK AC-DC CONVERTER



CONVENTIONAL VOLTAGE DOUBLER



GENERAL ELECTRIC/RCA

Fig. 17-9

Circuit Notes

Using a CA3140 BiMOS op amp and a single positive supply converts a conventional voltage doubler with two precision diodes into a precision peak-to-peak ac-to-dc voltage converter having wide dynamic range and wide bandwidth.

ALL RESISTANCE VALUES ARE IN OHMS

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PHOTODIODE CURRENT-TO-VOLTAGE CONVERTER

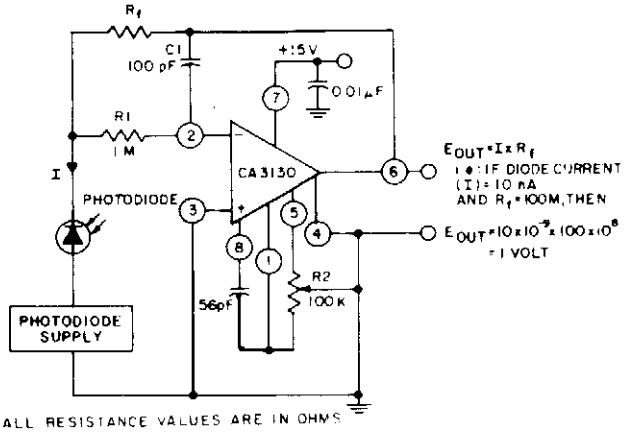


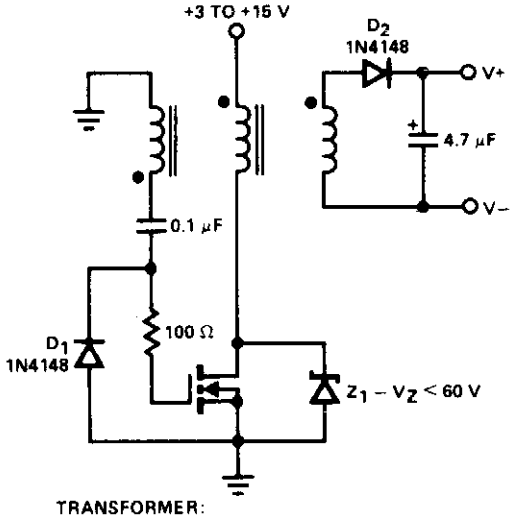
Fig. 17-10

GENERAL ELECTRIC/RCA

Circuit Notes

The circuit uses three CA3130 BiMOS op amps in an application sensitive to sub-picoampere input currents. The circuit provides a ground-referenced output voltage proportional to input current flowing through the photodiode.

SELF OSCILLATING FLYBACK CONVERTER



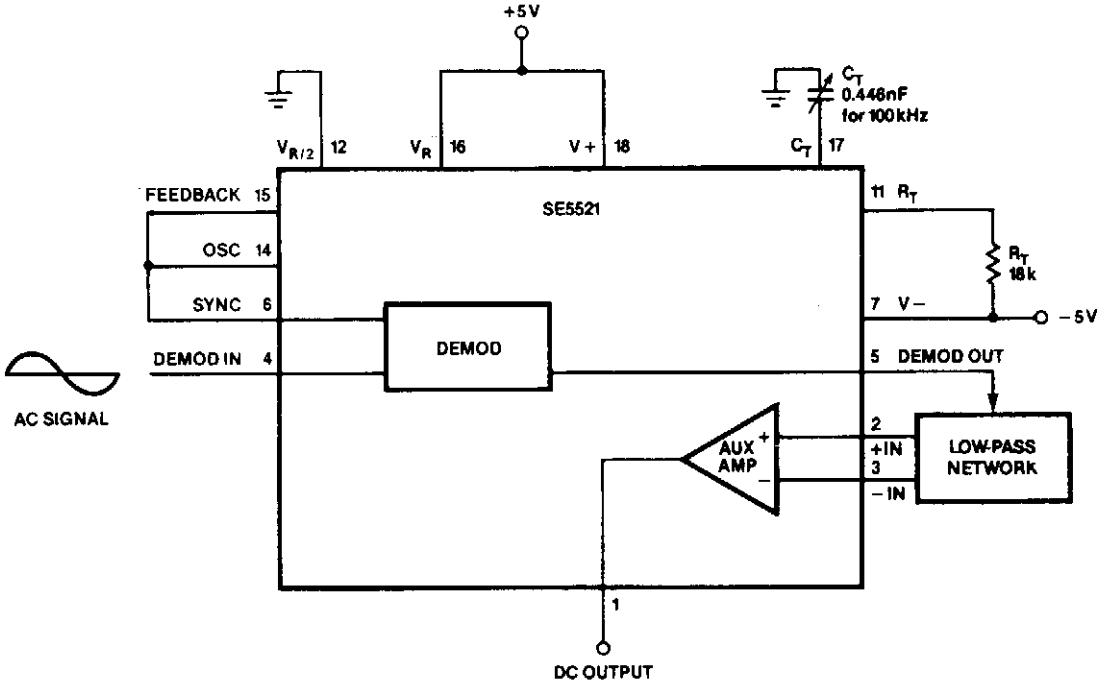
Circuit Notes

Low-power converter uses the core characteristics to determine frequency. With the transformer shown, operating frequency is 250 kHz. Diode D1 prevents negative spikes from occurring at the MOSFET gate, the 100 ohm resistor is a parasitic suppressor, and Z1 serves as a dissipative voltage regulator for the output and also clips the drain voltage to a level below the rated power FET breakdown voltage.

SILICONIX, INC.

Fig. 17-11

RMS-TO-DC CONVERTER



NOTE:
1. The DC output at Pin 1 varies linearly with the RMS input at Pin 4.
2. C_T is tweaked until the sync signal is in phase with the AC signal.

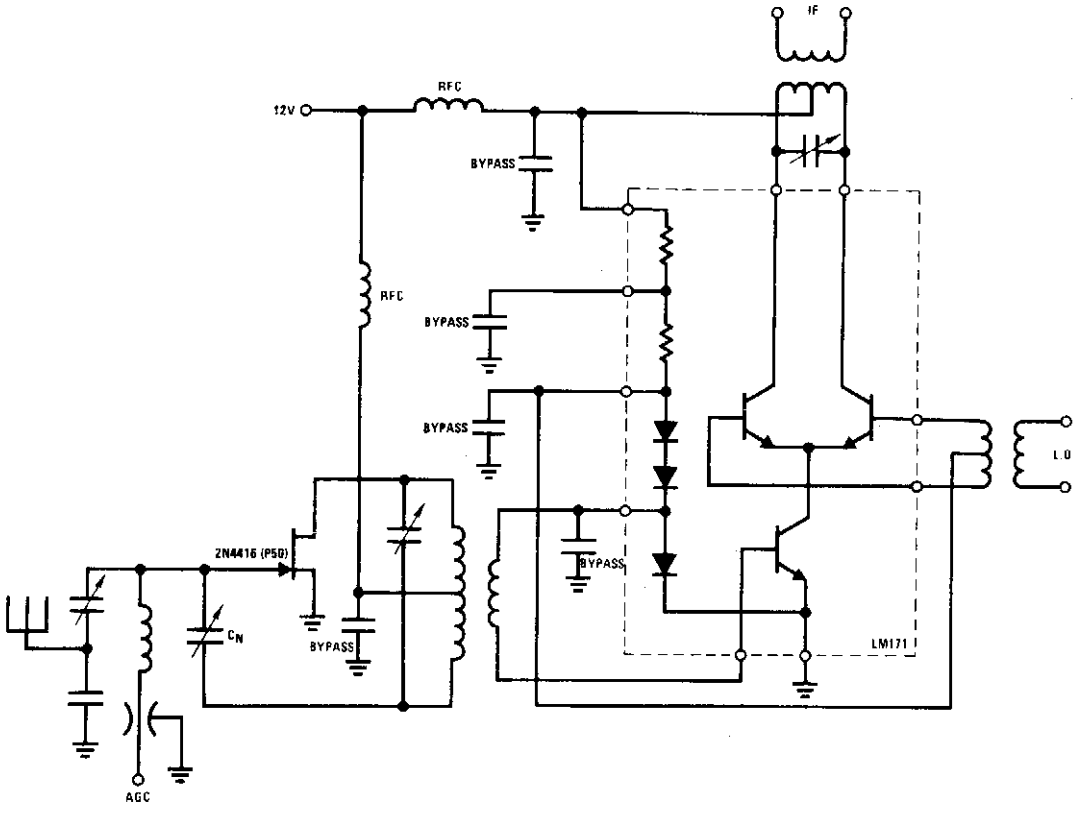
SIGNETICS

Fig. 17-12

Circuit Notes

An ac voltmeter may be easily constructed. Simplicity of the circuit and low component count make it particularly attractive. The demodulator output is a full-wave rectified signal from the ac input at Pin 4. The dc component on the rectified signal at Pin 5 varies linearity with the rms input at Pin 4 and thus provides an accurate rms-to-dc conversion at the output of the filter (Pin 1). C_T is a variable capacitor that is tweaked until the oscillator signal to the sync input of the demodulator is in phase with the ac signal at Pin 4.

100 MHz CONVERTER



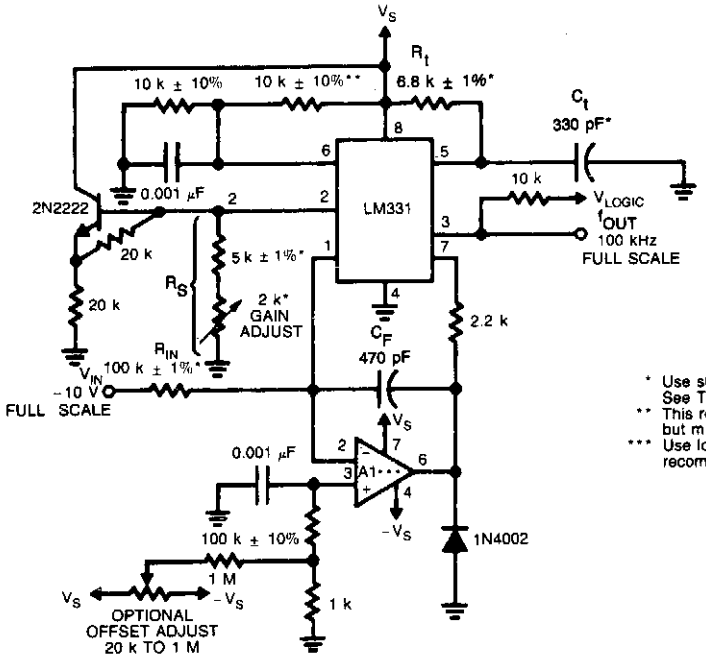
NATIONAL SEMICONDUCTOR CORP.

Fig. 17-13

Circuit Notes

The 2N4416 JFET will provide noise figures of less than 3 dB and power gain of greater than 20 dB. The JFET's outstanding low crossmodulation and low intermodulation distortion provides an ideal characteristic for an input stage. The output feeds into an LM171 used as a balanced mixer. This configuration greatly reduces local oscillator radiation both into the antenna and into the *if* strip and also reduces *rf* signal feedthrough.

PRECISION VOLTAGE-TO-FREQUENCY CONVERTER



* Use stable components with low temperature coefficients. See Typical Applications section.
** This resistor can be 5 kΩ or 10 kΩ for V_S = 8 V to 22 V, but must be 10 kΩ = 4.5 V to 8 V.
*** Use low offset voltage and low offset current op amps for A1: recommended types LF411A or LF356.

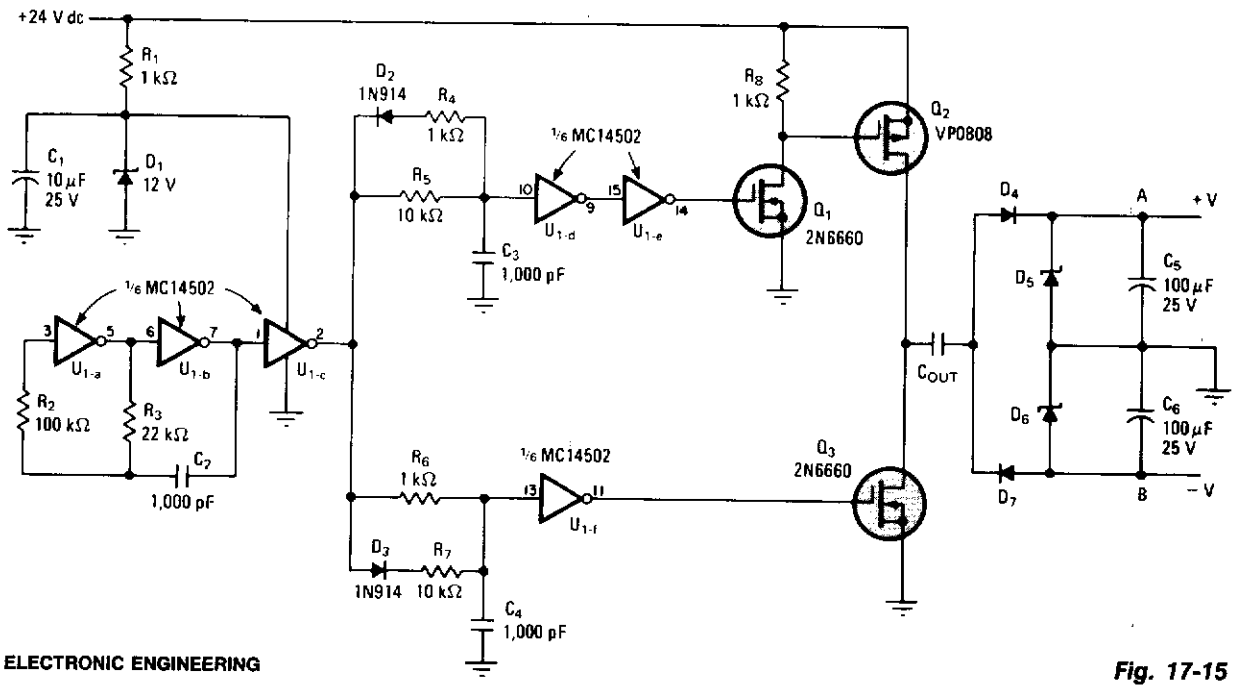
NATIONAL SEMICONDUCTOR CORP.

Fig. 17-14

Circuit Notes

In this circuit, integration is performed by using a conventional operational amplifier and feedback capacitor, C_F. When the integrator's output crosses the nominal threshold level at pin 6 of the LM131, the timing cycle is initiated. The average current fed into the op amp's summing point (pin 2) is $i \times (1.1\ R_tC_t) \times f$ which is perfectly balanced with $-V_{IN}/R_{IN}$. In this circuit, the voltage offset of the LM131 input comparator does not affect the offset or accuracy of the V-to-F converter as it does in the stand-alone V-to-F converter, nor does the LM131 bias current or offset current. Instead, the offset voltage and offset current of the operational amplifier are the only limits on how small the signal can be accurately converted.

BIPOLAR DC-DC CONVERTER REQUIRES NO INDUCTOR



ELECTRONIC ENGINEERING

Fig. 17-15

Circuit Notes

Inverters U1a and U1b form a 20-kilohertz oscillator whose square wave output—further shaped by D2, R4, and R5 and by D3, R6, and R7—drives power field-effect transistors Q2 and Q3. The p-channel and n-channel FETs conduct alternately, in a push-pull configuration. When Q2 conducts, the positive charge on C_{out} forces diode D4 to conduct as well, which produces a positive voltage, determined by zener diode D5, at terminal A. Similarly, when Q3, in its turn conducts, the negative charge on C_{out} forces D7 to do so as well. A negative voltage, therefore, develops at terminal B, whose level is set by D6.

18

Counters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

8-Digit Up/Down Counter
Ring Counter with Variable Timing
20 kHz Ring Counter
Binary Counter
100 MHz Frequency, Period Counter
Analog Counter Circuit
Attendance Counter
10 MHz Universal Counter

8-DIGIT UP/DOWN COUNTER

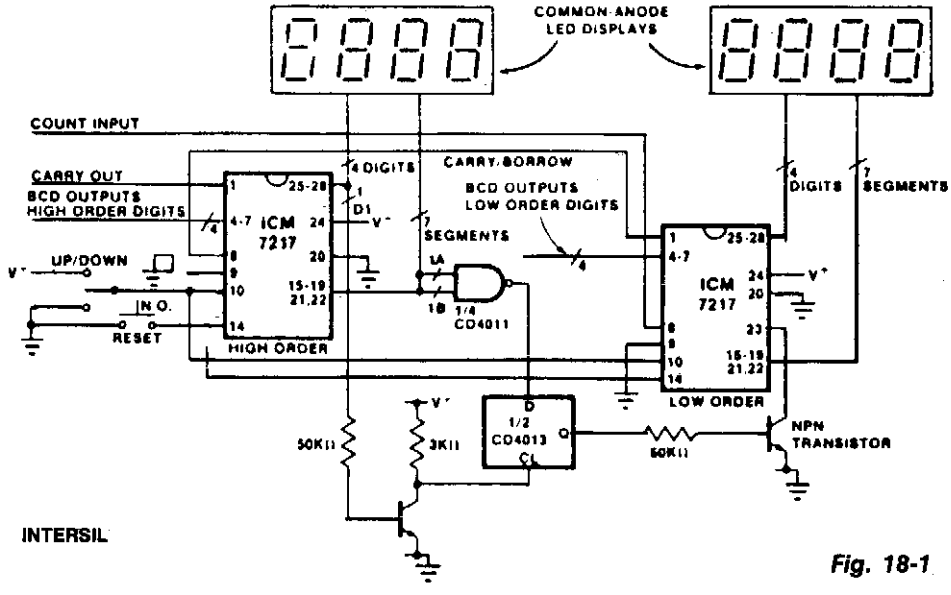
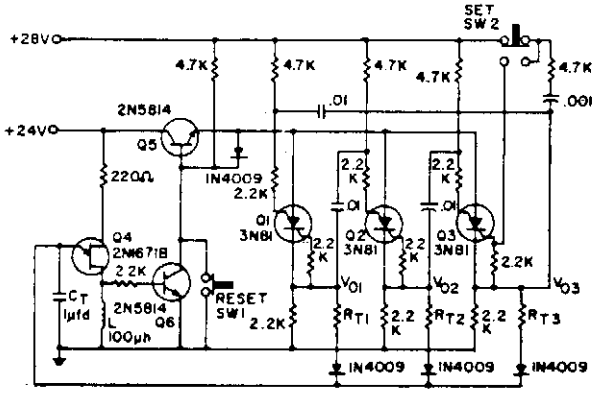


Fig. 18-1

Circuit Notes

This circuit shows how to cascade counters and retain correct leading zero blanking. The NAND gate detects whether a digit is active since one of the two segments a or b is active on any unblanked number. The flip flop is clocked by the least significant digit of the high order counter, and if this digit is not blanked, the Q output of the flip flop goes high and turns on the npn transistor, thereby inhibiting leading zero blanking on the low order counter.

RING COUNTER WITH VARIABLE TIMING



GENERAL ELECTRIC

Fig. 18-2

Circuit Notes

Shift pulses are generated by the unijunction transistors. The intervals between pulses are controlled by C_T and R_T. A different R_T can be selected for each stage of the counter as shown.

20 kHz RING COUNTER

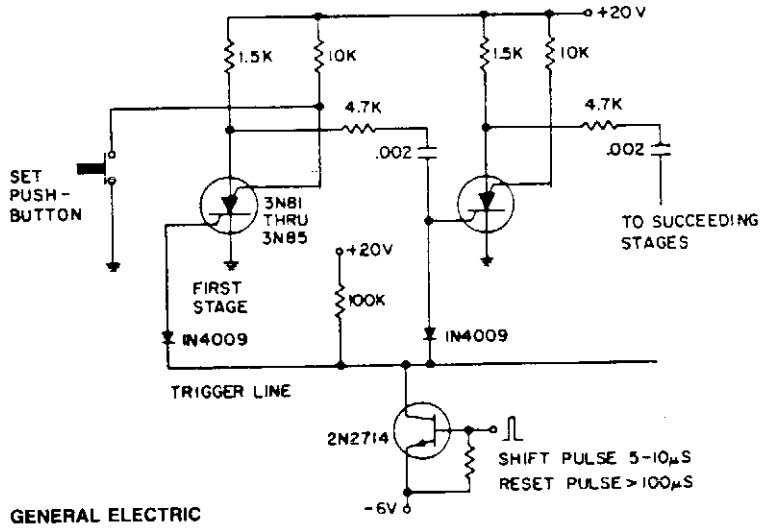


Fig. 18-3

Circuit Notes

The shift pulse turns off the conducting scs by reverse biasing the cathode gate. The charge stored on the coupling capacitor then triggers the next stage. An excessively long shift pulse charges up all the capacitors, turning off all stages. Grounding an anode gate will "set" that stage.

BINARY COUNTER

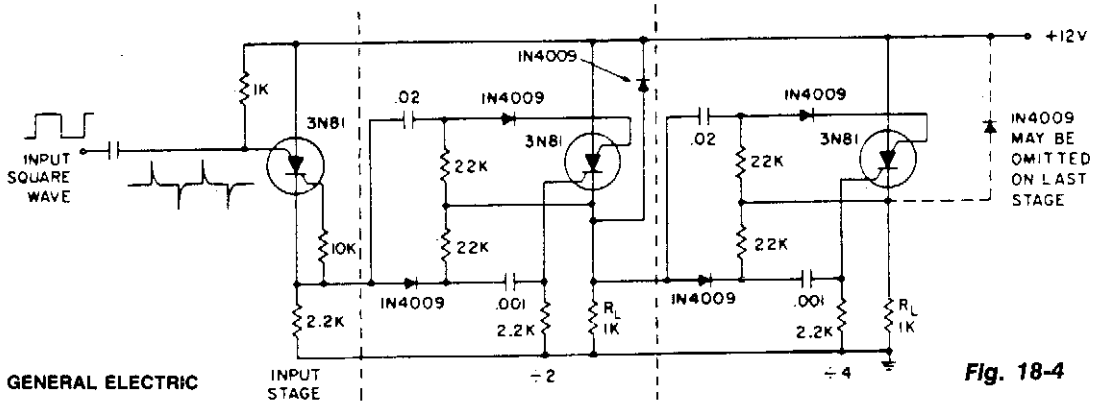


Fig. 18-4

Circuit Notes

Stages are triggered by the positive going edge. The scs is turned on at the cathode gate; turned off at the anode gate. The anode-to-cathode IN4009 suppresses positive transients while the scs is recovering. The input stage generates fast positive edges to trigger the counter.

100 MHz FREQUENCY, PERIOD COUNTER

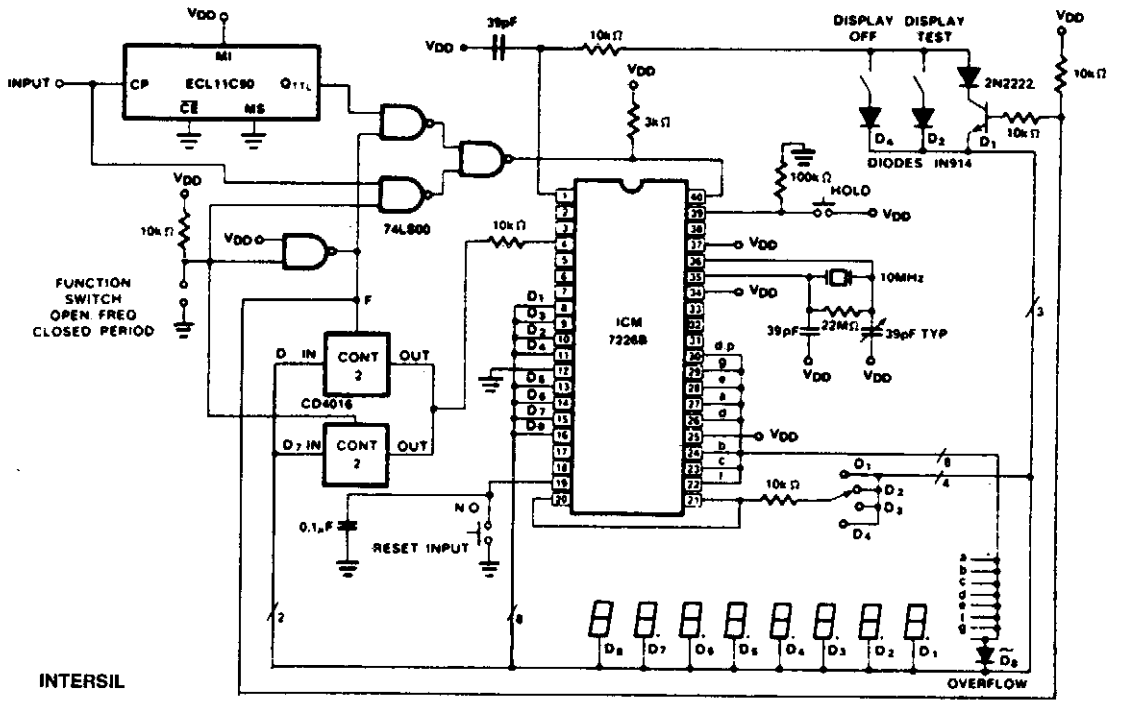
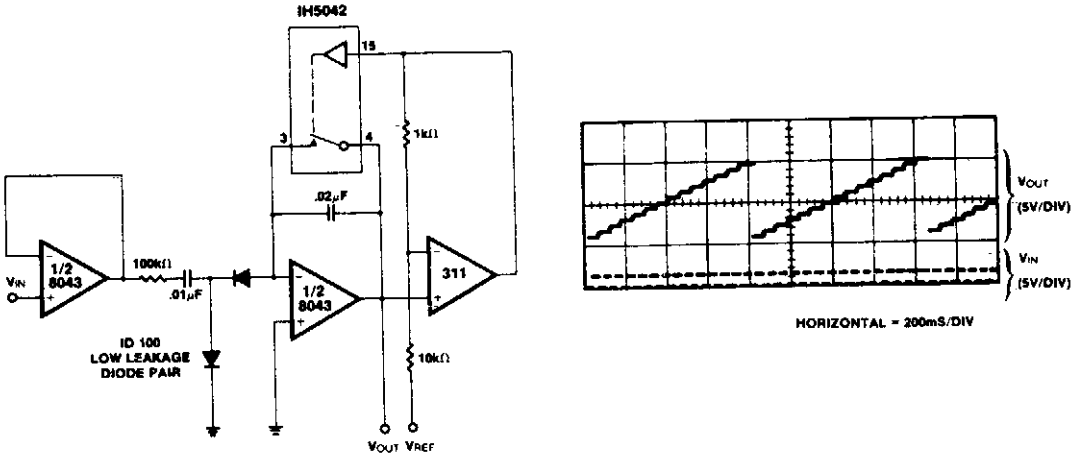


Fig. 18-5

Circuit Notes

The figure shows the use of a CD4016 analog multiplex to multiplex the digital outputs back to the FUNCTION input. Since the CD4016 is a digitally controlled analog transmission gate, no level shifting of the digit output is required. The CD4051's or CD4052's could also be used to select the proper inputs for the multiplexed input on the ICM7226 from 2 or 3 bit digital inputs. These analog multiplexers may also be used in systems in which the mode of operation is controlled by a microprocessor rather than directly from front panel switches. TTL multiplexers such as the 74LS153 or 74LS251 may also be used, but some additional circuitry will be required to convert the digit output to TTL compatible logic levels.

ANALOG COUNTER CIRCUIT



INTERSIL

Fig. 18-6

Circuit Notes

A straightforward circuit using a LM311 for the level detector and a CMOS analog gate to discharge the capacitor is shown. An important property of this type of counter is the ease with which the count can be changed; it is only necessary to change the voltage at which the comparator trips. A low cost A-D converter can also be designed using the same principle since the digital count between reset periods is directly proportional to the analog voltage used as a reference for the comparator. A considerable amount of hysteresis is used in the comparator. This ensures that the capacitor is completely discharged during the reset period. In a more sophisticated circuit, a dual comparator "window detector" could be used, the lower trip point is set close to ground to ensure complete discharge. The upper trip point could then be adjusted independently to determine the pulse count.

ATTENDANCE COUNTER

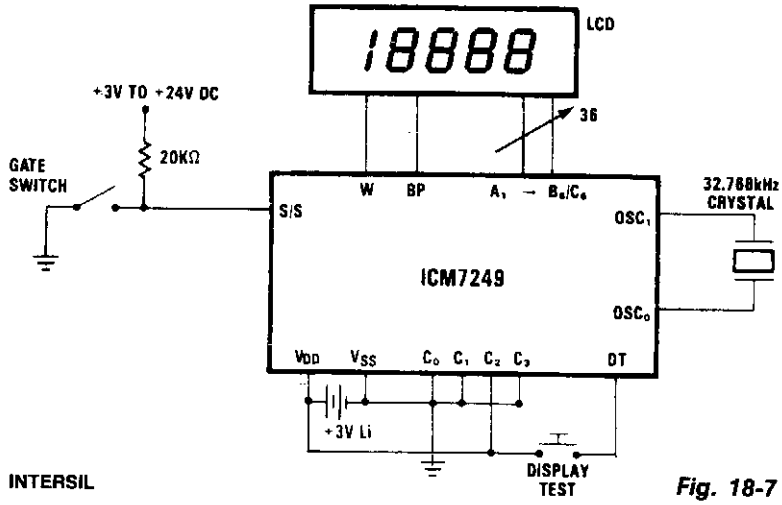


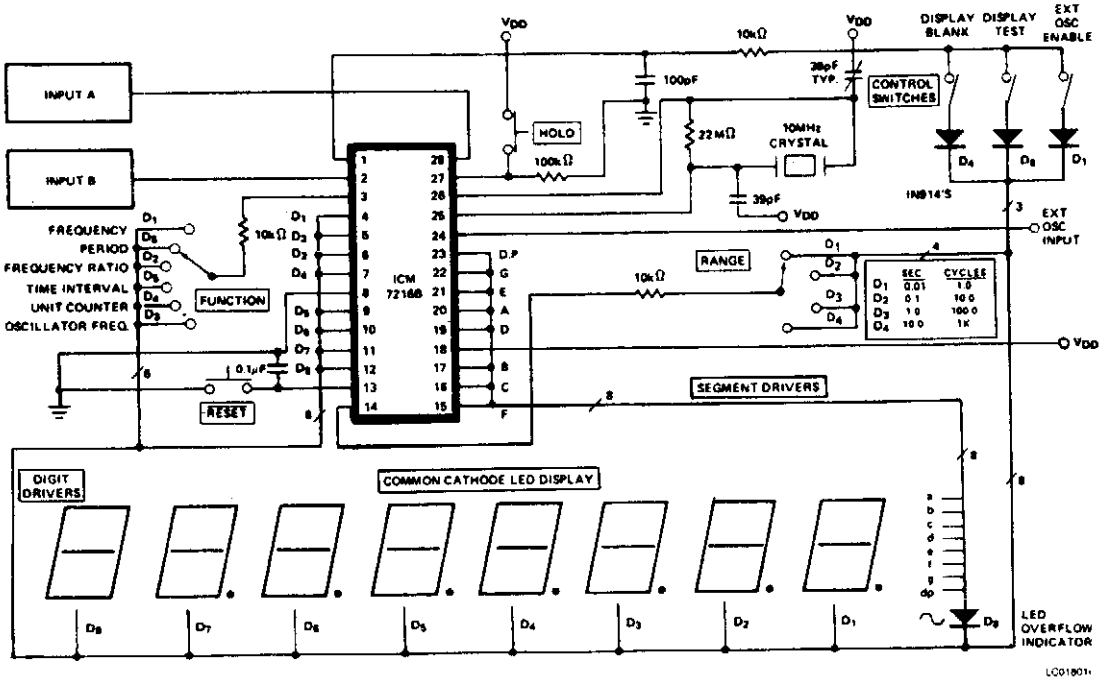
Fig. 18-7

Circuit Notes

The display shows each increment. By using mode 2, external debouncing of the gate switch is unnecessary, provided the switch bounce is less than 35ms. The 3 V lithium battery can be replaced without disturbing operation if a suitable capacitor is connected in parallel with it. The display should be disconnected, if possible, during the procedure to minimize current drain. The capacitor should be large enough to store charge for the amount of time needed to physically replace the battery ($t = VC/1$). A 100 μF capacitor initially charged to 3 V will supply a current of 1.0 μA for 50 seconds before its voltage drops to 2.5 V, which is the minimum operating voltage for the ICM7249.

Before the battery is removed, the capacitor should be placed in parallel, across the V_{DD} and GND terminals. After the battery is replaced, the capacitor can be removed and the display reconnected.

10 MHz UNIVERSAL COUNTER



INTERSIL

Fig. 18-8

Circuit Notes

The ICM7216A or B can be used as a minimum component complete Universal Counter. This circuit can use input frequencies up to 10 MHz at INPUT A and 2 MHz at INPUT B. If the signal at INPUT A has a very low duty cycle it may be necessary to use a 74121 monostable multivibrator or similar circuit to stretch the input pulse width to be able to guarantee that it is at least 50 ns in duration.

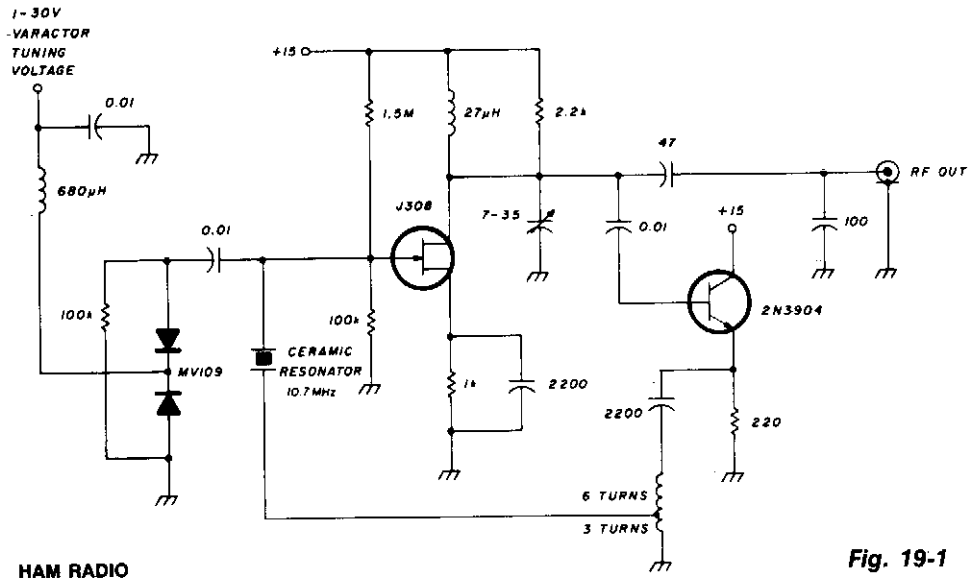
19

Crystal Oscillators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Varactor-Tuned 10 MHz Ceramic Resonator Oscillator
10 MHz Crystal-Controlled Oscillator
Low Power, 5V Driven, Temperature Compensated Crystal Oscillator (TXCO)
Crystal-Controlled LO for SSB Transmitter
Crystal Oscillator
Crystal Controlled Signal Source
1 MHz FET Crystal Oscillator
Pierce Crystal Oscillator
IC-Compatible Crystal Oscillator
Crystal Oscillator Provides Low Noise
Low-Frequency Crystal Oscillator–10 kHz–150 kHz
Overtone Crystal Oscillator
Colpitts Oscillator
Crystal-Controlled Oscillator
High-Frequency Crystal Oscillator
Crystal-Controlled Oscillator Operates from One Mercury Cell
High-Frequency Signal Generator
Crystal Tester
Crystal Stabilized IC Timer can Provide Subharmonic Frequencies

VARACTOR-TUNED 10 MHz CERAMIC RESONATOR OSCILLATOR



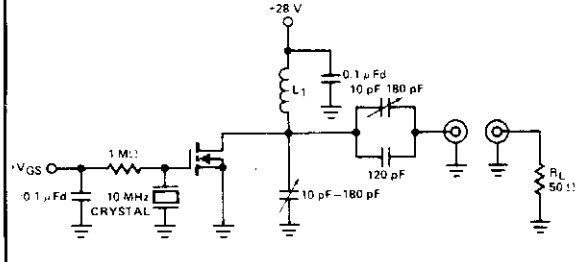
HAM RADIO

Fig. 19-1

Circuit Notes

The FET input amplifier has fixed bias with source feedback. This provides a very high input impedance with very low capacitance. The FET amplifier drives an emitter follower which, in spite of the fact that it has a low output impedance, feeds a transformer with a 3:1 turns ratio for a nine-fold impedance reduction. The result is an impedance at the ceramic resonator of a few ohms maximum. The varactor-tuned ceramic resonator oscillator has a significant frequency-temperature coefficient. The tuning range of the VCO is approximately 232 kHz, with a temperature coefficient of 350 Hz per degree centigrade. When using this circuit as a VCO, the entire 232 kHz range cannot be used because some of the tuning range must be sacrificed for the temperature dependence. If the required tuning range were 200 kHz, leaving 32 kHz for temperature variation, the resulting temperature variation would be more than 90°C.

10 MHz CRYSTAL OSCILLATOR



SILICONIX INC.

Fig. 19-2

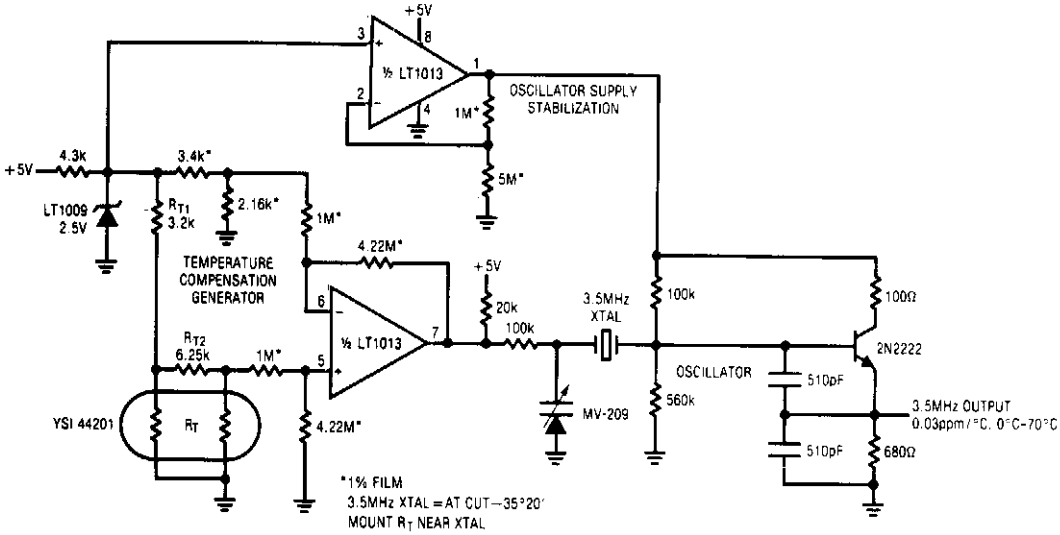
Circuit Notes

This xtal oscillator is a FET equivalent of a vacuum tube tuned to plate-tuned grid xtal oscillator. Feedback is via the drain to gate capacitance.

Parts List

L_1 – 18 turns #22 enameled wire on micrometals T-50-6 torroid core. ≈ 1.0 µH.

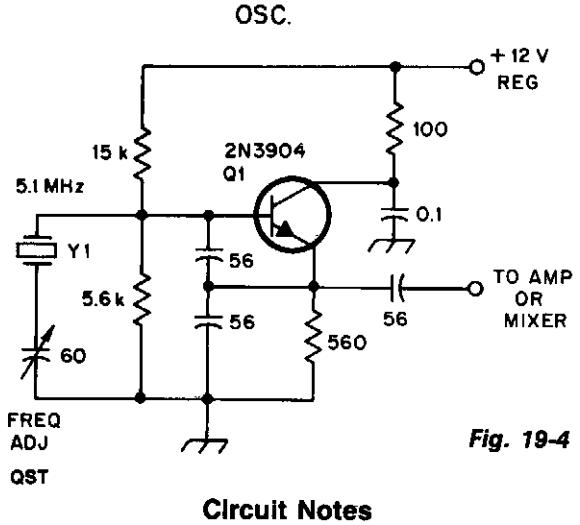
LOW POWER, 5 V DRIVEN, TEMPERATURE COMPENSATED CRYSTAL OSCILLATOR (TXCO)



LINEAR TECHNOLOGY CORP.

Fig. 19-3

CRYSTAL-CONTROLLED LOCAL OSCILLATOR FOR SSB TRANSMITTER



QST

Fig. 19-4

Circuit Notes

This oscillator may contain several switched crystals to provide channelized operation. A buffer amplifier may be added, if desired.

CRYSTAL OSCILLATOR

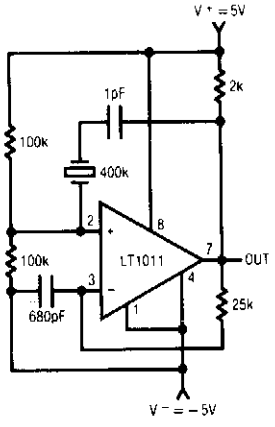


Fig. 19-5

LINEAR TECHNOLOGY CORPORATION

Circuit Notes

This circuit uses an LT1011 comparator biased in its linear mode and a crystal to establish its resonant frequency. This circuit can achieve a few hundred kHz, temperature independent clock frequency with nearly 50% duty cycle.

CRYSTAL-CONTROLLED SIGNAL SOURCE

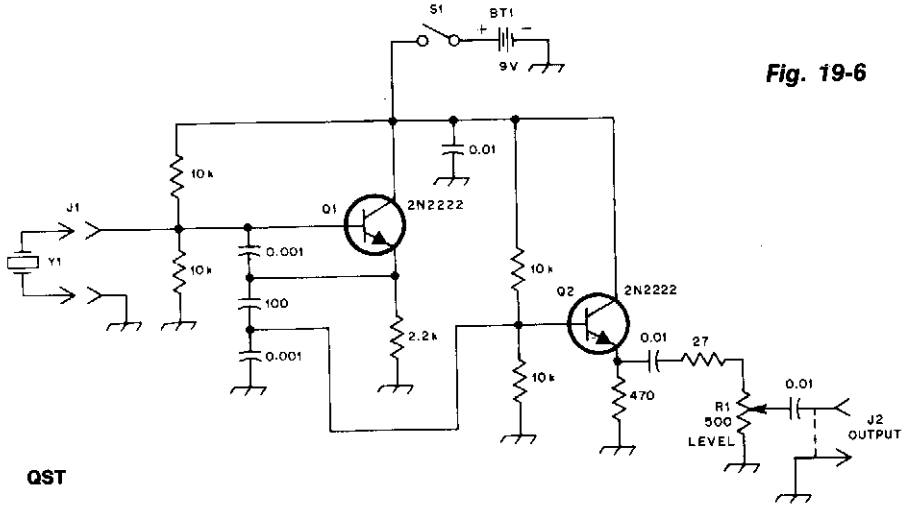


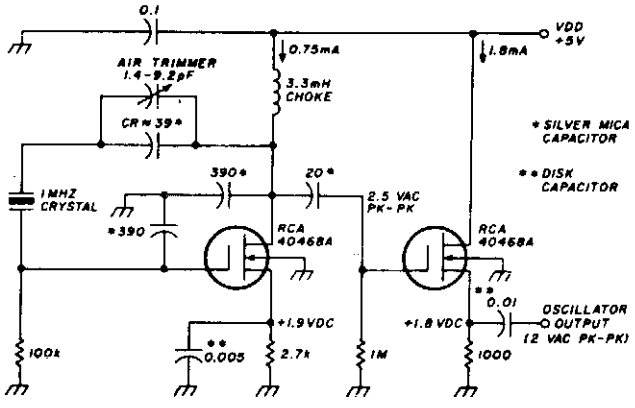
Fig. 19-6

QST

Circuit Notes

This general purpose signal source serves very well in signal-tracing applications. The output level is variable to more than 1 Vrms into a 50 Ω load. Almost any crystal in the 1 to 15 MHz range can be used. Q1 forms a Colpitts oscillator with the output taken from the emitter. A capacitive voltage divider (across the 2.2 K emitter resistor) reduces the voltage applied to the buffer amplifier, Q2. The buffer and emitter follower, provides the low input impedance necessary to drive 50 Ω loads.

1 MHz FET CRYSTAL OSCILLATOR



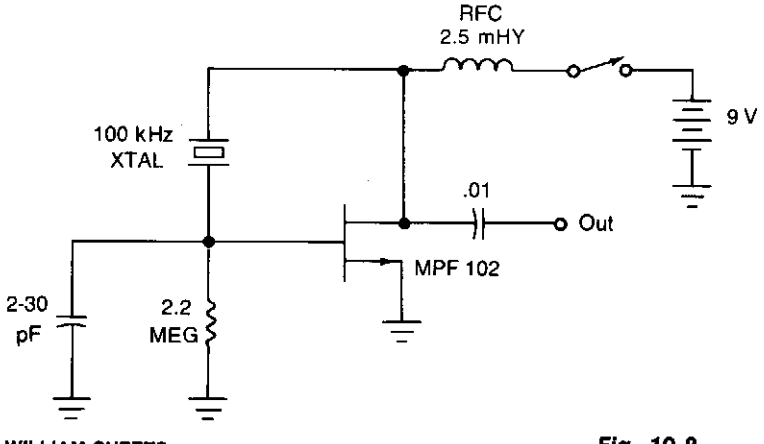
HAM RADIO

Fig. 19-7

Circuit Notes

This stable oscillator circuit exhibits less than 1 Hz frequency change over a V_{DD} range of 3–9 volts. Stability is attributed to the use of MOSFET devices and the use of stable capacitors.

PIERCE CRYSTAL OSCILLATOR



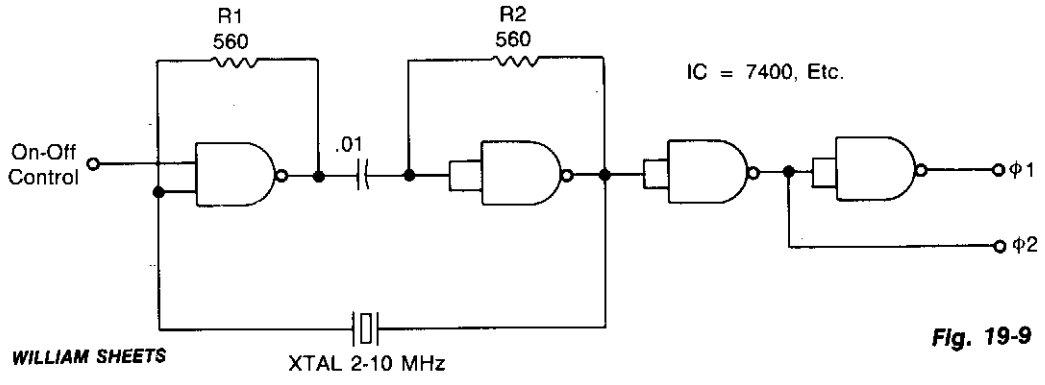
WILLIAM SHEETS

Fig. 19-8

Circuit Notes

The JFET Pierce oscillator is stable and simple. It can be the clock of a microprocessor, a digital timepiece or a calculator. With a probe at the output, it can be used as a precise injection oscillator for troubleshooting. Attach a small length of wire at the output and this circuit becomes a micropower transmitter.

IC-COMPATIBLE CRYSTAL OSCILLATOR



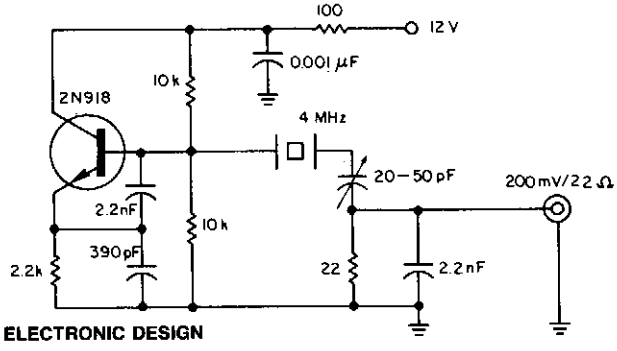
WILLIAM SHEETS

Fig. 19-9

Circuit Notes

Resistors R1 and R2 temperature-stabilize the NAND gates; they also ensure that the gates are in a linear region for starting. Capacitor C1 is a dc block; it must have less than 1/10 ohm impedance at the operating frequency. The crystal runs in a series-resonant mode. Its series resistance must be low; AT-cut crystals for the 1- to 10-MHz range work well. The output waveshape has nearly a 50% duty cycle, with chip-limited rise times. The circuit starts well from 0° to 70°C.

CRYSTAL OSCILLATOR PROVIDES LOW NOISE



ELECTRONIC DESIGN

Fig. 19-10

Circuit Notes

The oscillator delivers an output of high spectral purity without any substantial sacrifice of the usual stability of a crystal oscillator. The crystal in addition to determining the oscillator's frequency, is used also as a low-pass filter for the unwanted harmonics and as a bandpass filter for the sideband noise. The noise bandwidth is limited to less than 100 Hz. All higher harmonics are substantially suppressed—60 dB down for the third harmonic of the 4-MHz fundamental oscillator frequency.

LOW-FREQUENCY CRYSTAL OSCILLATOR—10 kHz to 150 kHz

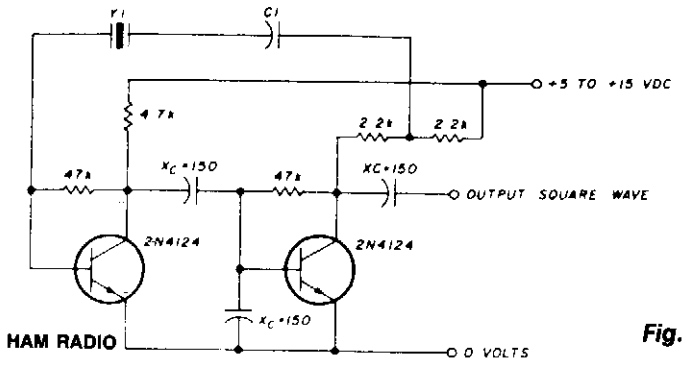
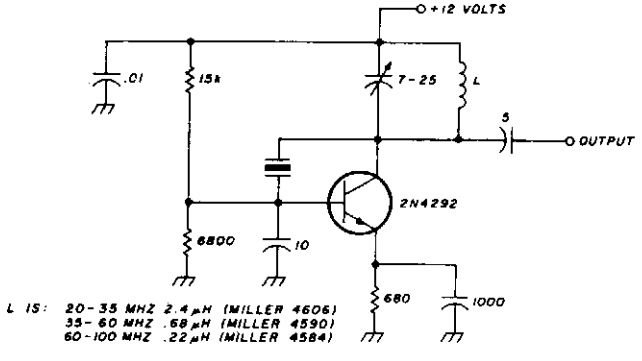


Fig. 19-11

Circuit Notes

C1 in series with the crystal may be used to adjust the oscillator output frequency. Value may range between 20 pF and 0.01 μF, or may be a trimmer capacitor and will approximately equal the crystal load capacitance. X values are approximate and can vary for most circuits and frequencies; this is also true for resistance values. Adequate power supply decoupling is required; local decoupling capacitors near the oscillator are recommended. All leads should be extremely short in high frequency circuits.

OVERTONE CRYSTAL OSCILLATOR



HAM RADIO

Fig. 19-12

Circuit Notes

This oscillator is designed for overtone crystals in the 20–100 MHz range operating in the third and fifth mode. Operating frequency is determined by the tuned circuit.

COLPITTS OSCILLATOR

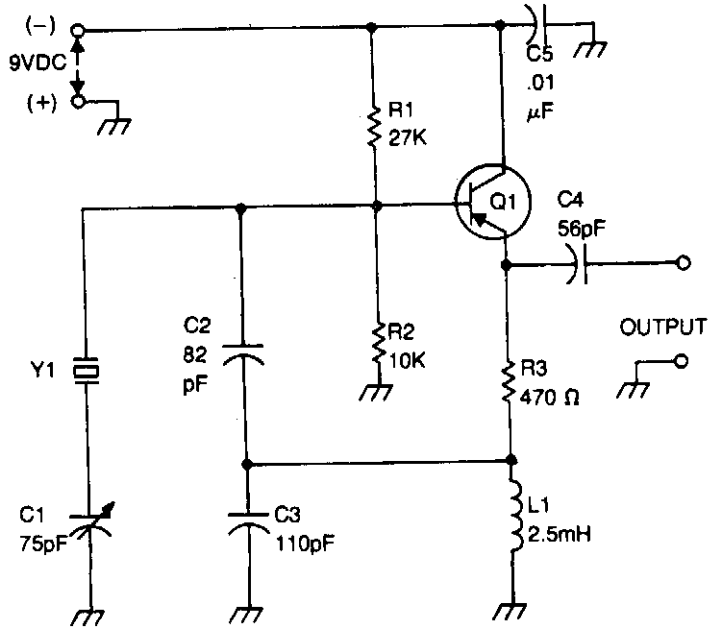


Fig. 19-13

TAB BOOKS, INC.

Circuit Notes

Bias for the pnp bipolar transistor is provided by resistor voltage divider network R1/R2. The collector of the oscillator transistor is kept at ac ground by capacitor C5, placed close to the transistor. Feedback is provided by capacitor voltage divider C2/C3.

CRYSTAL-CONTROLLED OSCILLATOR

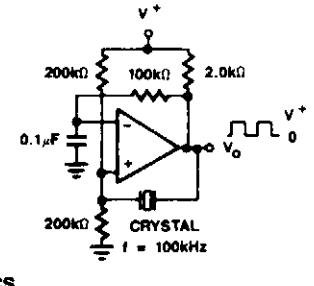
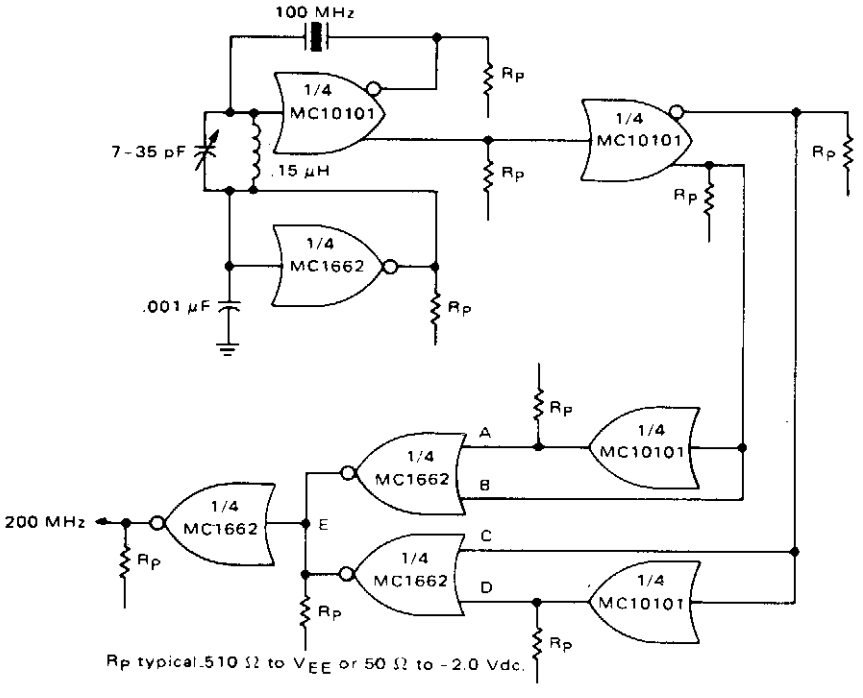


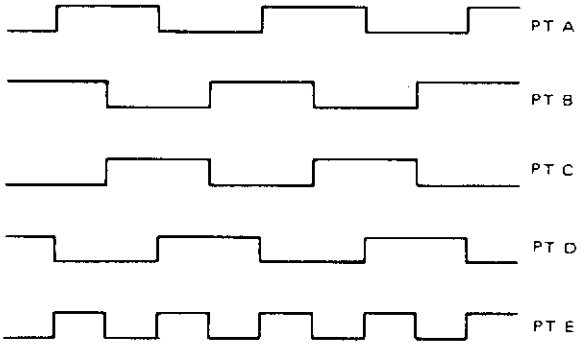
Fig. 19-14

SIGNETICS

HIGH-FREQUENCY CRYSTAL OSCILLATOR



200 MHz Crystal Oscillator



Frequency Doubler Waveforms

MOTOROLA, INC.

Fig. 19-15

HIGH-FREQUENCY CRYSTAL OSCILLATOR, Continued.

Circuit Notes

A high speed oscillator is possible by combining an MECL 10 K crystal oscillator with an MECL III frequency doubler as shown. One section of the MC10101 is connected as a 100 MHz crystal oscillator with the crystal in series with the feedback loop. The LC tank circuit tunes the 100 MHz harmonic of the crystal and may be used to calibrate the circuit to the exact frequency. A second section of the MC10101 buffers the crystal oscillator and gives complementary 100 MHz signals. The frequency doubler consists of two MC10101 gates as phase shifters and two MC1662 NOR gates. For a 50% duty cycle at the output, the delay to the true and complement 100 MHz signals should be 90°. This may be built precisely with 2.5 ns delay lines for the 200 MHz output or approximated by the two MC10101 gates. The gates are easier to incorporate and cause only a slight skew in output signal duty cycle. The MC1662 gates combine the 4 phase 100 MHz signals as shown in Figure B. The outputs of the MC1662's are wire-OR connected to give the 200 MHz signal. MECL III gates are used because of the bandwidth required for 200 MHz signals. One of the remaining MC1662 gates is used as a V_{BB} bias generator for the oscillator. By connecting the NOR output to the input, the circuit stays in the center of the logic swing or at V_{BB}. A 0.001 μF capacitor ensures the V_{BB} circuit does not oscillate.

CRYSTAL-CONTROLLED OSCILLATOR OPERATES FROM ONE MERCURY CELL

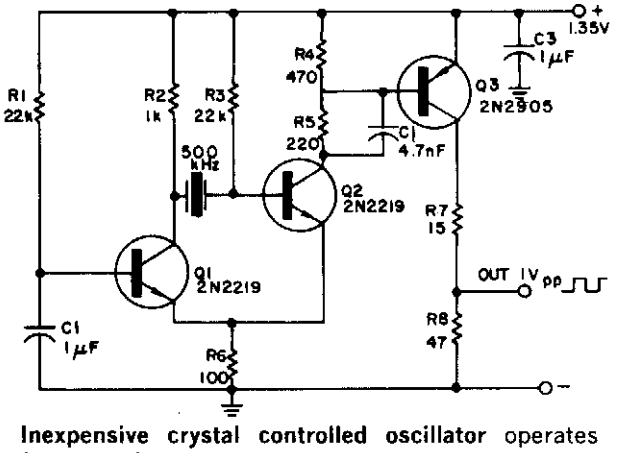


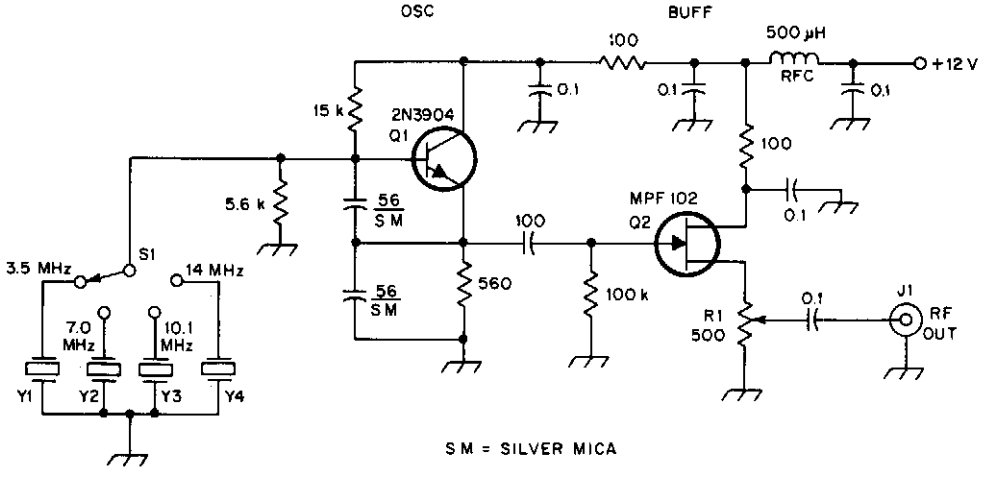
Fig. 19-16

ELECTRONIC DESIGN **Inexpensive crystal controlled oscillator** operates from a 1.35-volt source.

Circuit Notes

The circuit is powered by a single 1.35 V mercury cell and provides a 1 V square-wave output. As shown, the crystal is a tuned circuit between transistors Q1 and Q2, which are connected in the common-emitter configuration. Positive feedback provided by means of R permits oscillation. The signal at the collector of Q2 is squared by Q3, which switches between cutoff and saturation. R7 permits short-circuit-proof operation.

HIGH FREQUENCY SIGNAL GENERATOR



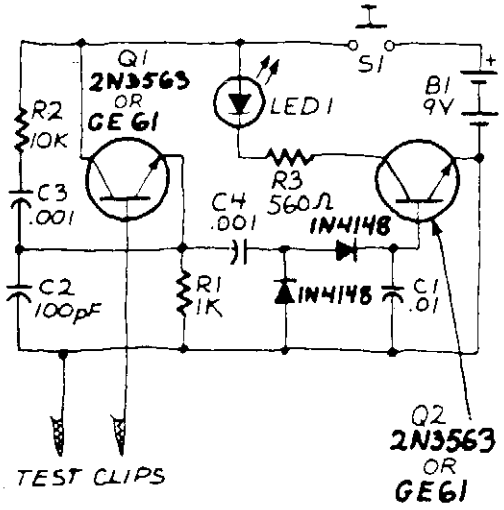
QST

Fig. 19-17

Circuit Notes

A tapped-coil Colpitts oscillator is used at Q1 to provide four tuning ranges from 1.7 to 3.1 MHz, 3.0 to 5.6 MHz, 5.0 to 12 MHz and 11.5 to 31 MHz. A Zener diode (D2) is used at Q1 to lower the operating voltage of the oscillator. A small value capacitor is used at C5 to ensure light coupling to the tuned circuit. Q2 is a source-follower buffer stage. It helps to isolate the oscillator from the generator-output load. The source of Q2 is broadly tuned by means of RFC1. Energy from Q2 is routed to a fed-back, broadband class-A amplifier. A 2 dB attenuator is used at the output of T1 to provide a 50 ohm termination for Q3 and to set the generator-output impedance at 50 ohms. C16, C17 and RFC2 form a brute-force RF decoupling network to keep the generator energy from radiating outside the box on the 12 V supply.

CRYSTAL TESTER



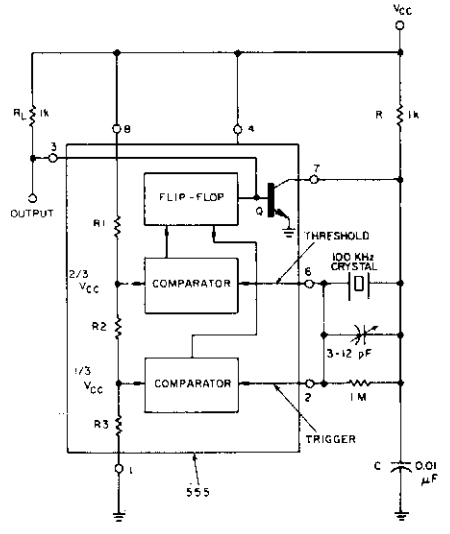
RADIO-ELECTRONICS

Fig. 19-18

Circuit Notes

Transistor Q1, a 2N3563, and its associated components form an oscillator circuit that will oscillate if, and only if, a good crystal is connected to the test clips. The output from the oscillator is then rectified by the two 1N4148 diodes and filtered by C1, a .01 μF capacitor. The positive voltage developed across the capacitor is applied to the base of Q2, another 2N3563, causing it to conduct. When that happens, current flows through LED1, causing it to glow. Since only a good crystal will oscillate, a glowing LED indicates that the crystal is indeed OK. The circuit is powered by a standard nine-volt transistor-radio battery and the SPST pushbutton power-switch is included to prolong battery life.

CRYSTAL-STABILIZED IC TIMER CAN PROVIDE SUBHARMONIC FREQUENCIES



ELECTRONIC DESIGN

Fig. 19-19

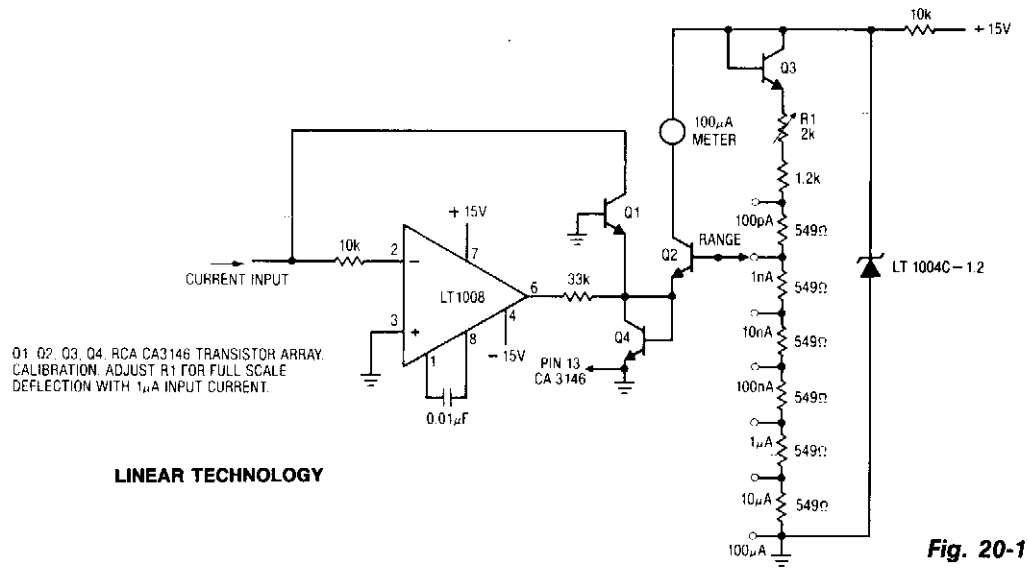
The trimmer across the crystal can finely tune the circuit's oscillating frequency.

20
Current Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Ammeter with Six Decade Range
Current Sensing in Supply Rails
Pico Ammeter
Electrometer Amplifier with Overload Protection
Guarded Input Picoammeter Circuit
Ammeter with Six Decade Range
Picoammeter Circuit

AMMETER WITH SIX DECADE RANGE



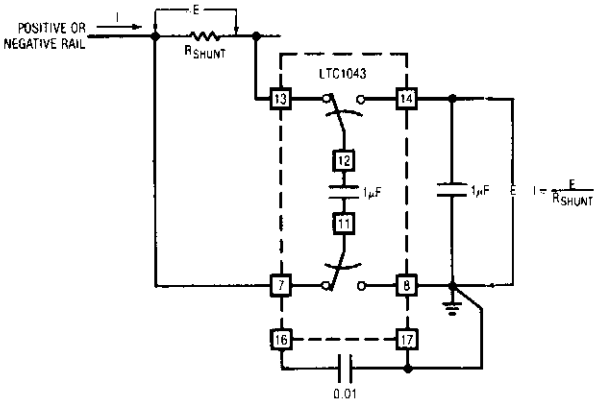
LINEAR TECHNOLOGY

Fig. 20-1

Circuit Notes

The Ammeter measures currents from 100 pA to 100 μA without the use of expensive high value resistors. Accuracy at 100 μA is limited by the offset voltage between Q1 and Q2 and, at 100 pA, by the inverting bias current of the LT1008.

CURRENT SENSING IN SUPPLY RAILS



Circuit Notes

The LTC1043 can sense current through a shunt in either of its supply rails. This capability has wide application in battery and solar-powered systems. If the ground-referred voltage output is unloaded by an amplifier, the shunt can operate with very little voltage drop across it, minimizing losses.

LINEAR TECHNOLOGY CORPORATION

Fig. 20-2

PICO AMMETER

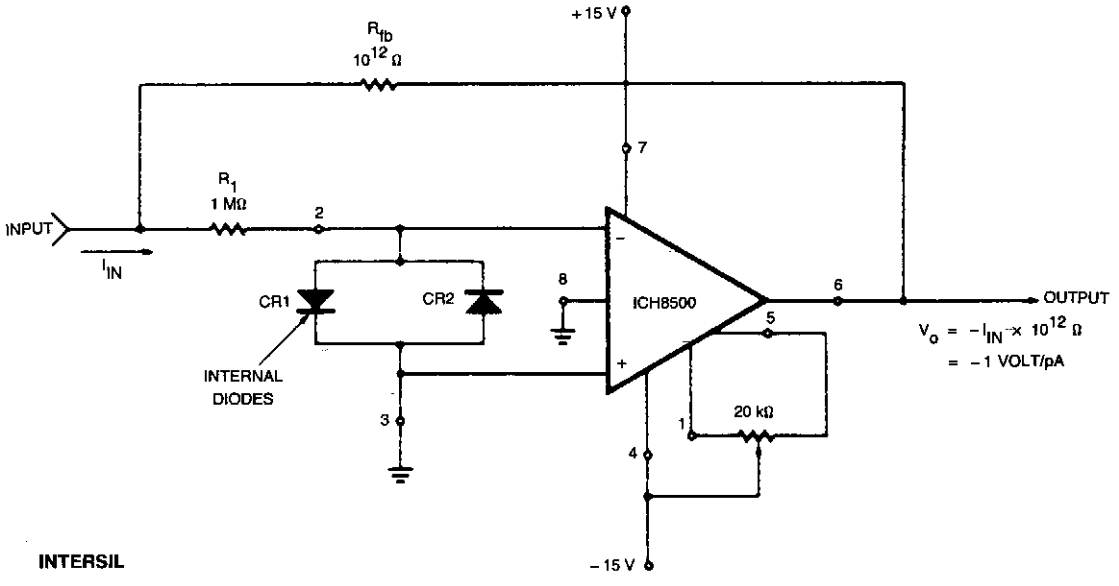
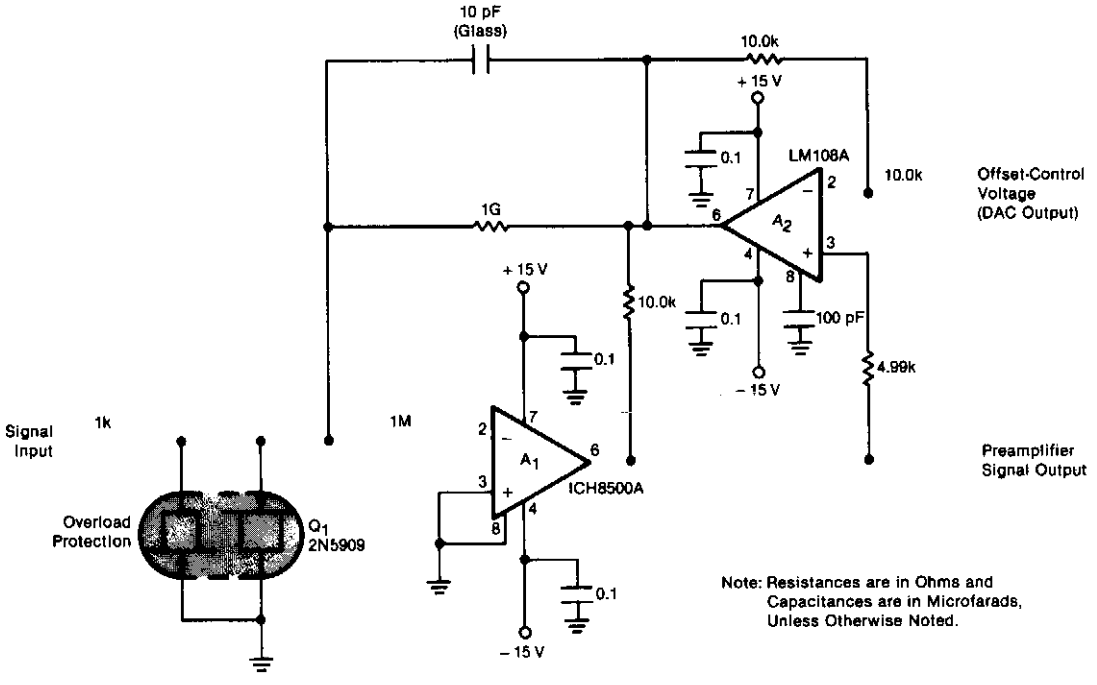


Fig. 20-3

Circuit Notes

Care must be taken to eliminate any stray currents from flowing into the current summing node. This can be accomplished by forcing all points surrounding the input to the same potential as the input. In this case the potential of the input is at virtual ground, or OV. Therefore, the case of the device is grounded to intercept any stray leakage currents that may otherwise exist between the ±15 V input terminals and the inverting input summing junctions. Feedback capacitance should be kept to a minimum in order to maximize the response time of the circuit to step function input currents. The time constant of the circuit is approximately the produce of the feedback capacitance C_{fb} times the feedback resistor R_{fb}. For instance, the time constant of the circuit is 1 sec if C_{fb} = 1pF. Thus, it takes approximately 5 sec (5 time constants) for the circuit to stabilize to within 1% of its final output voltage after a step function of input current has been applied. C_{fb} of less than 0.2 to 0.3 pF can be achieved with proper circuit layout.

ELECTROMETER AMPLIFIER WITH OVERLOAD PROTECTION



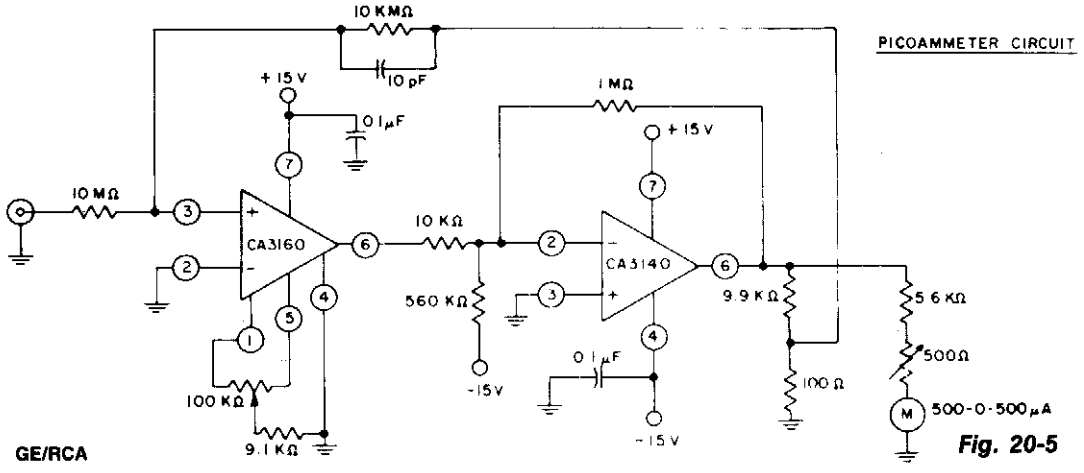
NASA TECH BRIEFS

Fig. 20-4

Circuit Notes

The preamplifier is protected from excessive input signals of either polarity by the 2N5909 junction field-effect transistor. A nulling circuit makes it possible to set the preamplifier output voltage to zero at a fixed low level (up to $\pm 10^{-8}$A) of the input current. (This level is called the standing current and corresponds to the zero-signal level of the instrumentation.) The opposing (offset) current is generated in the 10^9 feedback resistor to buck the standing current. Different current ranges are reached by feeding the preamplifier output to low and high gain amplifier chains. To reduce noise, each chain includes a 1.5 Hz corner active filter.

GUARDED INPUT PICOAMMETER CIRCUIT



GE/RCA

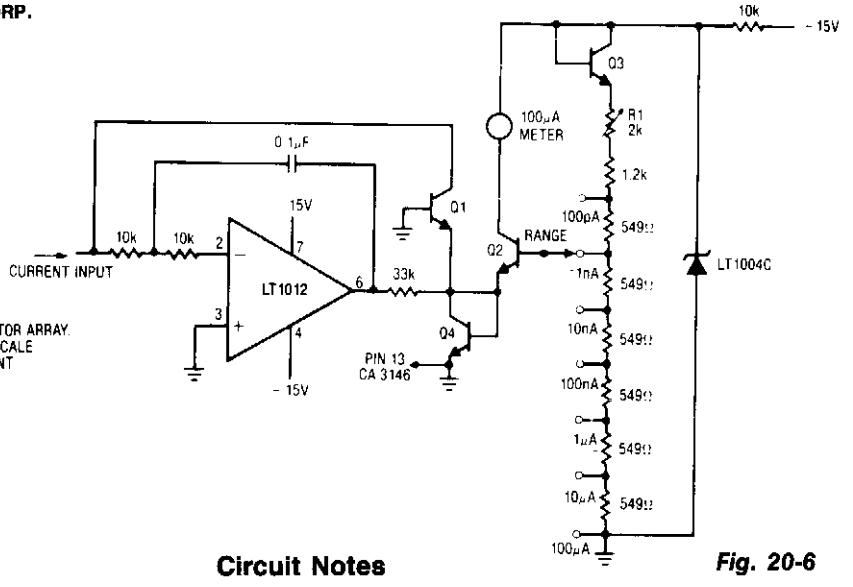
Fig. 20-5

Circuit Notes

The circuit utilizes CA3160 and CA3140 BiMOS op amps to provide a full-scale meter deflection of ±3 pA. The CA3140 serves as an X100 gain stage to provide the required plus and minus output swing for the meter and feedback network. Terminals 2 and 4 of the CA3160 are at ground potential, thus its input is operated in the "guarded mode."

AMMETER WITH SIX DECADE RANGE

LINEAR TECHNOLOGY CORP.



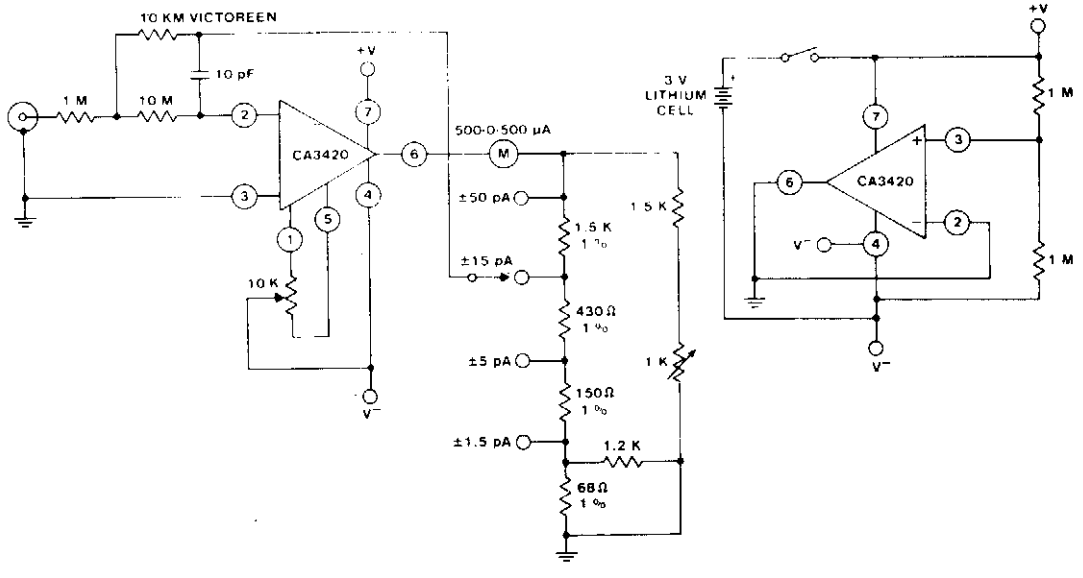
Q1, Q2, Q3, Q4, RCA CA3146 TRANSISTOR ARRAY.
CALIBRATION: ADJUST R1 FOR FULL SCALE DEFLECTION WITH 1μA INPUT CURRENT

Fig. 20-6

Circuit Notes

The Ammeter measures currents from 100 pA to 100 μA without the use of expensive high value resistors. Accuracy at 100 μA is limited by the offset voltage between Q1 and Q2 and, at 100 pA, by the inverting bias current of the LT1008.

PICOAMMETER CIRCUIT



GENERAL ELECTRIC /RCA

Fig. 20-7

Circuit Notes

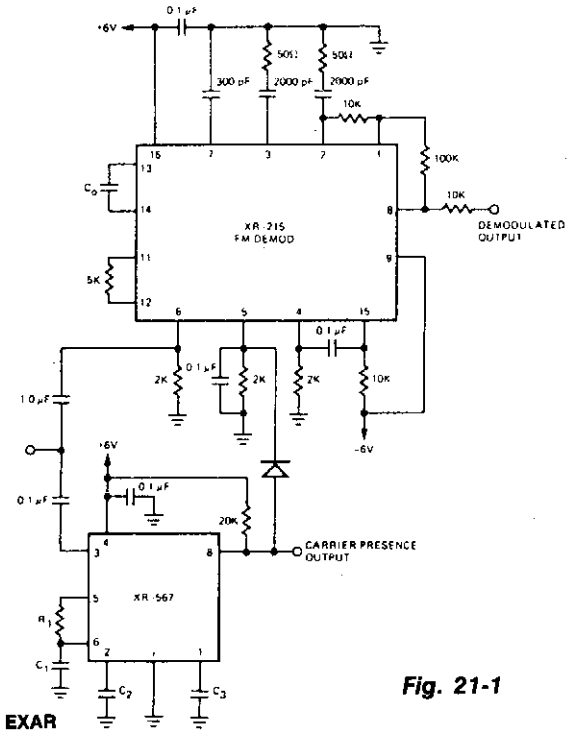
The circuit uses the exceptionally low input current (0.1pA) of the CA3420 BiMOS op amp. With only a single 10 megohm resistor, the circuit covers the range from ±50 pA maximum to a full-scale sensitivity of ±1.5 pA. Using an additional CA3420, a low-resistance center tap is obtained from a single 3-volt lithium battery.

21
Demodulators

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Narrow Band FM Demodulator with Carrier Detect
Stereo Demodulator
AM Demodulator
FM Demodulator

NARROW BAND FM DEMODULATOR WITH CARRIER DETECT



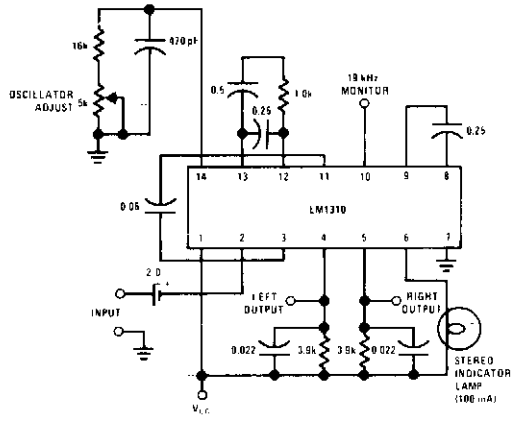
EXAR

Fig. 21-1

Circuit Notes

For FM demodulation applications where the bandwidth is less than 10% of the carrier frequency, an XR-567 can be used to detect the presence of the carrier signal. The output of the XR-567 is used to turn off the FM demodulator when no carrier is present, thus acting as a squelch. In the circuit shown, an XR-215 FM demodulator is used because of its wide dynamic range, high signal/noise ratio and low distortion. The XR-567 will detect the presence of a carrier at frequencies up to 500 kHz.

STEREO DEMODULATOR



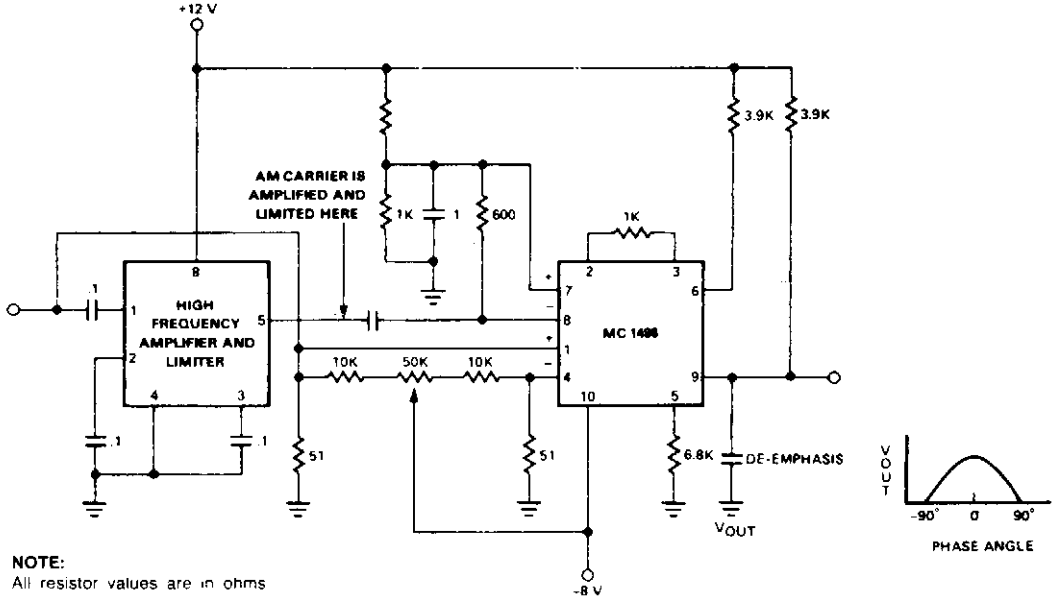
NATIONAL SEMICONDUCTOR CORP.

Fig. 21-2

Circuit Notes

This circuit uses a single IC LM1310 to provide left and right outputs from a composite MPX stereo signal. Oscillator adjust R1 is set for 76 kHz (19 kHz at pin 10). C1 should be a silver mica or NPO ceramic capacitor.

AM DEMODULATOR



NOTE:
All resistor values are in ohms

Fig. 21-3

SIGNETICS

Circuit Notes

Amplifying and limiting of the AM carrier is accomplished by the if gain block providing 55 dB of gain or higher with a limiting of 40 μV. The limited carrier is then applied to the detector at the carrier ports to provide the desired switching function. The signal is then demodulated by the synchronous AM demodulator (1496) where the carrier frequency is attentuated due to the balanced nature of the device. Care must be taken not to overdrive the signal input so that distortion does not appear in the recorded audio. Maximum conversion gain is reached when the carrier signals are in phase as indicated by the phase-gain relationship. Output filtering is also necessary to remove high frequency sum components of the carrier from the audio signal.

FM DEMODULATOR

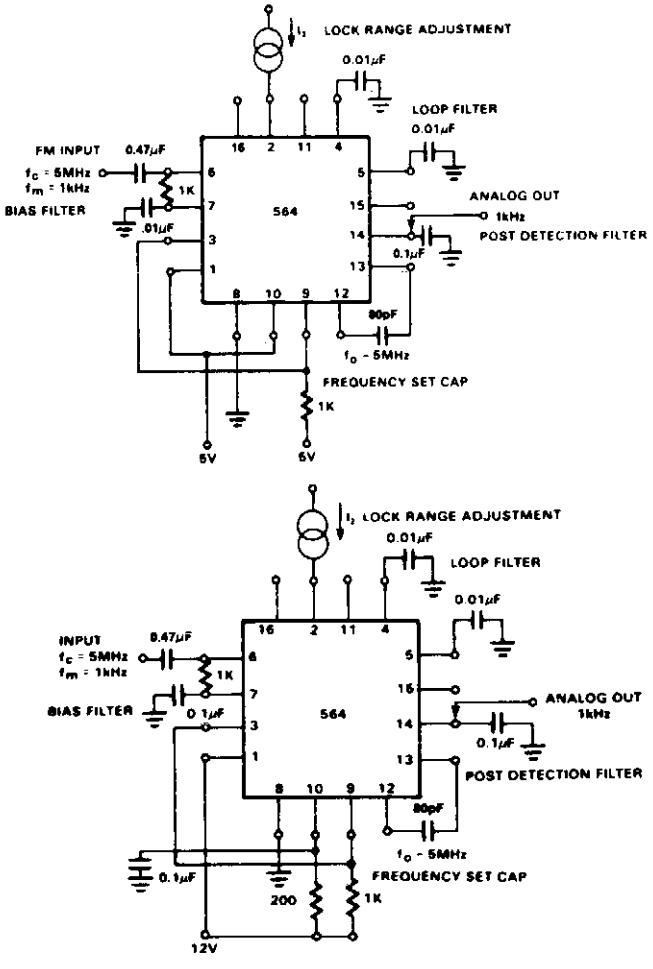


Fig. 21-4

SIGNETICS

Circuit Notes

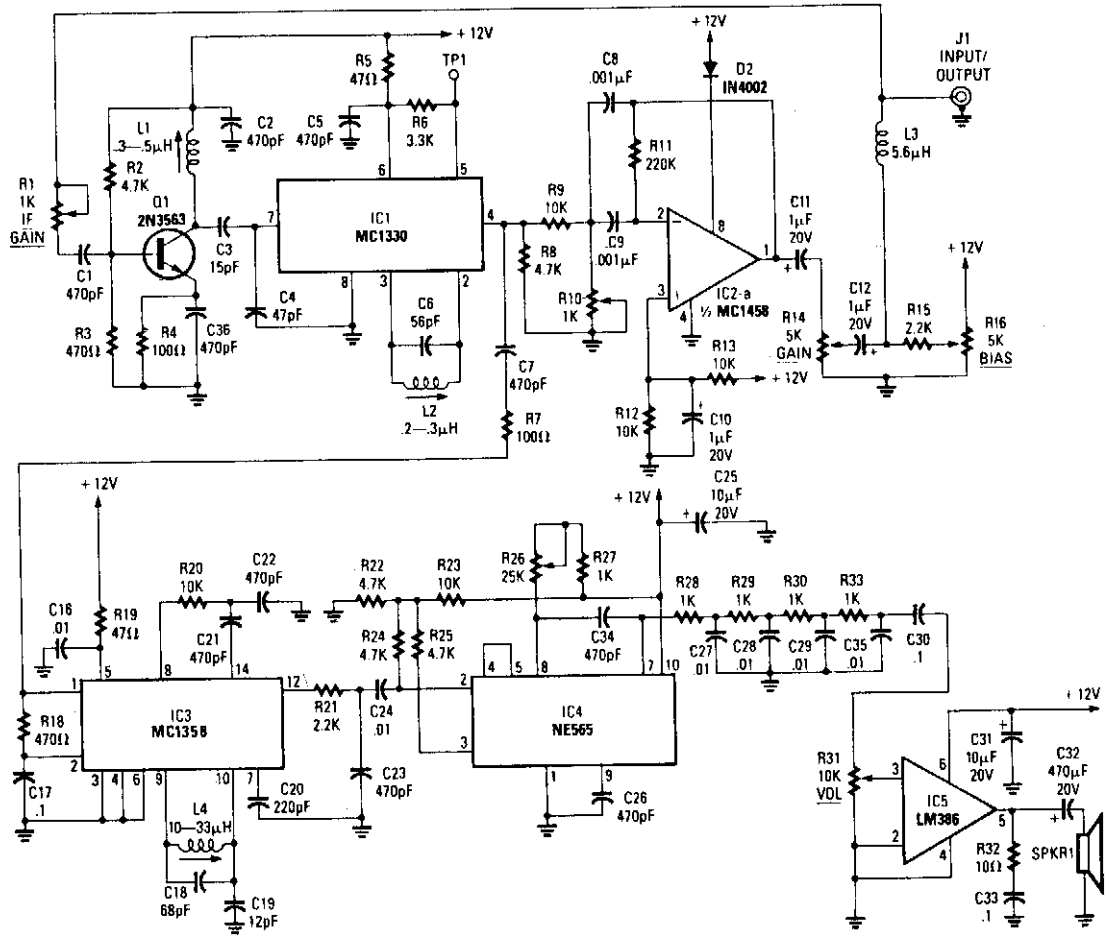
The NE564 is used as an FM demodulator. The connections for operation at 5 V and 12 V are shown in Figures 21-4A and 21-4B. The input signal is ac coupled with the output signal being extracted at Pin 14. Loop filtering is provided by the capacitors at Pins 4 and 5 with additional filtering being provided by the capacitor at Pin 14. Since the conversion gain of the VCO is not very high, to obtain sufficient demodulated output signal the frequency deviation in the input signal should be 1% or higher.

22
Descramblers and Decoders

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Sine Wave Descrambler
Outband Descrambler
Gated Pulse Descrambler
SCA Decoder
Dual Time Constant Tone-Decoder
Stereo TV Decoder
Time Division Multiplex (TDM) Stereo Decoder
Frequency Division Multiplex (FDM) Stereo Decoder
SCA (Background Music) Decoder

SINE WAVE DESCRAMBLER



—A COMPLETE SINEWAVE DESCRAMBLER. Easy to build, and relatively easy to align, this circuit completely removes the 15.75-kHz scrambling sinewave.

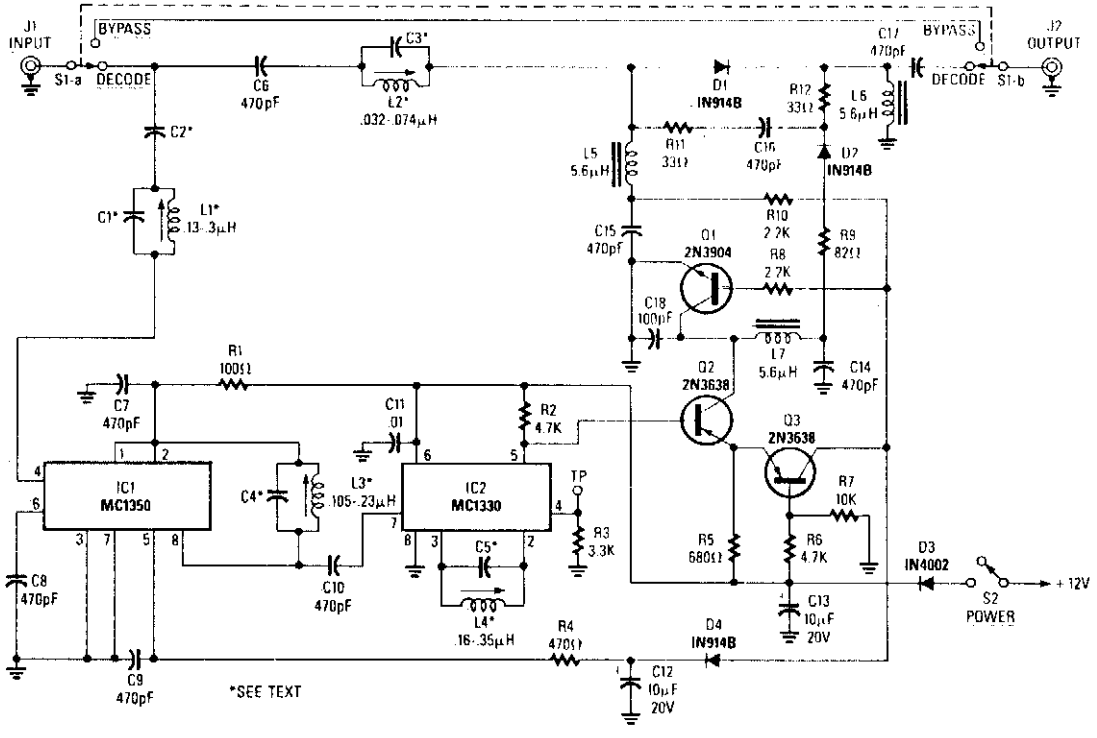
RADIO-ELECTRONICS

Fig. 22-1

Circuit Notes

This decoder features a sine wave recovery channel and uses a PIN diode attenuator driven by the sine wave recovery system to cancel out the sine wave sync suppression signal. Kit available from North Country Radio, P.O. Box 53, Wykagyl Station, New York 10804.

OUTBAND DESCRAMBLER



—FOR THE OUTBAND DECODER shown here to work, the cable company must provide at least a 1 millivolt signal. Values for C1–C5 and L1–L4 are found in Table 1.

TABLE 1—CAPACITOR AND COIL VALUES

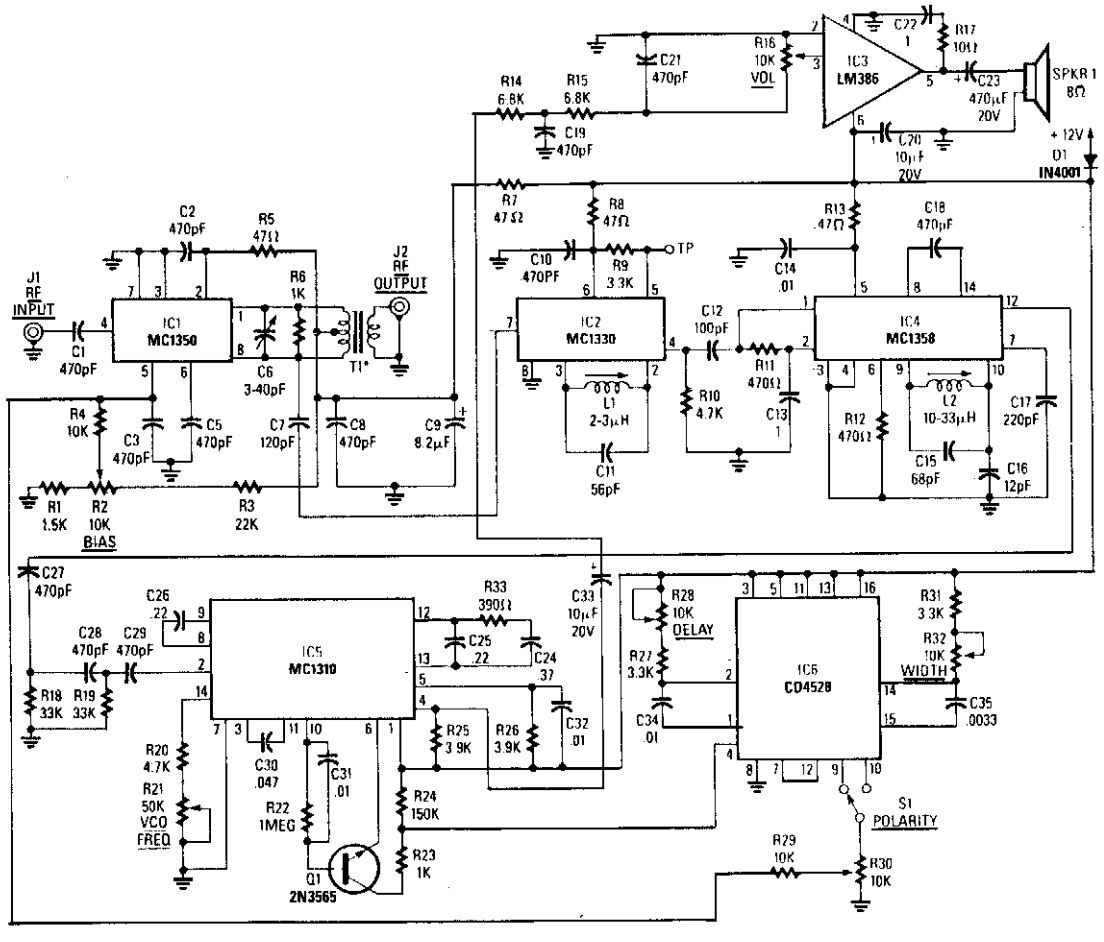
	50 MHz	90–114 MHz
C1	5 pF	5 pF
C2	47 pF	12 pF
C3	200 pF	82 pF
C4	56 pF	12 pF
C5	56 pF	10 pF
L1	0.2 µH	0.2 µH
L2	0.05 µH	0.03 µH
L3	0.175 µH	0.2 µH
L4	0.175 µH	0.24 µH

Fig. 22-2

Circuit Notes

This circuit consists of an amplifier for the synch channel and a video detector which controls an attenuator so that the gain of the systems is increased during synch intervals. Kit available from North Country Radio, P.O. Box 53, Wykagyl Station, New York 10804.

GATED PULSE DESCRAMBLER



—DESCRAMBLE GATED-PULSE SIGNALS using this easy-to-build circuit. Information for winding transformer T1 and coil L1 can be found in the text.

RADIO-ELECTRONICS

Fig. 22-3

Circuit Notes

This circuit consists of an amplifier and video detector with a second subcarrier detector for synch recovery purposes. A pulse-former circuit modulates the gain of the main channel increasing it during synch intervals. Provision for subcarrier audio descrambling is also provided. Kit available from North Country Radio, P.O. Box 53, Wykagyl Station, New York 10804.

SCA DECODER

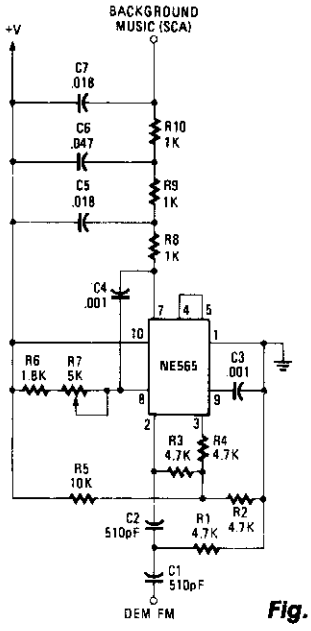


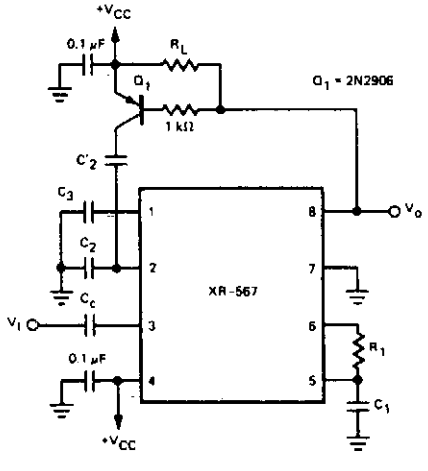
Fig. 22-4

RADIO-ELECTRONICS

Circuit Notes

The circuit uses a Signetics NE565 PLL (Phase-Locked Loop) as a detector to recover the SCA signal. The input to the SCA decoder circuit is connected to an FM receiver at a point between the FM discriminator and the de-emphasis filter network. The PLL, IC1, is tuned to 67 kHz by R7, a 5 K potentiometer. Tuning need not be exact since the circuit will seek and lock onto the subcarrier. The demodulated signal from the FM receiver is fed to the input of the 565 through a high-pass filter consisting of two 510 pF capacitors (C1 and C2) and a 4.7 K resistor (R1). Its purpose is to serve as a coupling network and to attenuate some of the main channel spill. The demodulated SCA signal at pin 7 passes through a three-stage de-emphasis network as shown. The resulting signal is around 50 mV, with the response extending to around 7 kHz.

DUAL TIME CONSTANT TONE DECODER



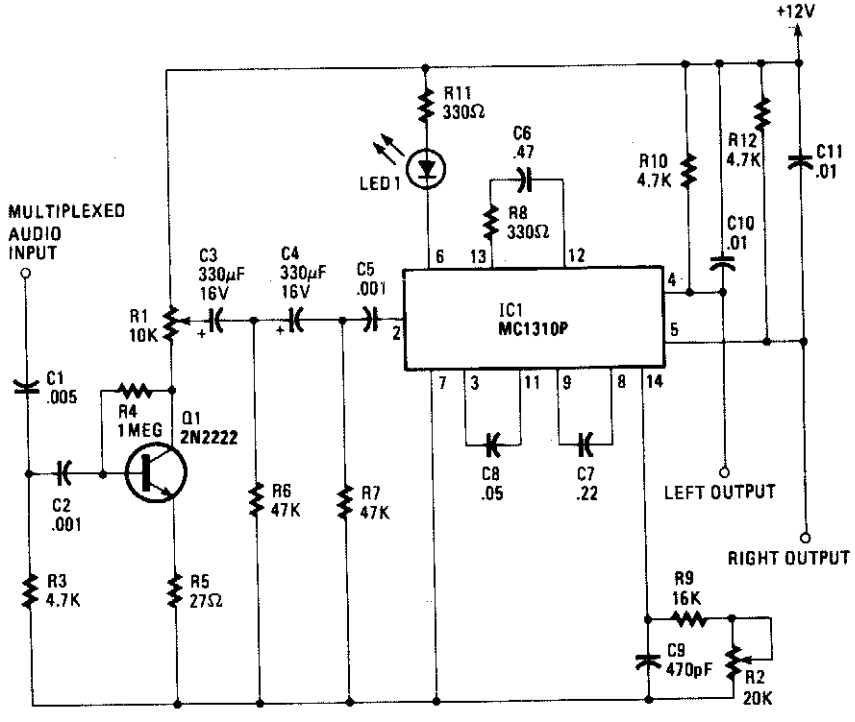
EXAR

Fig. 22-5

Circuit Notes

For some applications it is important to have a tone decoder with narrow bandwidth and fast response time. This can be accomplished by the dual time constant tone decoder circuit shown. The circuit has two low-pass loop filter capacitors, C_2 and C'_2. With no input signal present, the output at pin 8 is high, transistor Q_1 is off, and C'_2 is switched out of the circuit. Thus, the loop low-pass filter is comprised of C_2, which can be kept as small as possible for minimum response time. When an in-band signal is detected, the output at pin 8 will go low, Q_1 will turn on, and capacitor C'_2 will be switched in parallel with capacitor C_2. The low-pass filter capacitance will then be $C_2 + C'_2$. The value of C'_2 can be quite large in order to achieve narrow bandwidth. During the time that no input signal is being received, the bandwidth is determined by capacitor C_2.

STEREO TV DECODER



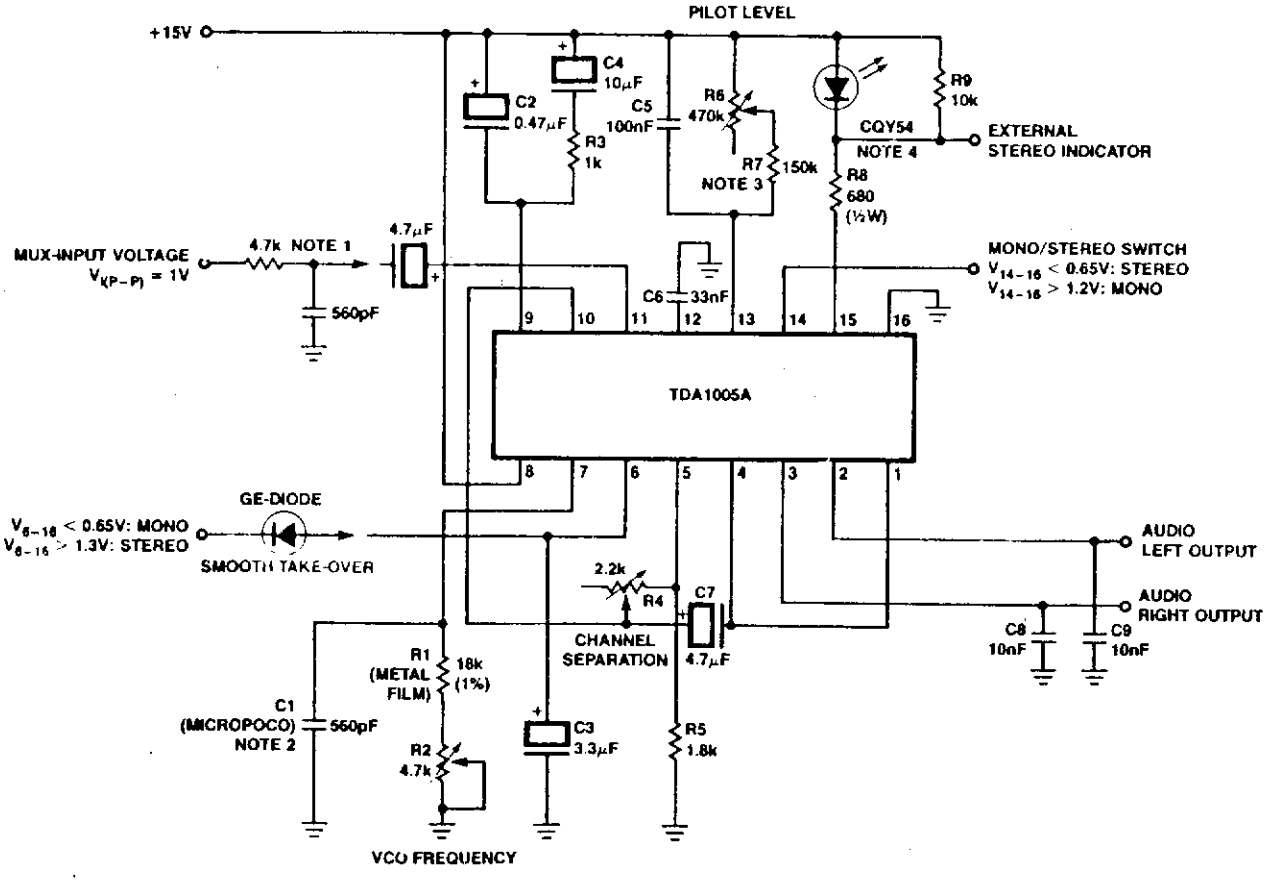
RADIO-ELECTRONICS

Fig. 22-6

Circuit Notes

The composite input signal is preamplified by transistor Q1 and is then coupled to the high-pass filter composed of C3, C4, R6, and R7. The filtered audio is then passed to IC1, an MC1310P "Coilless Stereo Demodulator." That IC is normally used to demodulate broadcast-band FM signals, but by changing the frequency of its on-board VCO (Voltage Controlled Oscillator) slightly (from 19 kHz to 15.734 kHz), we can use that IC to detect stereo-TV signals. A block diagram of the MC1310P is shown in Fig. 22-5. Notice that the components connected to pin 14 control the VCO's frequency, hence the pilot-detect and carrier frequencies. For use in an FM receiver, the VCO would run at four times the 19 kHz pilot frequency (76 kHz), but for our application, it will run at four times the 15.734 kHz pilot frequency of stereo TV, or 62.936 kHz. The MC1310P divides the master VCO signal by two in order to supply the 31.468 kHz carrier that is used to detect the L – R audio signal. The L – R signal undergoes normal FM detection, and at that point we've got two audio signals: L + R and L – R. The decoder block in the IC performs the addition and subtraction to produce the separate left and right signals. R10 and C10 form a de-emphasis network that compensates for the 75 μs pre-emphasis that the left channel underwent; R12 and C11 perform the same function for the right channel.

TIME DIVISION MULTIPLEX (TDM) STEREO DECODER



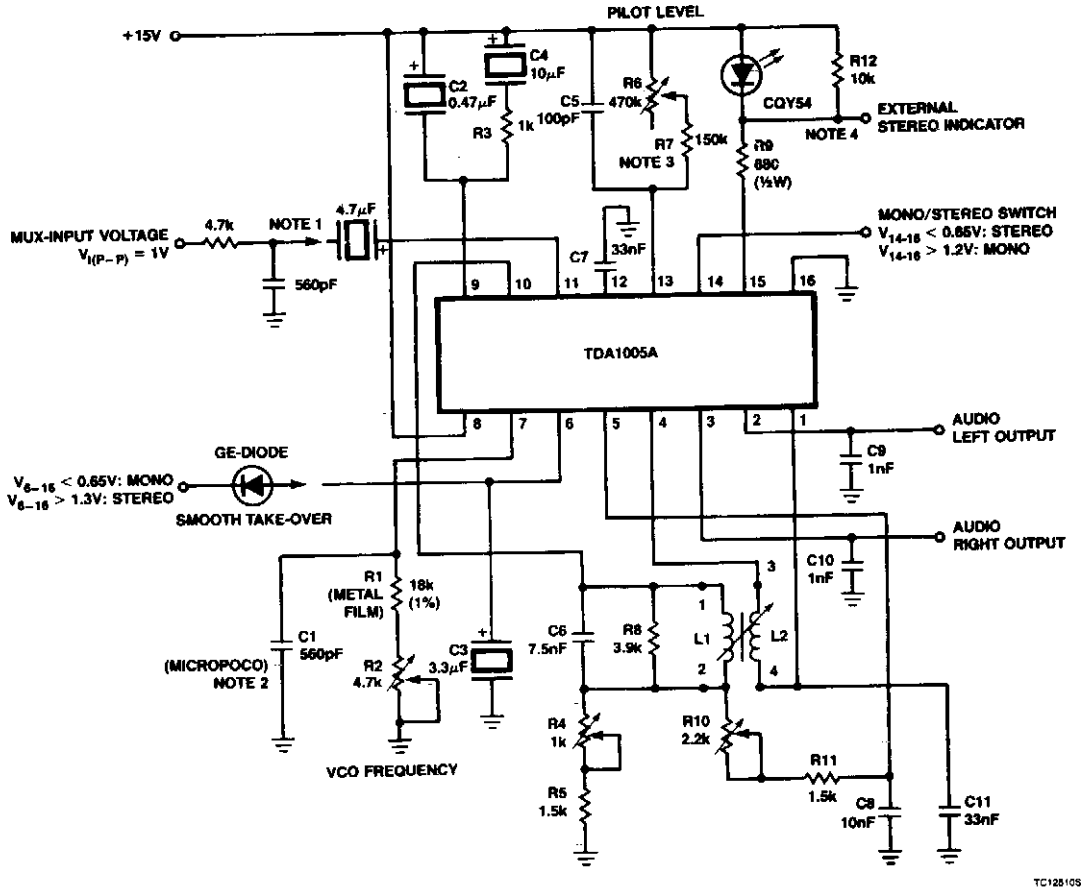
NOTES:

1. For other input structures see Figures 7 to 11, shown here with RC-filter (Figure 8).
2. Micropoco capacitor has a temperature coefficient of $125.10^{-6} \pm 60.10^{-6}°C^{-1}$.
3. In simplified circuits a fixed resistor (e.g. 620k) can be used for a guaranteed switching level of ≤ 16mV.
4. Either the LED circuit or an external stereo indicator can be used.

SIGNETICS

Fig. 22-7

FREQUENCY DIVISION MULTIPLEX (FDM) STEREO DECODER



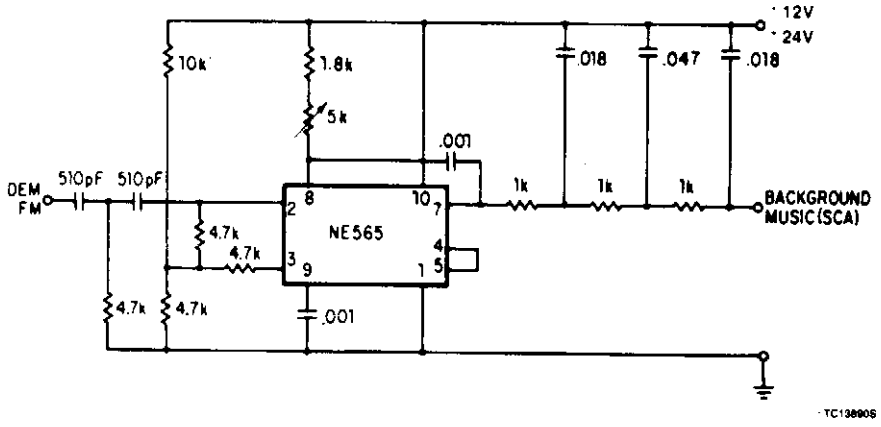
Coil data:
L_1 L_2 = 2.6mH
Q_{1-2} = 35; Q_{MIN} = 30
N_{1-2} = 357½ turns;
N_{3-4} = 297½ turns: scrambled wound with wire diameter 0.09mm, $\frac{E_{3-4}}{E_{1-2}} \times 100\% = 82\%$

NOTES:
1. For other input structures see Figures 6 to 11; shown here with RC-filter (Figure 8).
2. The micropoco capacitor has a temperature coefficient of $125.10^{-6} \pm 60.10^{-6}$ k^{-1}.
3. In simplified circuits a fixed resistor (e.g. 620k) can be used for a guaranteed switching level of ≤ 16mV.
4. Either the LED circuit or an external stereo indicator can be used.

SIGNETICS

Fig. 22-8

SCA (Background Music) DECODER



SIGNETICS

Fig. 22-9

Circuit Notes

A resistive voltage divider is used to establish a bias voltage for the input (Pins 2 and 3). The demodulated (multiplex) FM signal is fed to the input through a two-stage high-pass filter, both to effect capacitive coupling and to attenuate the strong signal of the regular channel. A total signal amplitude, between 80 mV and 300 mV, is required at the input. Its source should have an impedance of less than 10,000 ohm. The Phase-Locked Loop is tuned to 67 kHz with a 5000 ohm potentiometer, only approximate tuning is required since the loop will seek the signal. The demodulated output (Pin 7) passes through a three-stage low-pass filter to provide de-emphasis and attenuate the high-frequency noise which often accompanies SCA transmission. Note that no capacitor is provided directly at Pin 7; thus, the circuit is operating as a first-order loop. The demodulated output signal is in the order of 50 mV and the frequency response extends to 7 kHz.

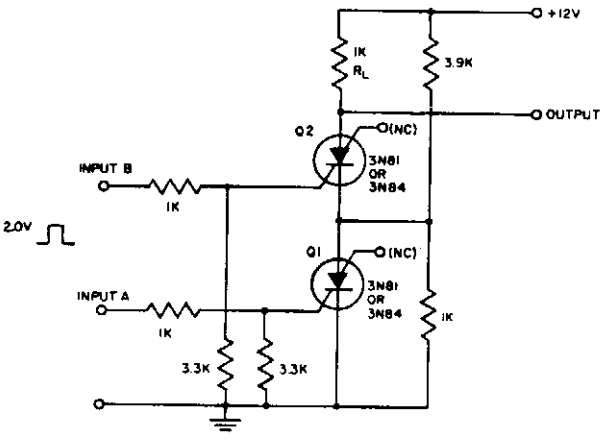
23

Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Pulse Sequence Detector
Voltage Level Detector
Zero-Crossing Detector
Peak Detector
Level Detector
High Frequency Peak Detector
Tachometer, Single Pulse Generator, Power Loss Detector, Peak Detector
Phase Detector with 10-Bit Accuracy
Frequency Limit Detector
Pulse Coincidence Detector

PULSE SEQUENCE DETECTOR



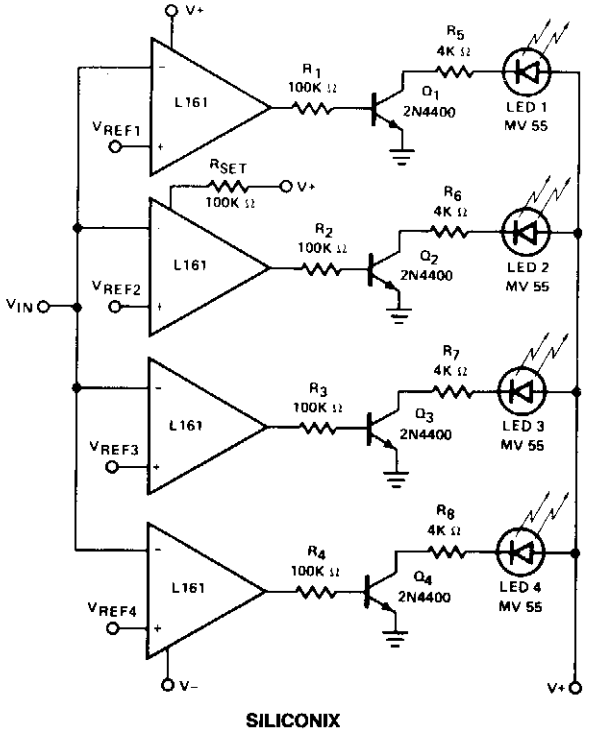
GENERAL ELECTRIC

Fig. 23-1

Circuit Notes

The resistor divider connected between Q1 and Q2 supplies I_H to Q1 after input A triggers it. It also prevents input B from triggering Q2 until Q1 conducts. Consequently, the first B input pulse after input A is applied will supply current to R_L.

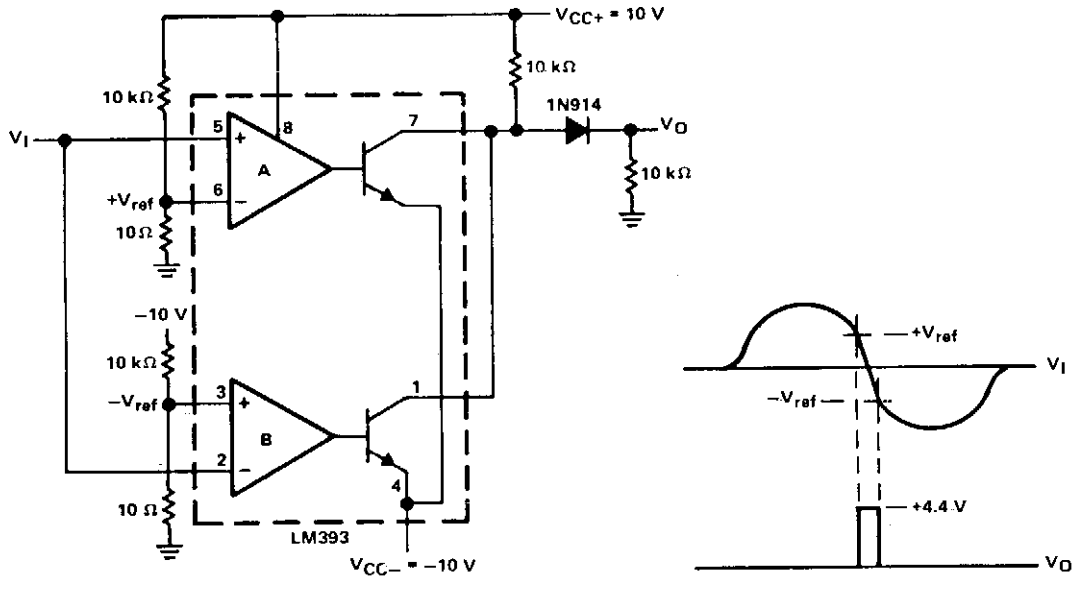
VOLTAGE LEVEL DETECTOR



SILICONIX

Fig. 23-2

ZERO-CROSSING DETECTOR



TEXAS INSTRUMENTS

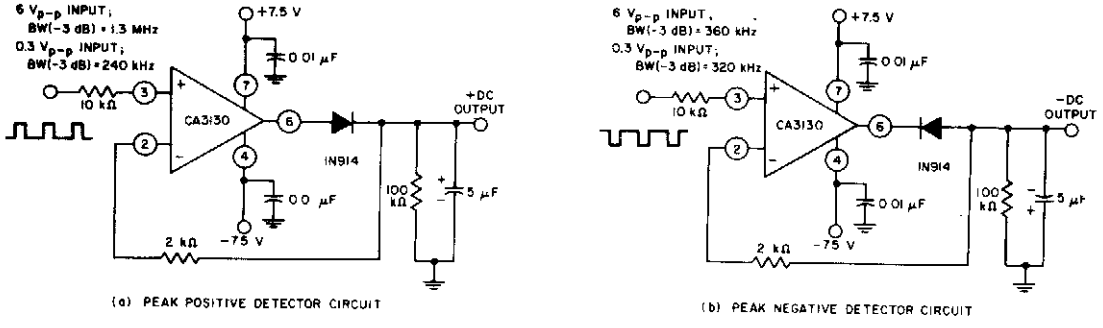
Fig. 23-3

Circuit Notes

This zero-crossing detector uses a dual LM393 comparator, and easily controls hysteresis by the reference levels which are set on the comparator inputs. The circuit illustrated is powered by ±10-V power supplies. The input signal can be an ac signal level up to +8 V. The output will be a positive going pulse of about 4.4 V at the zero-crossover point. These parameters are compatible with TTL logic levels.

The input signal is simultaneously applied to the non-inverting input of comparator A and the inverting input of comparator B. The inverting input of comparator A has a +10 mV reference with respect to ground, while the non-inverting input of comparator B has a −10 mV reference with respect to ground. As the input signal swings positive (greater than +10 mV), the output of comparator "A" will be low while comparator "B" will have a high output. When the input signal swings negative (less than −10 mV), the reverse is true. The result of the combined outputs will be low in either case. On the other hand, when the input signal is between the threshold points (±10 mV around zero crossover), the output of both comparators will be high. If more hysteresis is needed, the ±10 mV window may be made wider by increasing the reference voltages.

PEAK DETECTOR



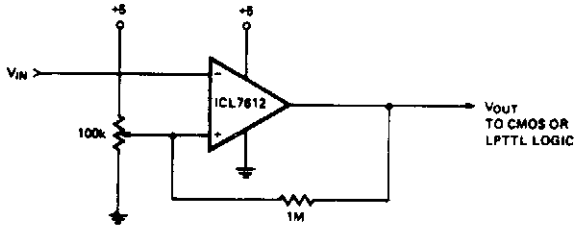
GENERAL ELECTRIC/RCA

Fig. 23-4

Circuit Notes

Circuits are easily implemented using the CA3130 BiMOS op amp. For large-signal inputs the bandwidth of the peak-negative circuit is less than that of the peak-positive circuit. The second stage of the CA3130 limits bandwidth in this case.

LEVEL DETECTOR



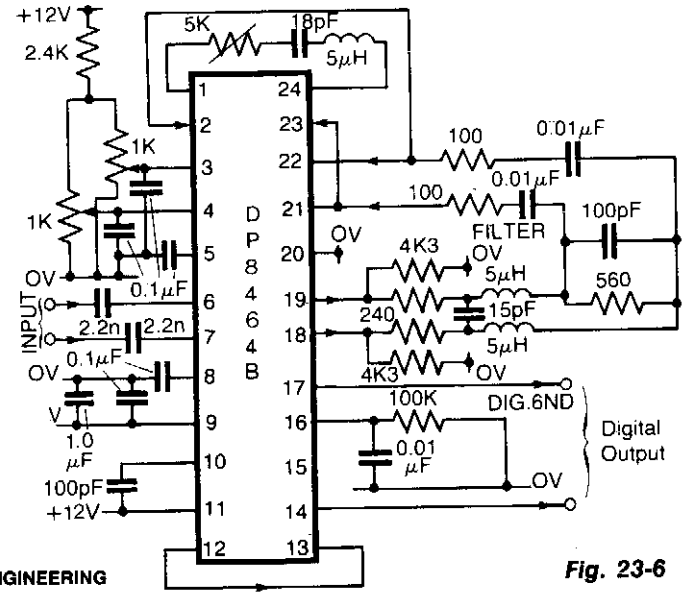
INTERSIL

Fig. 23-5

Circuit Notes

By using the ICL7612 in these applications, the circuits will follow rail to rail inputs.

HIGH FREQUENCY PEAK DETECTOR



ELECTRONIC ENGINEERING

Fig. 23-6

Circuit Notes

National Semiconductor's DP8464B is primarily intended for use in disk systems as a pulse detector. However it can be easily used as a general purpose peak detector for analogue signals up to 5 MHz. The chip can handle signals between 20 and 66 mV peak-to-peak. The circuit includes a filter with constant group delay characteristics to band limit the signal. Typically the −3 dB point for this filter will be at about 1.5 times the highest frequency of interest. This differentiator network between pins 1 and 24 can be as simple as a capacitor, or can be more complex to band limit the differentiator response.

TACHOMETER, SINGLE PULSE GENERATOR, POWER LOSS DETECTOR, PEAK DETECTOR

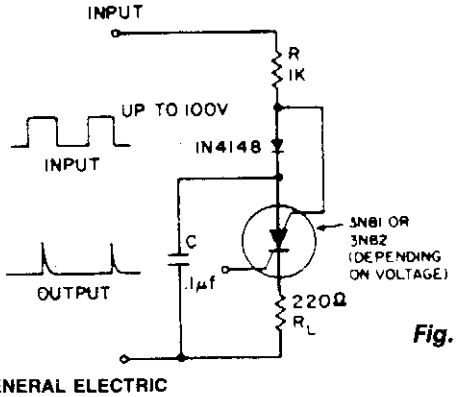
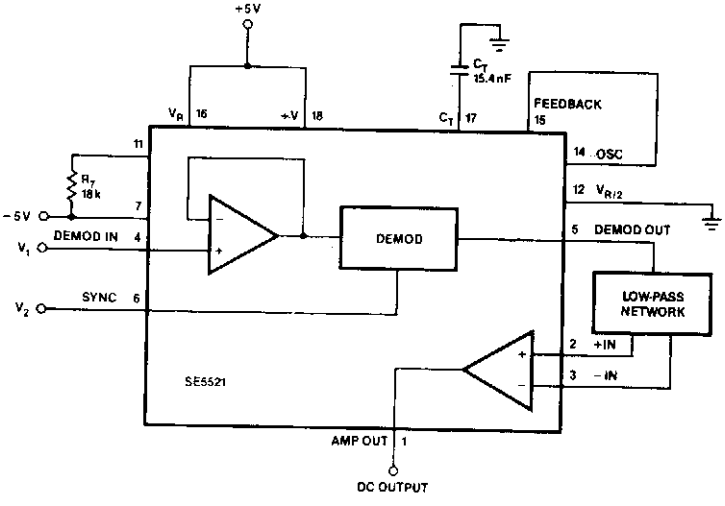


Fig. 23-7

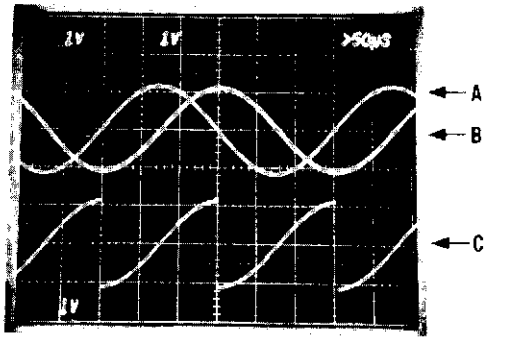
A positive going input charges C through the IN4148 and R. The diode keeps the scs off. A negative going input supplies anode-gate current triggering on the scs discharging C through R_L.

GENERAL ELECTRIC

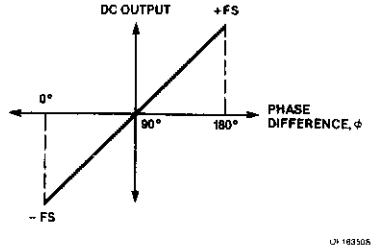
PHASE DETECTOR WITH 10-BIT ACCURACY



a. Phase Detector Measures Phase Difference Between Signals V_1 and V_2 and Provides dc Output at Pin 1



b. When V_1 and V_2 in (a) are at Quadrature (Traces A and B), the DC Component of Demodulator Output (Trace C) is at 0V



c. The dc Output and Phase Vary Linearly

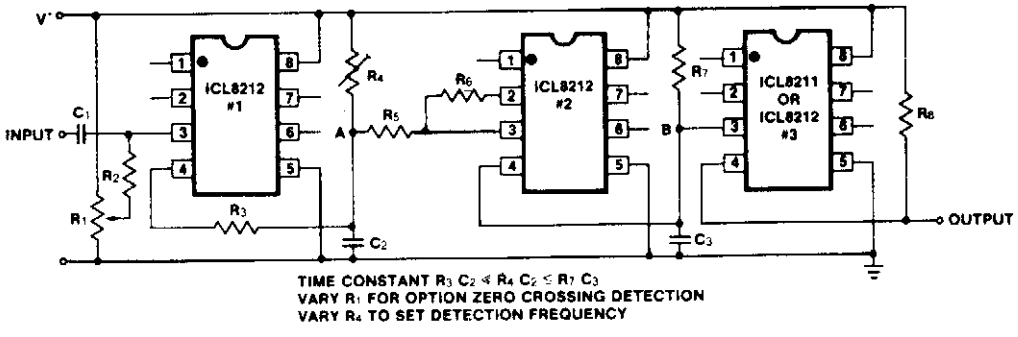
SIGNETICS

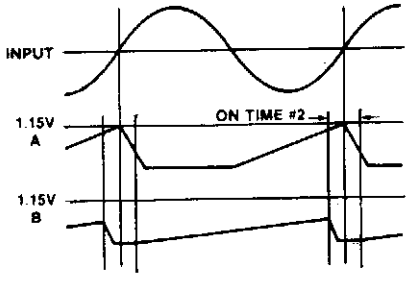
Fig. 23-8

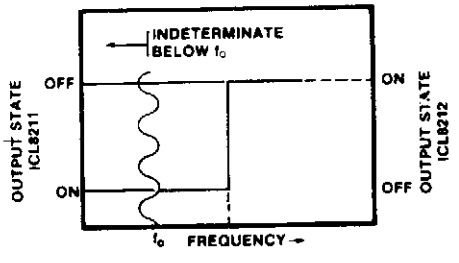
Circuit Notes

Signals of identical frequency are applied to sync input (Pin 6) and to the demodulator input (Pin 4), respectively, the demodulator functions as a phase detector with output dc component being proportional to phase difference between the two inputs. The signals must be referenced to 0 V for dual supply operation or to $V_R/2$ for single supply operation. At ± 5-V supplies, the demodulator can easily handle 7-V peak-to-peak signals. The low-pass network configured with the uncommitted amplifier dc output at Pin 1 of the device. The dc output is maximum (+ full-scale) when V_1 and V_2 are 180° out of phase and minimum (– full-scale) when the signals are in phase.

FREQUENCY LIMIT DETECTOR







INTERSIL

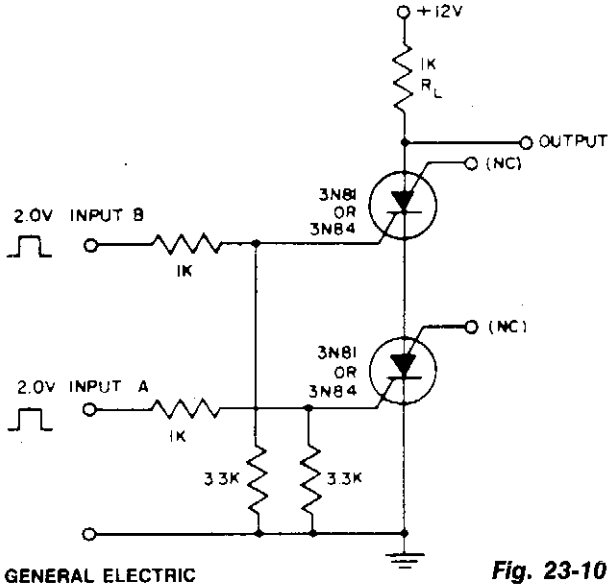
Fig. 23-9

Circuit Notes

Simple frequency limit detectors providing a GO/NO-GO output for use with varying amplitude input signals may be conveniently implemented with the ICL8211/8212. In the application shown, the first ICL8212 is used as a zero-crossing detector. The output circuit consisting of R_3, R_4 and C_2 results in a slow output positive ramp. The negative range is much faster than the positive range. R_5 and R_6 provide hysteresis so that under all circumstances the second ICL8212 is turned on for sufficient time to discharge C_3. The time constant of R_7C_3 is much greater than R_4C_2. Depending upon the desired output polarities for low and high input frequencies, either an ICL8211 or an ICL8212 may be used as the output driver.

The circuit is sensitive to supply voltage variations and should be used with a stabilized power supply. At very low frequencies the output will switch at the input frequency.

PULSE COINCIDENCE DETECTOR



GENERAL ELECTRIC

Fig. 23-10

Circuit Notes

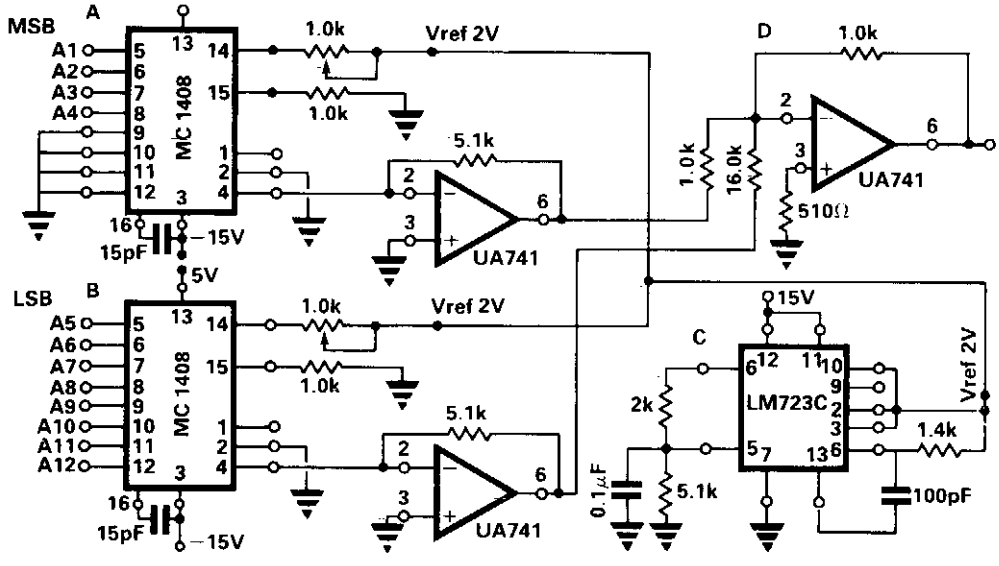
Unless inputs A and B (2- to 3-V amplitude) occur simultaneously no voltage exists across R_L. Less than 1 microsecond overlap is sufficient to trigger the scs. Coincidence of negative inputs is detected with gates G_A instead of G_C by using the scs in a complementary SCR configuration.

24
Digital-to-Analog Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Two 8-Bit DACs Make a 12-Bit DAC
12-Bit DAC with Variable Step Size

TWO 8-BIT DACS MAKE A 12-BIT DAC



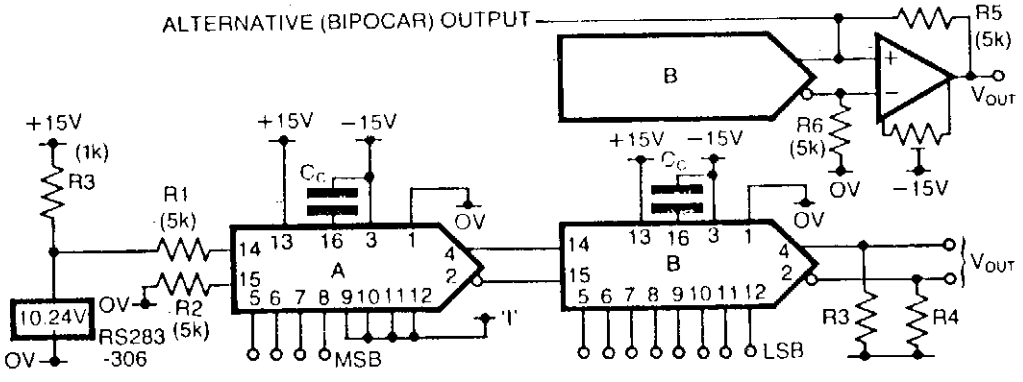
ELECTRONIC ENGINEERING

Fig. 24-1

Circuit Notes

Two MC1408—8-bit D/A converters, A and B in the circuit diagram, are used. The four least-significant bits of A are tied to zero. The four most significant bits of the 12-bit data are connected to the remaining four input pins. The eight least significant bits of the 12-bit data are connected to the eight input pins of B. The four most significant bits of the 12-bit data together have a weight of 16 relative to the remaining eight bits. Hence, the output from B is reduced by a factor of 16 and summed with the output from A using the summing op-amp configuration D. Voltage regulator chip, LM7236, is used to provide an accurate reference voltage, 2 V, for the MC1408. The full-scale voltage of the converter is $\frac{1}{16} \times 9.9609 + 1 \times (9.375)$ or 9.9976 V. The step size of the converter is 2.4 mV.

12-BIT DAC WITH VARIABLE STEP SIZE



ELECTRONIC ENGINEERING

Fig. 24-2

Circuit Notes

The step size of the converter is variable by selection of the high order data bits. The first DAC, A, has a stable reference current supplied via the 10.24 V reference IC and R1. R2 provides bias cancellation. As shown, only the first 4 MSB inputs are used, giving a step size of 225/256 × 2.048/16 = 0.127 mA. This current supplies the reference for DAC B whose step size is then 0.1275/256 = 0.498 μA. Complementary voltage outputs are available for unipolar output and using R3 = R4 = 10 K, V_{out} is ±10.2 V approximately, with a step size (1 LSB) of approximately 5 mV. If desired an op amp can be added to the output to provide a low impedance output with bipolar output symmetrical about ground, if R5 = R6 within 0.05%. Note that offset null is required, and all resistors except R2 and R3 should be 1% high stability types.

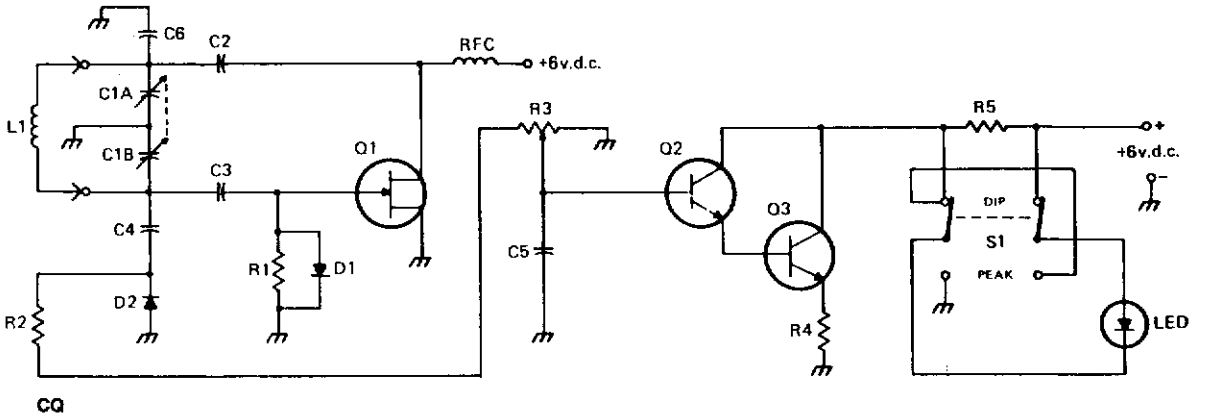
By using lower order address lines than illustrated for DAC A, a smaller step size (and therefore full-scale output) can be obtained. Unused high order bits can be manipulated high or low to change the relative position of the full-scale output.

25
Dip Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Little Dipper

LITTLE DIPPER



Parts List

L1—See coil data.
C1A,1B—Dual capacitor 100 pF per section (ETCO SV409 or similar)
C2,C3—100 pF mica, mylar, etc., low voltage
C4—10 pF mica, mylar, etc., low voltage
C5—.01 uF ceramic, low voltage
C6—5 pF mica, mylar, etc., low voltage
D1,D2—1N914 silicon diode or similar
R1—100 K ohms ¼ watt
R2—220 K ohms ¼ watt
R3—500 K ohms potentiometer
R4—10 ohms ¼ watt
R5—270 ohms ¼ watt
Q1—MPF 102 FET
Q2—Any general-purpose NPN transistor with a Beta (Hfe) of 40 or so (2N3904 or similar)
Q3—Any general-purpose NPN transistor capable of 20 mA collector current or more, Beta 40 or so (2N3904, 2N2222 or similar)
RFC—1 mH miniature ferrite core choke (value not critical)
LED—Panel mounting LED Radio Shack 276-068 or similar.
SW1—Sub-miniature DPDT slide switch or similar

Miscellaneous—6 volt AC adapter (Radio Shack 273-1454A)
Coaxial DC power jack (RS 274-1565)
Calibrated Dial knob (RS 274-413)
Dual phono jack (RS 274-332)

Fig. 25-1

Circuit Notes

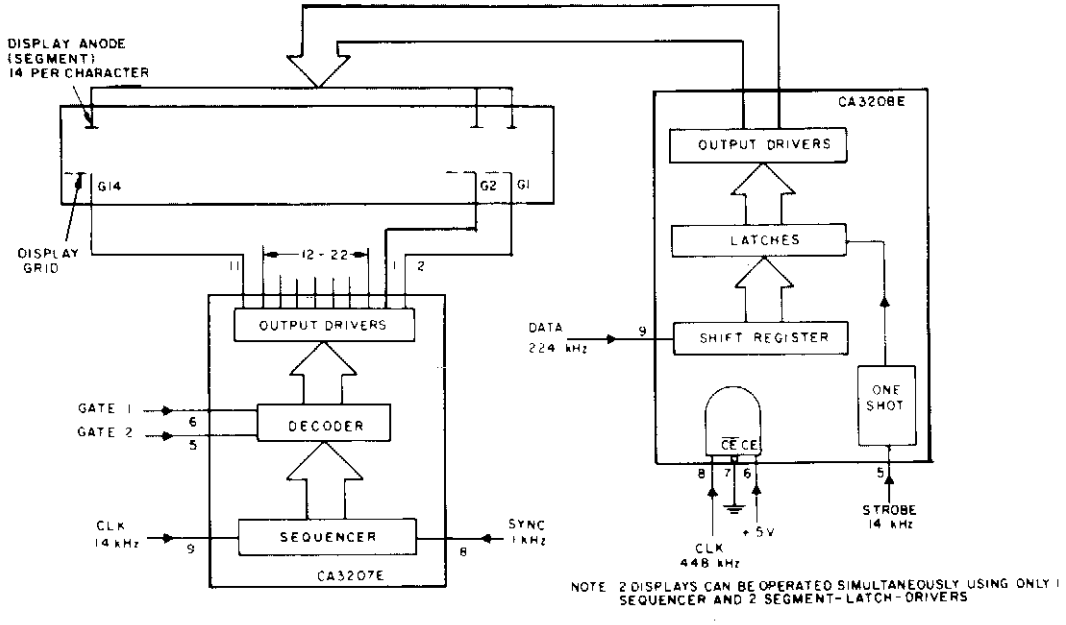
The circuit consists of two basic circuits, the oscillator and the detector. The oscillator uses an FET in a Colpitts configuration. The energy circulating in the oscillator tank is coupled through C4 to the detector circuit, where a small diode (D2) rectifies it, feeding a dc voltage to the Darlington pair (Q2, A3) controlled by the sensitivity control (R3). Any small variations in the bias of the amplifier will cause large variations of current through the LED indicator in the DIP mode; however, in the PEAK mode the current produces a corresponding voltage drop through R5 and the action of the LED is reversed. The circuit shown will work practically on any frequency from LF to VHF if the appropriate components are used.

26
Display Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Vacuum Fluorescent Display
Expanded Scale Meter, Dot or Bar
Low-Cost Bar-Graph Indicator for ac Signals
LED Bar-Graph Driver

VACUUM FLUORESCENT DISPLAY

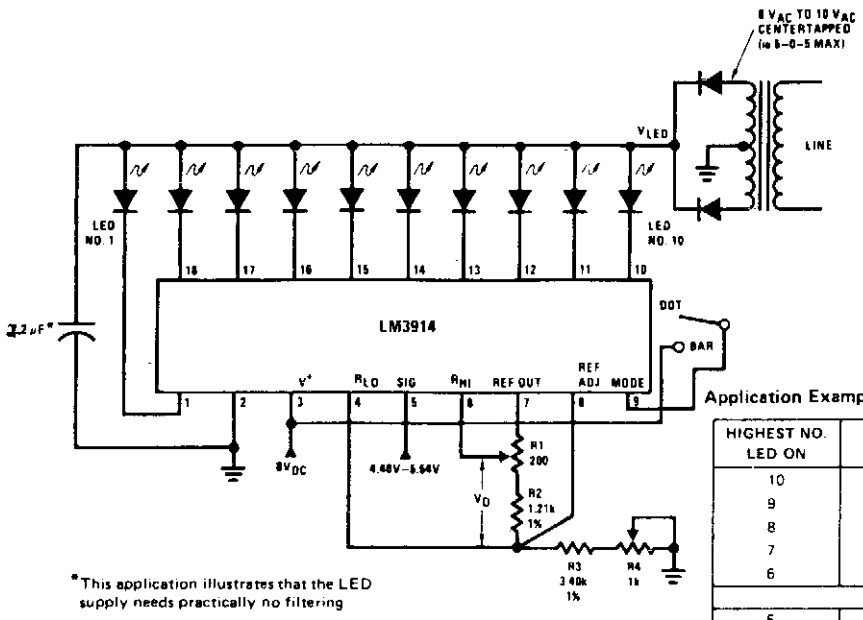


GENERAL ELECTRIC/RCA

Fig. 26-1

This circuit uses the CA3207 sequence driver and CA3208 segment latch-driver in combination to drive display devices of up to 14 segments with up to 14 characters of display. The CA3207 selects the digit or character to be displayed in sequence, CA3208 turns on the required alphanumeric segments.

EXPANDED SCALE METER, DOT OR BAR



*This application illustrates that the LED supply needs practically no filtering

Calibration: With a precision meter between pins 4 and 6 adjust R1 for voltage V_D of 1.20V. Apply 4.94V to pin 5, and adjust R4 until LED No. 5 just lights. The adjustments are non-interacting.

Application Example: Grading 5V Regulators

HIGHEST NO. LED ON	COLOR	VOUT(MIN)
10	Red	5.54
9	Red	5.42
8	Yellow	5.30
7	Green	5.18
6	Green	5.06
5V		
5	Green	4.94
4	Green	4.82
3	Yellow	4.7
2	Red	4.58
1	Red	4.46

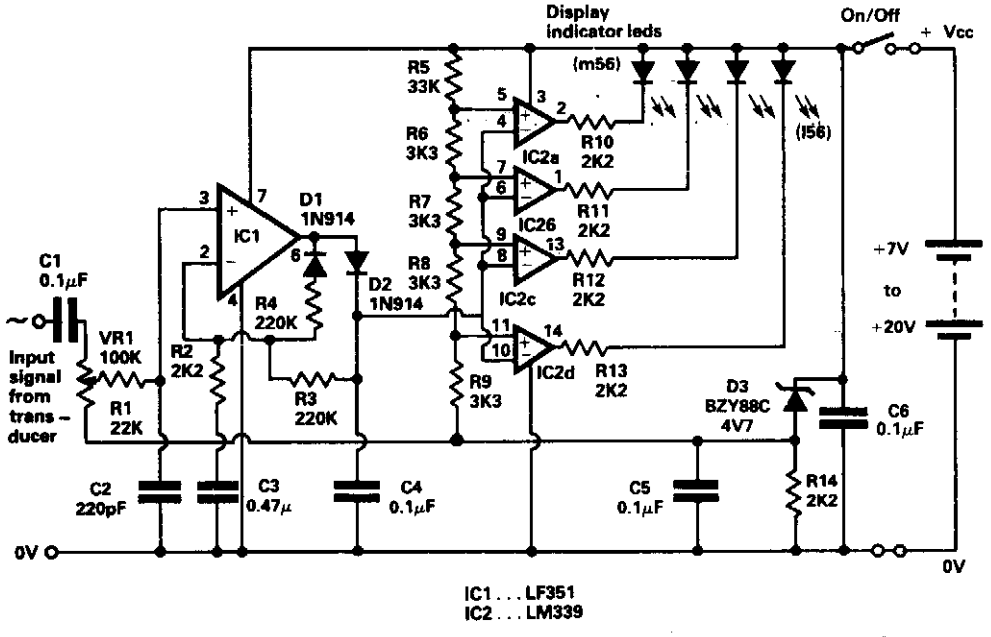
NATIONAL SEMICONDUCTOR CORP.

Fig. 26-2

Circuit Notes

A bar graph driver IC LM314 drives an LED display. The LEDs may be separate or in a combined (integral) bar graph display.

LOW-COST BAR-GRAPH INDICATOR FOR AC SIGNALS



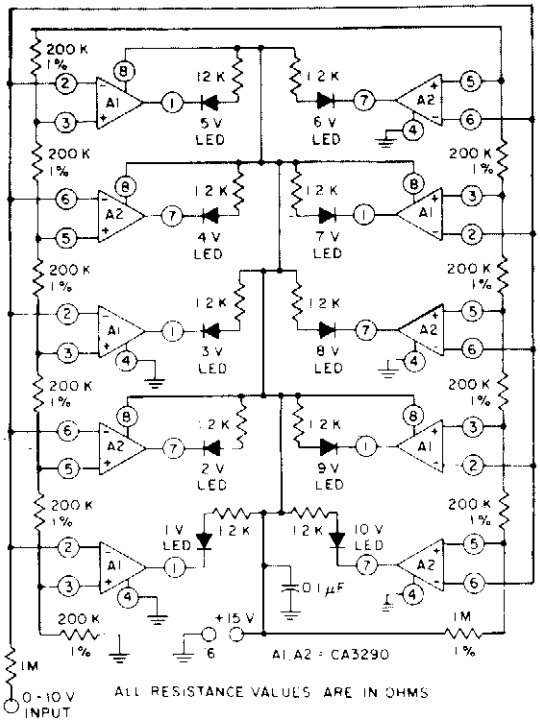
ELECTRONIC ENGINEERING

Fig. 26-3

Circuit Notes

Indicator was designed for displaying the peak level of small ac signals from a variety of transducers including microphones, strain gauges and photodiodes. The circuit responds to input signals contained within the audio frequency spectrum, i.e., 30 Hz to 20 kHz, although a reduced response extends up to 40 kHz. Maximum sensitivity for the component values shown, with VR1 fully clockwise, is 30 mV peak-to-peak. The indicator can be calibrated by setting VR1 when an appropriate input signal is applied.

LED BAR-GRAPH DRIVER



GENERAL ELECTRIC/RCA

Fig. 26-4

Circuit Notes

The circuit uses CA3290 BiMOS dual voltage comparators. Non-inverting inputs of A1 and A2 are tied to voltage divider reference. The input signal is applied to the inverting inputs. LEDs are turned "on" when input voltage the reaches the voltage on the reference divider.

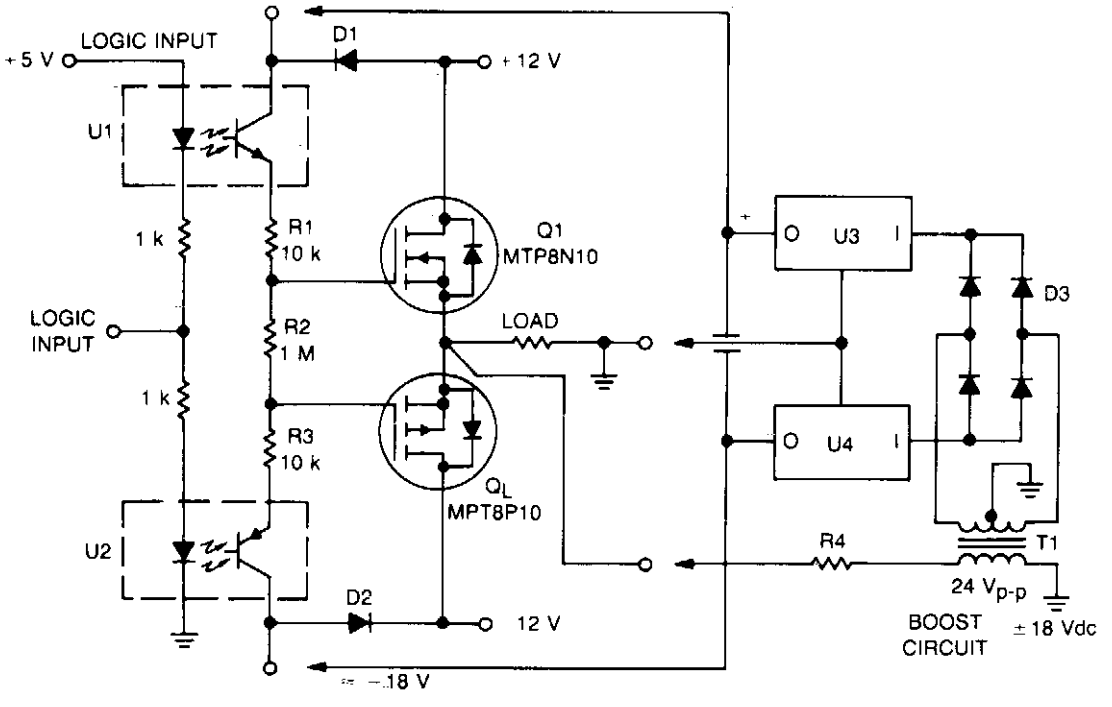
27

Drive Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Line Driver Provides Full Rail Excursions
Five Transistor Amplifier Boosts Fast Pulses into 50-Ohm Coaxial Cable
50-Ohm Transmission Line Driver
600-Ohm Balanced Driver for Line Signals
High Output 600-Ohm Line Driver

LINE DRIVER PROVIDES FULL RAIL EXCURSIONS



MOTOROLA

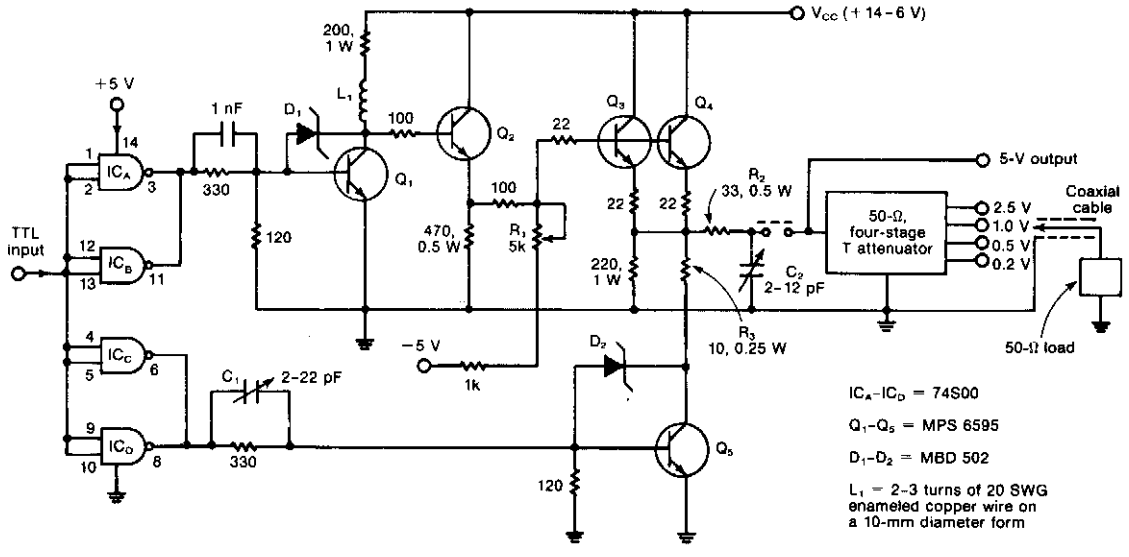
Fig. 27-1

Circuit Notes

The logic input is applied to opto-isolators U1 and U2 with, respectively, npn and pnp emitter follower outputs. Dc balance is adjusted by potentiometer R2. The emitter followers drive the gates of Q1 and Q2, the complementary TMOS pairs. With a ±12 V supply, the swing at the common source output point is about 12 V peak-to-peak.

By adding a ±18-V boost circuit, as shown, the output swing can approach the rail swing. This circuit applies the output to transformer T1, which is rectified by diode bridge D3, regulated by U3 and U4, and then applied to the collectors of U1 and U2. Diodes D1 and D2 are forward-biased when 12-V supplies are used, but they are back-biased when the 18-V boost is used.

FIVE-TRANSISTOR AMPLIFIER BOOSTS FAST PULSES INTO 50-OHM COAXIAL CABLE



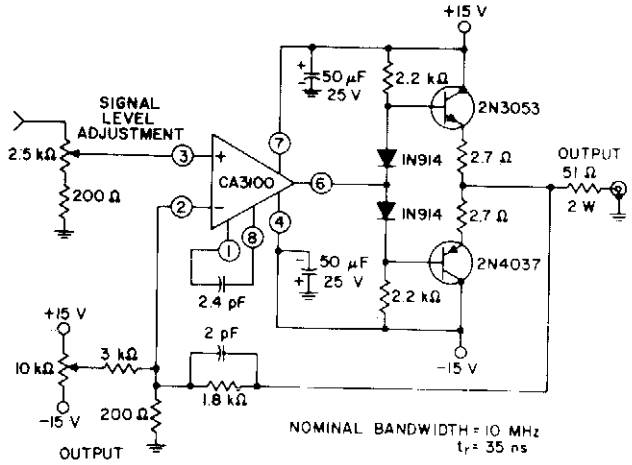
ELECTRONIC DESIGN

Fig. 27-2

Circuit Notes

The circuit works from dc to 50 MHz and will deliver pulses as short as 10 ns. It is driven by a TTL signal through a 740S00 quad Schottky NAND gate, IC_A through IC_D. Transistor Q1, wired as a common-emitter amplifier, drives transistor Q2, a simple emitter follower. Transistors Q3 and Q4, wired in parallel, also form an emitter follower and drive the output. When Q3 and Q4 are both turned off, transistor Q5 works as a low-impedance sink. Schottky diodes D1 and D2 prevent Q1 and Q5 from becoming saturated. To adjust the circuit, potentiometer R1 is set to optimize the output pulse's fall time. Inductor L1, a peaking coil, should be adjusted to improve the rise time to within a permissible 5% overshoot. Likewise, capacitor C1 can be varied to control preshooting. Further output pulse shaping is accomplished with the help of capacitor C2. Resistors R2 and R3 ensure a proper 50-ohm impedance at the amplifier's output when the pulse is on or off, respectively.

50-OHM TRANSMISSION LINE DRIVER



GENERAL ELECTRIC/RCA

Fig. 27-3

Circuit Notes

This circuit uses a wideband, high slew rate CA3100 BiMOS op amp. The slew rate for this amplifier is 28 V/μs. Output swing is 9 volts peak-to-peak into a terminated line, measured at the termination.

600-OHM BALANCED DRIVER FOR LINE SIGNALS

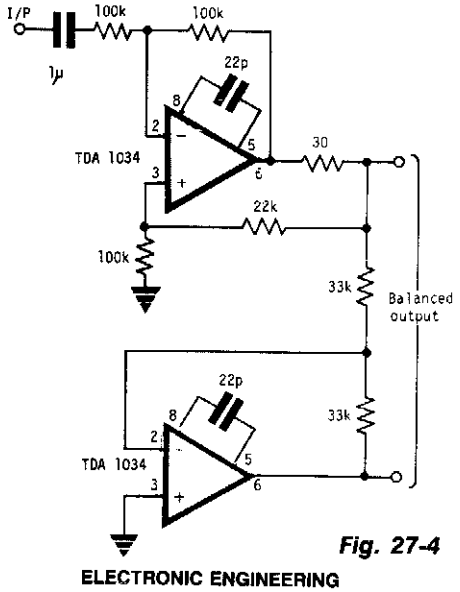


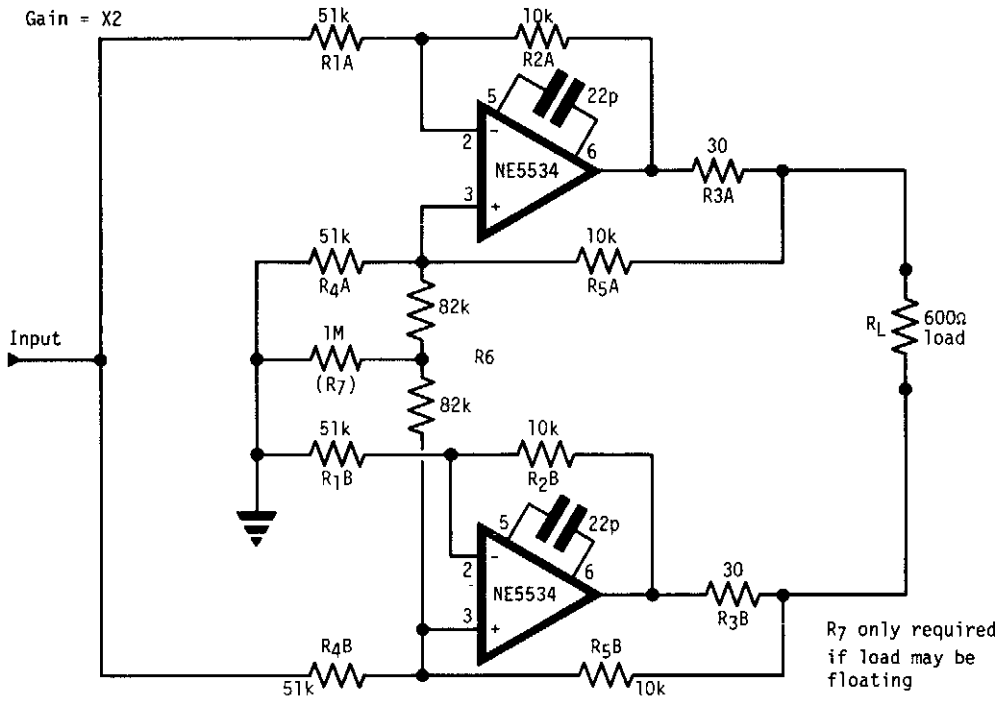
Fig. 27-4

ELECTRONIC ENGINEERING

Circuit Notes

This circuit uses current and voltage feedback. This circuit will handle +24 dBm with ±12 volts supply using TDA 1034s.

HIGH OUTPUT 600-OHM LINE DRIVER



ELECTRONIC ENGINEERING

Fig. 27-5

Circuit Notes

The circuit has a "floating" output, i.e., it behaves like an isolated transformer winding, with the output amplitude remaining unchanged whether the center or either end of the load is grounded. This is achieved by making Z-out, common mode, infinite. The circuit consists of two current-sources in push-pull. Since each has infinite output Z, the common mode output impedance is also infinite. Connecting a resistor between the non-inverting terminals of the op amps reduces the differential Z-out without affecting the Z-common-mode. Since the output is floating, if the load is also floating there is no output ground reference, which results in malfunction. This can be corrected by reducing the common-mode Z slightly. R7 fulfills this function. All resistors should be of close tolerance to give a good balance. The line driver provides +24 dB from ±12 V or +16 dB from ±6 V supplies.

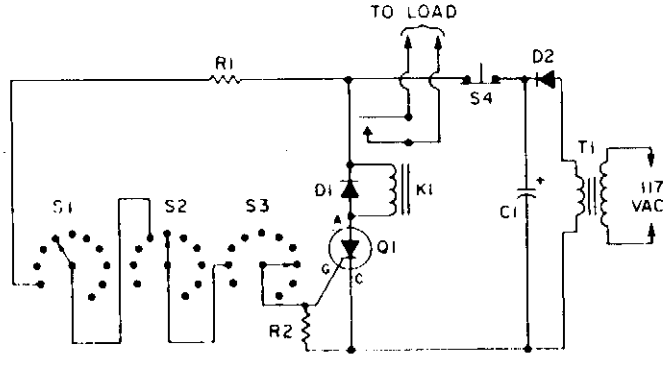
28
Electronic Locks

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Three-Dial Combination Lock
Electronic Combination Lock

THREE-DIAL COMBINATION LOCK

C1—500-uF, 25-VDC electrolytic capacitor
D1, D2—1N4002 diode
K1—relay with 6-volt coil rated @ 250-ohms, with SPST contacts
Q1—2N5050 SCR
R1, R2—4,700-ohm, ½-watt resistor, 5%
S1, S2, S3—single pole, 10-position rotary or thumbwheel switches
S4—normally closed SPST pushbutton switch
T1—120-VAC to 6.3-VAC @ 300mA power transformer



TAB BOOKS, INC.

Fig. 28-1

Circuit Notes

Here's an effective little combination lock that you can put together in one evening's time. To open the lock, simply dial in the correct combination on the three rotary or thumbwheel switches. With the correct combination entered, current flows through R1 into Q1's gate terminal, causing the SCR to latch in a conductive state. This sends a current through relay K1, which responds by closing its contacts and actuating whatever load is attached. After opening the lock, twirl the dials of S1 through S3 away from the correct combination so that nobody gets a look at it. The lock will remain open and your load will remain on because the SCR is latched on. To lock things up, it's only necessary to interrupt the flow of anode current through the SCR by pressing pushbutton S4.

ELECTRONIC COMBINATION LOCK

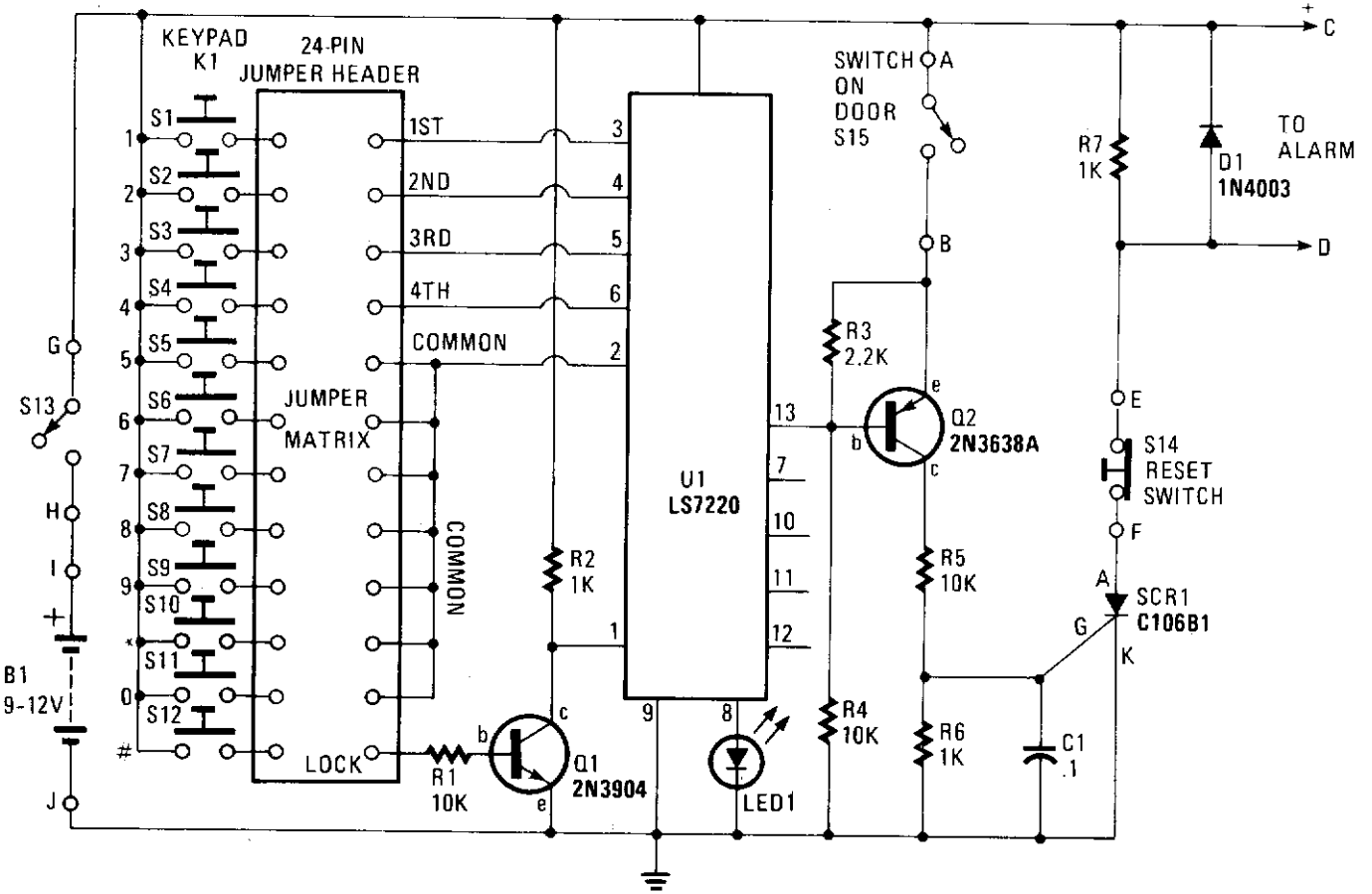


Fig. 28-2

Circuit Notes

When button S12 (#) is pressed, a positive voltage fed through R1 appears at the base of transistor Q1, turning it on. When Q1 is conducting, pin 1 of U1 is brought to ground (low) or the battery's negative terminal. With pin 1 low, two things occur: Pin 8 of U1 goes high (+9 volts dc), turning on LED 1—indicating that the circuit has been armed—and pin 13 goes from high to low. Transistor Q2 requires a low signal or negative voltage on its base in order to conduct. It also needs a positive voltage on its emitter and a negative voltage on the collector. As long as the door switch (S15) remains open (with the door itself closed), Q2's emitter will not receive the necessary positive voltage. If, however, an unauthorized person opens the door, thus closing switch S15 and placing a positive voltage on the emitter of Q1, the following sequence occurs:

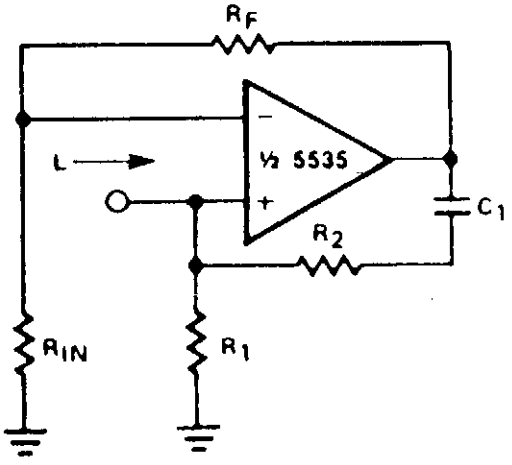
1. Transistor Q2 conducts, receiving the necessary biasing current through a current-divider network consisting of resistors R3 and R4.
2. As Q2 conducts, a voltage drop is developed across the voltage dividers made up of resistors R5 and R6. With R5 at 10,000 ohms and R6 at 1000 ohms, approximately one volt appears at the gate of SCR1. That's enough voltage to trigger the SCR's gate.

29
Emulator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Simulated Inductor
Resistor Multiplier
Capacitor Multiplier
JFET ac Coupled Integrator

SIMULATED INDUCTOR



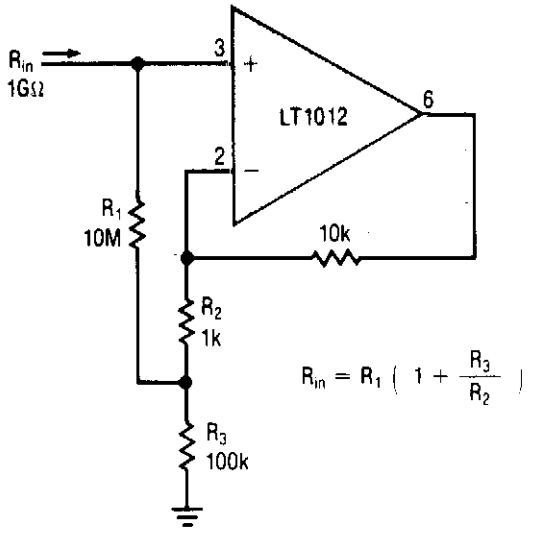
SIGNETICS

Fig. 29-1

Circuit Notes

With a constant current excitation, the voltage dropped across an inductance increases with frequency. Thus, an active device whose output increases with frequency can be characterized as an inductance. The circuit yields such a response with the effective inductance being equal to: $L = R_1R_2C$. The Q of this inductance depends upon R_1 being equal to R_2. At the same time, however, the positive and negative feedback paths of the amplifier are equal leading to the distinct possibility of instability at high frequencies. R_1 should therefore always be slightly smaller than R_2 to assure stable operation.

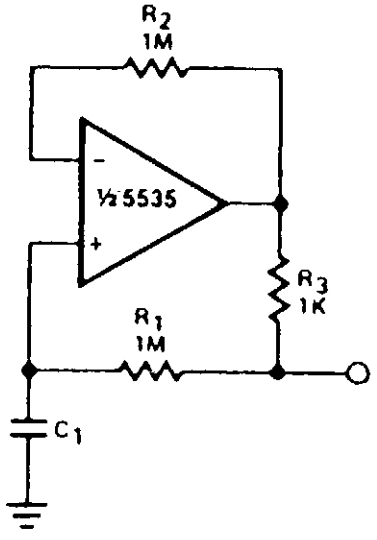
RESISTOR MULTIPLIER



LINEAR TECHNOLOGY CORP.

Fig. 29-2

CAPACITANCE MULTIPLIER



TC10620S

NOTE:
All resistor values are in ohms.

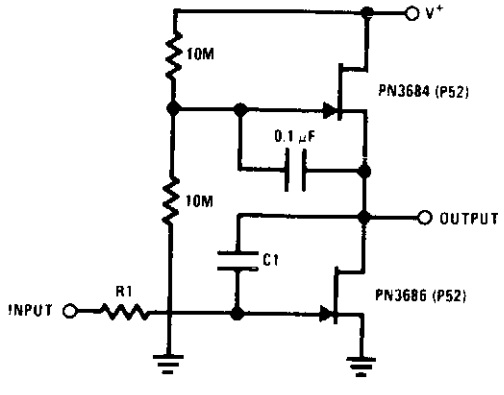
SIGNETICS

Fig. 29-3

Circuit Notes

The circuit can be used to simulate large capacitances using small value components. With the values shown and C - 10 μF, an effective capacitance of 10,000 μF was obtained. The Q available is limited by the effective series resistance. So R1 should be as large as practical.

JFET ac COUPLED INTEGRATOR



NATIONAL SEMICONDUCTOR CORP.

Fig. 29-4

Circuit Notes

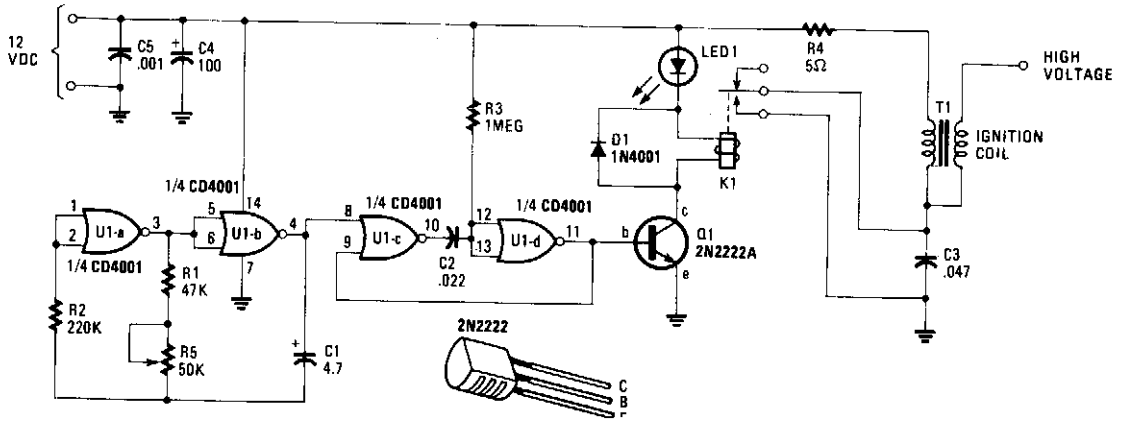
This circuit utilizes the "μ-amp" technique to achieve very high voltage gain. Using C1 in the circuit as a Miller integrator, or capacitance multiplier, allows this simple circuit to handle very long time constants.

30
Fence Chargers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Battery-Powered Fence Charger
Solid-State Electric Fence Charger
Electric Fence Charger

BATTERY-POWERED FENCE CHARGER



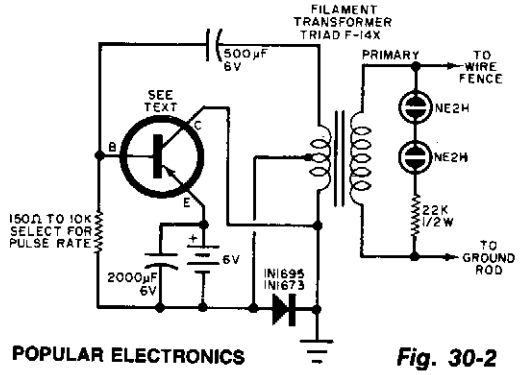
HANDS-ON ELECTRONICS

Fig. 30-1

Circuit Notes

In essence, the circuit is nothing more than an auto ignition coil and a set of points which accomplishes the same thing. A pulsing circuit made from a single CMOS NOR integrated circuit (U1), opens and closes the relay contacts to simulate the action of the original breaker points. The relay pulser is divided into two clocking functions. The first circuit is a free-running squarewave generator that determines the rate or frequency of the pulses that activate the relay. It is essentially a pair of NOR gates connected as inverters and placed in a feedback loop, they are U1-b. The oscillating period of the feedback loop is determined by timing components C1, R1, and variable resistor R5.

ELECTRIC FENCE CHARGER



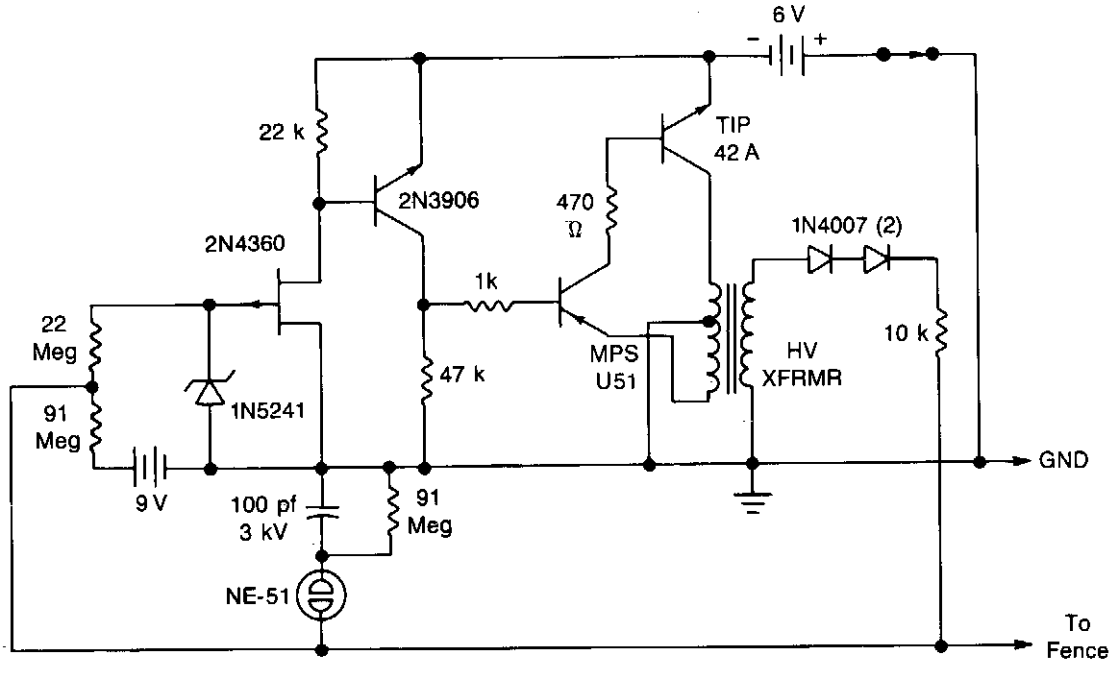
POPULAR ELECTRONICS

Fig. 30-2

Circuit Notes

Any good power transistor can be used in this circuit. The base resistor should be adjusted to obtain a pulse rate of about 50 pulses per minute. The range of values shown can go from 10 pulses to 100 pulses per minute. The single fence wire must be insulated at each supporting pole and should be mounted low enough to prevent an animal from crawling under the wire. The two neon lamps indicate when the unit is operating.

SOLID-STATE ELECTRIC FENCE CHARGER



WILLIAM SHEETS

Fig. 30-3

Circuit Notes

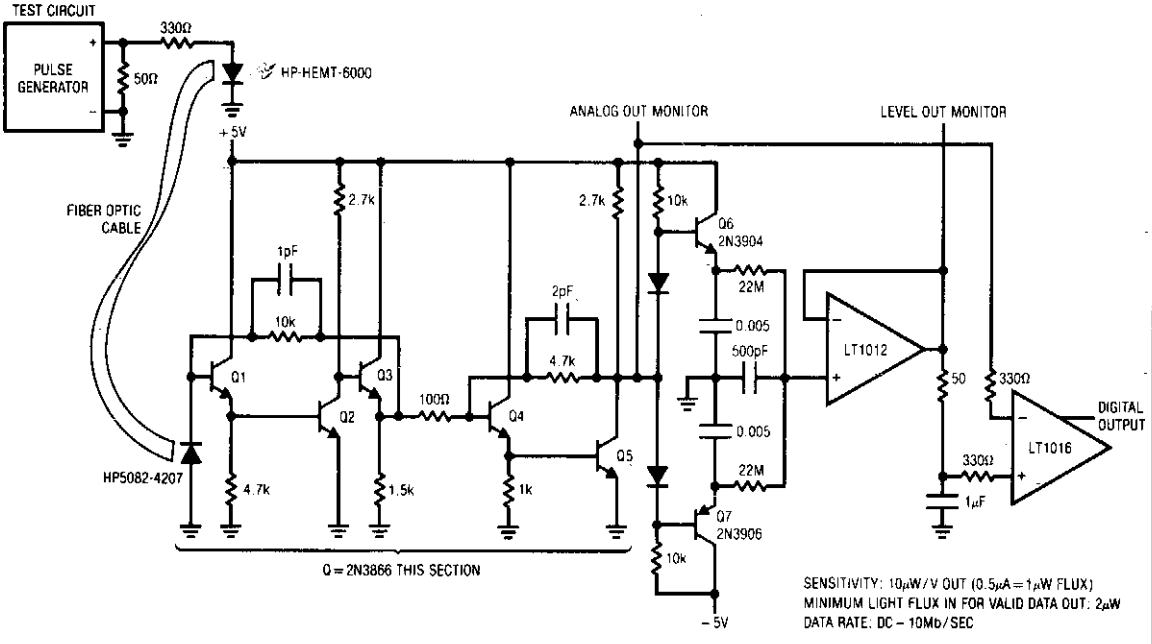
A touch-sensing circuit keeps the high-voltage generator cut off until something touches the fence wire. Contact with the fence sensing circuit wire starts the high-voltage generator which applies a series of 500 microsecond pulses at approximately 300 volts to the fence wire. Pulse repetition rate is determined by the intruder's resistance to earth ground. The lower the resistance, the higher the pulse rate. A ground rod is inserted several inches into the ground near the fence wire. In the sensing mode the neon lamp should not flicker or light. If it does, it indicates leakage between the fence wire and ground. If sensitivity is too great, it can be reduced by changing the 91 Meg resistor to 47 or 22 Meg as required.

31
Fiberoptics Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Fiberoptic Interface
10 MHz Fiberoptic Receiver
DC Variable Speed Motor Control via Fiberoptics

10 MHz FIBEROPTIC RECEIVER



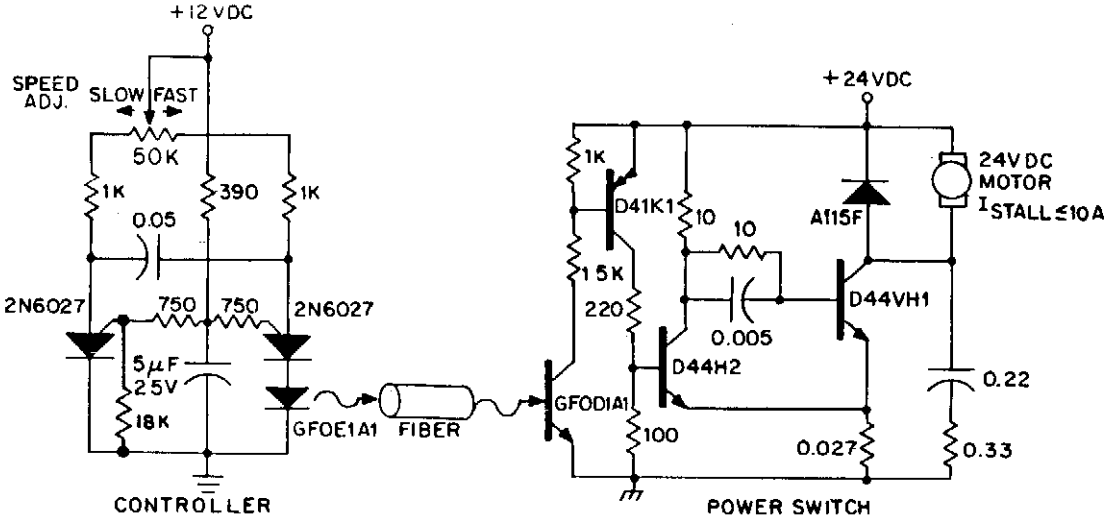
LINEAR TECHNOLOGY CORPORATION

Fig. 31-1

Circuit Notes

The receiver will accurately condition a wide range of light inputs at up to 10 MHz data rates. The optical signal is detected by the PIN photodiode and amplified by a broadband fed-back stage, Q1-Q3. A second, similar, stage gives further amplification. The output of this stage (Q5's collector) biases a 2-way peak detector (Q6-Q7). The maximum peak is stored in Q6's emitter capacitor while the minimum excursion is retained in Q7's emitter capacitor. The dc value of Q5's output signal's mid-point appears at the junction of the 0.005 μF capacitor and the 22 M ohm unit. This point will always sit midway between the signal's excursions, regardless of absolute amplitude. This signal-adaptive voltage is buffered by the low bias LT1012 to set the trigger voltage at the LT1016's positive input. The LT1016's negative input is biased directly from Q5's collector.

DC VARIABLE SPEED MOTOR CONTROL VIA FIBEROPTICS



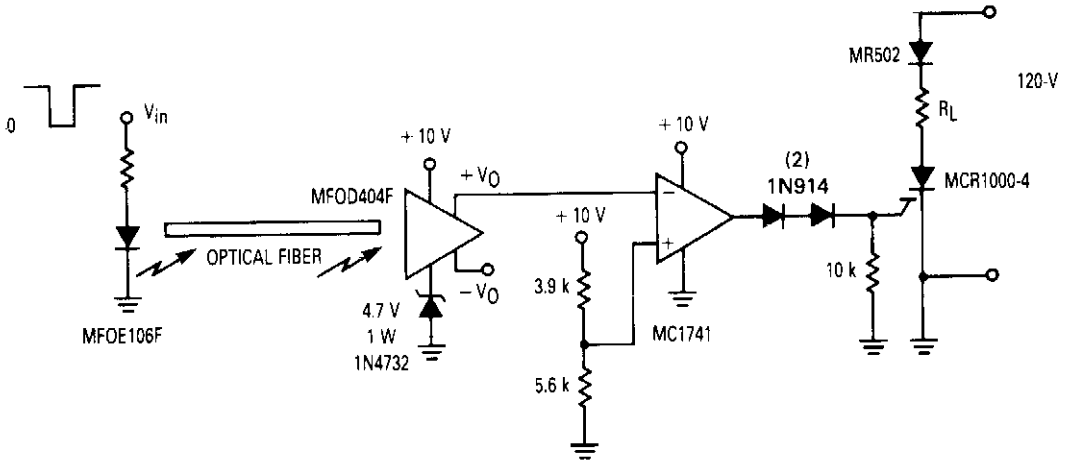
GENERAL ELECTRIC

Fig. 31-2

Circuit Notes

Dc power can also be controlled via fiberoptics. The circuit provides an insulated speed control path for a small dc actuator motor (≤ ½ hp). Control logic is a self-contained module requiring about 300 mW at 12 V, which can be battery powered. The control module furnishes infrared pulses, at a rate of 160 Hz, with a duty cycle determined by the position of the speed adjust potentiometer. The programmable unijunction multivibrator provides approximately 10 mA pulses to the GFOE1A1 at duty cycles adjustable over a range of 1% to 99%. The infrared pulses are detected by the GFOD1A1, amplified by the D39C1 pnp Darlington, and supplied to the power drive switch, which is connected in a Schmitt trigger configuration to supply the motor voltage pulses during the infrared pulses. Thus, the motor's average supply voltage is pulse width modulated to the desired speed, while its current is maintained between pulses by the A115F freewheeling diode. The snubber network connected in parallel with the power switch minimizes peak power dissipation in the output transistor, and enhancing reliability. Larger hp motors can be driven by adding another stage of current gain, while longer fiber range lengths can be obtained with an amplifier transistor driving the GFOE1A1.

FIBEROPTIC INTERFACE



MOTOROLA

Fig. 31-3

Circuit Notes

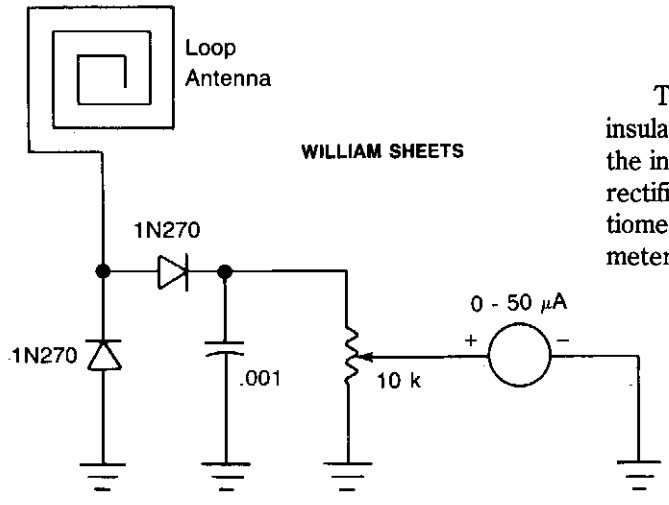
An op amp is used to interface between a fiberoptic system and the MOS SCR to multi-cycle, half-wave control of a load. This receiver has two complementary outputs, one at a quiescent level of about 0.6 V and the second at 3 V. By adding a 4.7 V zener in series with the return bus, the effective V_{CC} becomes 5.3 V and also the 0.6 V output level is translated up to about 5.3 V. This level is compatible with the reference input (5.9 V) of the single-ended powered op-amp acting as a comparator.

32
Field Strength Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Field Strength Meter
Field Strength Meter II
RF Sniffer
High Sensitivity Field Strength Meter
Transmission Indicator
LF or HF Field Strength Meter

FIELD-STRENGTH METER



Circuit Notes

The antenna consists of about 20 cm of insulated stranded wire glued or taped around the inside of a small plastic box. RF current is rectified by two diodes, and a 10 k potentiometer provides variable attenuation for the meter.

Fig. 32-1

FIELD-STRENGTH METER II

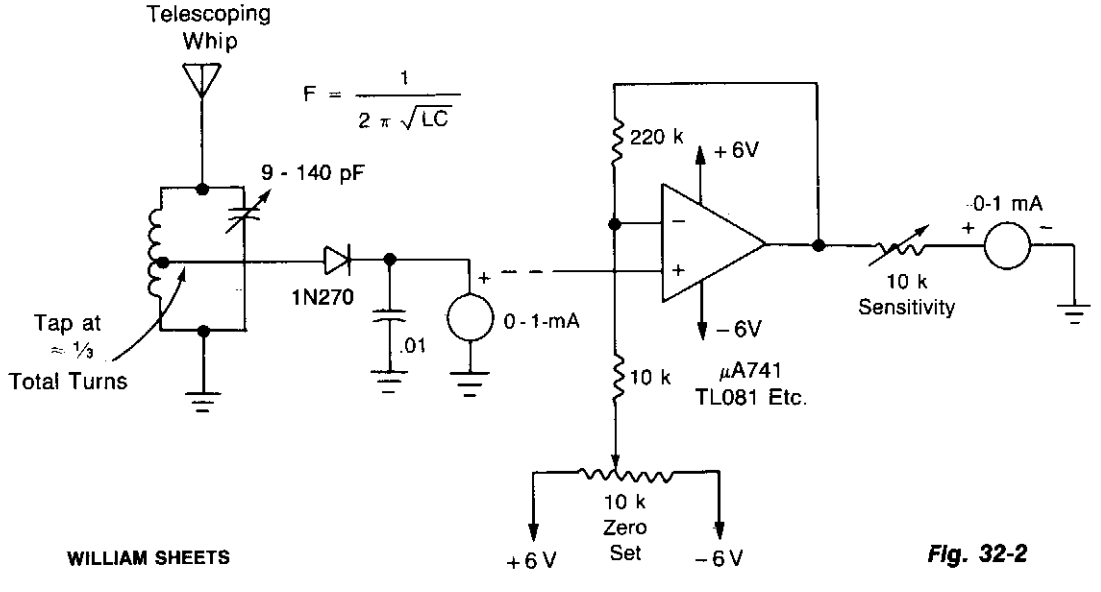
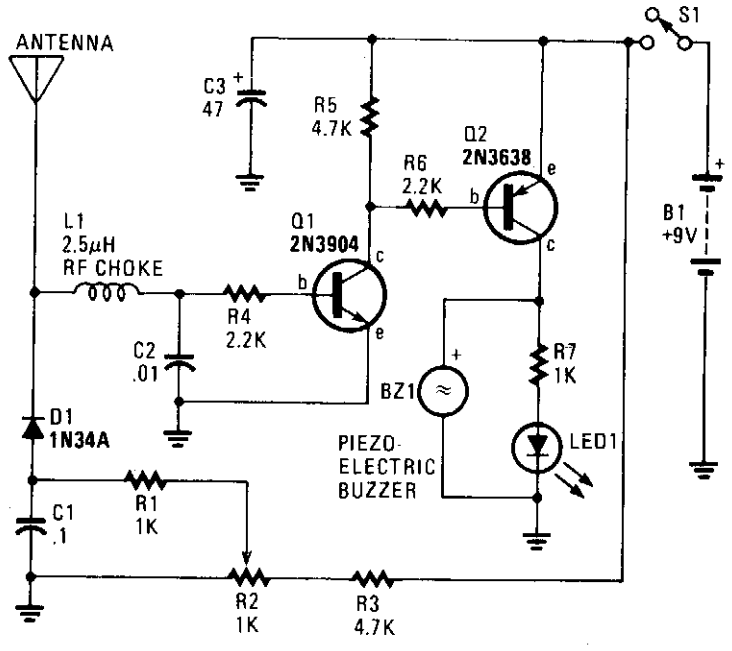


Fig. 32-2

Circuit Notes

"Minimum-parts" field-strength meter is shown here. For more distant testing, add the dc amplifier.

RF SNIFFER



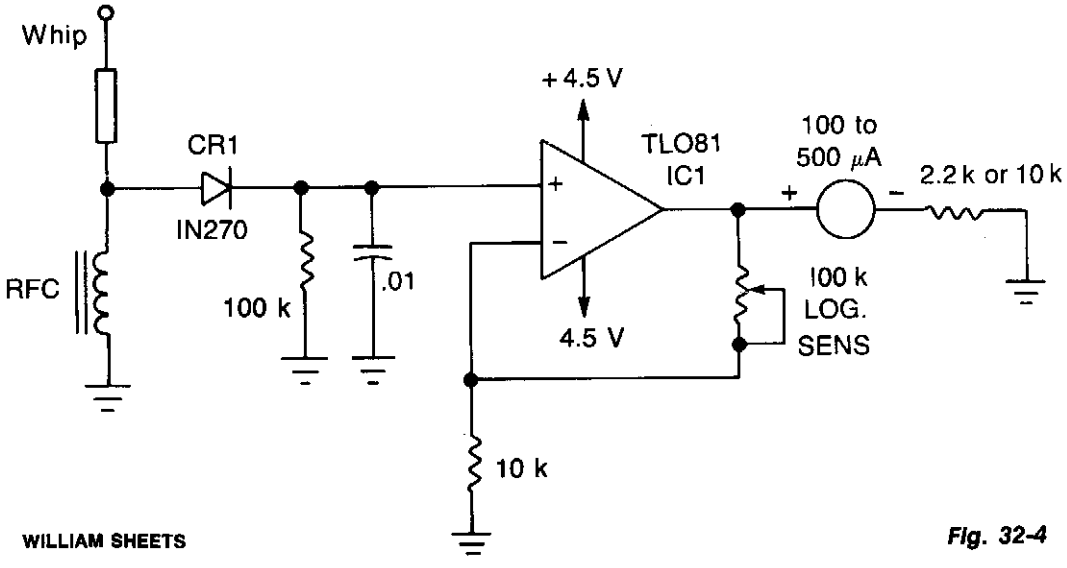
HANDS-ON ELECTRONICS

Fig. 32-3

Circuit Notes

This circuit responds to RF signals from below the standard broadcast band to well over 500 MHz, and provides a visual and audible indication when a signal is received. The circuit is designed to receive low-powered signals as well as strong sources of energy by adjusting the bias on the pick-up diode, D1, with R2. A very sensitive setting can be obtained by carefully adjusting R2 until the LED just begins to light and a faint sound is produced by the Piezo sounder.

HIGH-SENSITIVITY FIELD STRENGTH METER



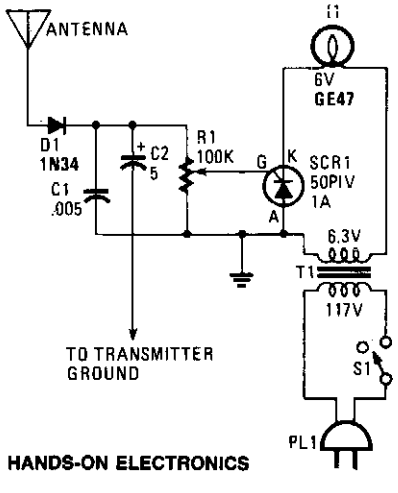
WILLIAM SHEETS

Fig. 32-4

Circuit Notes

A TL081 (IC1 op amp is used to increase sensitivity. RF signal is detected by CR1 and is then amplified by IC1. Full-scale sensitivity is set with the 100 K potentiometer.

TRANSMISSION INDICATOR



HANDS-ON ELECTRONICS

Fig. 32-5

Circuit Notes

Everytime the push-to-talk button is closed the light will go on. The antenna samples the output RF from the transmitter. That signal is then rectified (detected) by germanium diode D1, and used to charge capacitor C2. The dc output is used to trigger a small silicon-controlled rectifier (SCR1), which permits the current to flow through the small pilot lamp. For lower-power applications, such as CB radio, the antenna will have to be close-coupled to the antenna.

LF OR HF FIELD STRENGTH METER

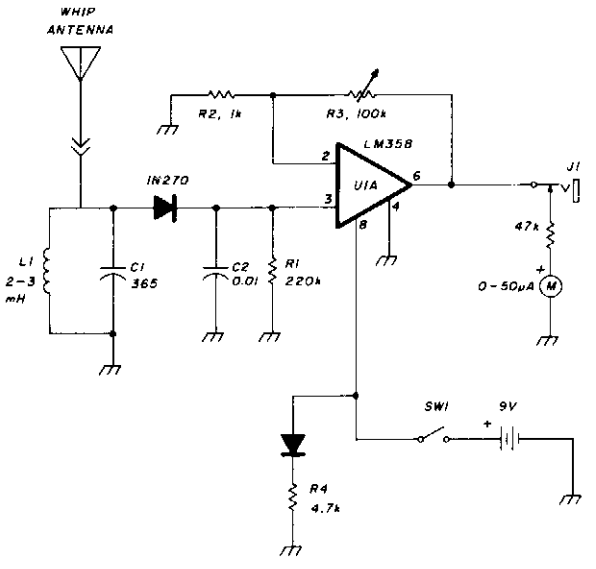


Table 1.

L1	C1 (variable)	Frequency Range	Ham Band
50 μH	30-365 pF	1- 4 MHz	160, 80 meters
3 μH	30-365 pF	5-16 MHz	40, 30, 20 meters
0.9 μH	30-365 pF	9-30 MHz	30, 20, 15, 12, 10 meters
2.5 mH	—	Broadband at reduced gain	

HAM RADIO

Fig. 32-6

Circuit Notes

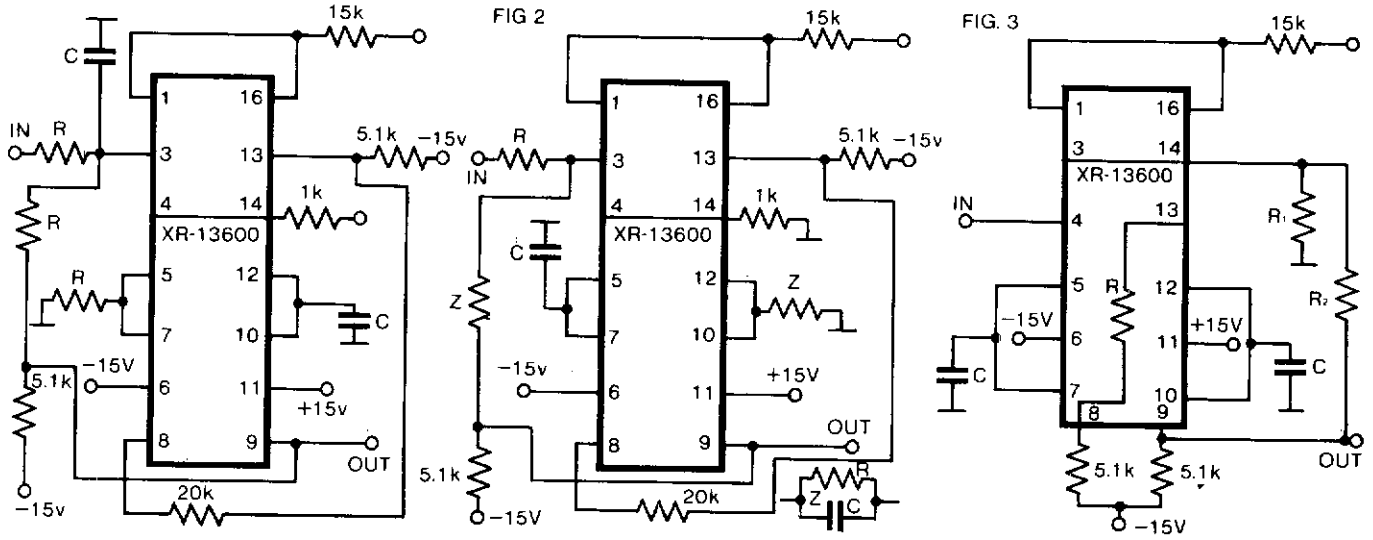
C1 and L1 resonate on the 1750 meter band, with coverage from 150 kHz to 500 kHz. L1 can be slug-tuned for 160-to-190 kHz coverage alone or a 2.5 mH choke can be used for L1, if desired, using C1 for tuning. A 1N270 germanium diode rectifies the RF signal and C2 is charged at the peak RF level. This dc level is amplified by an LM358. The gain is determined by R2 and R3, 1 100-kilohm linear potentiometer that varies the dc gain from 1 to 100, driving the 50 microampere meter. This field strength meter need not be limited to LF use. The Table shows the L1 and C1 values for HF operation and broadband operation.

33
Filter Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low Cost Universal Active Filter
State Variable Filter
Wideband Two-Pole High-Pass Filter
Active Low-Pass Filter with Digitally Selected Break Frequency
Digitally Tuned Low Power Active Filter
Razor Sharp CW Filter
Fifth Order Chebyshev Multiple Feedback Low-Pass Filter
Precision, Fast Settling Low-Pass Filter
Programmable Bandpass Using Twin-T Bridge
Active Bandpass Filter (f0 = 1000 Hz)
Bandpass Filter
Active Bandpass Filter
Bandpass and Notch Filter
Multiple-Feedback Bandpass Filter

LOW COST UNIVERSAL ACTIVE FILTER



ELECTRONIC ENGINEERING

Fig. 33-1

Circuit Notes

The circuit as shown in Fig. 1 gives the bandpass operation the transfer function calculated from

$$F_{BP}(s) = \frac{S/\omega_o}{K}$$

where $K = 1 + s/Q\omega_o + s^2/\omega_o^2$.
The cut-off frequency, ω_o, and the Q-factor are given by

$$\omega_o = g/C \text{ and } Q = gR/2$$

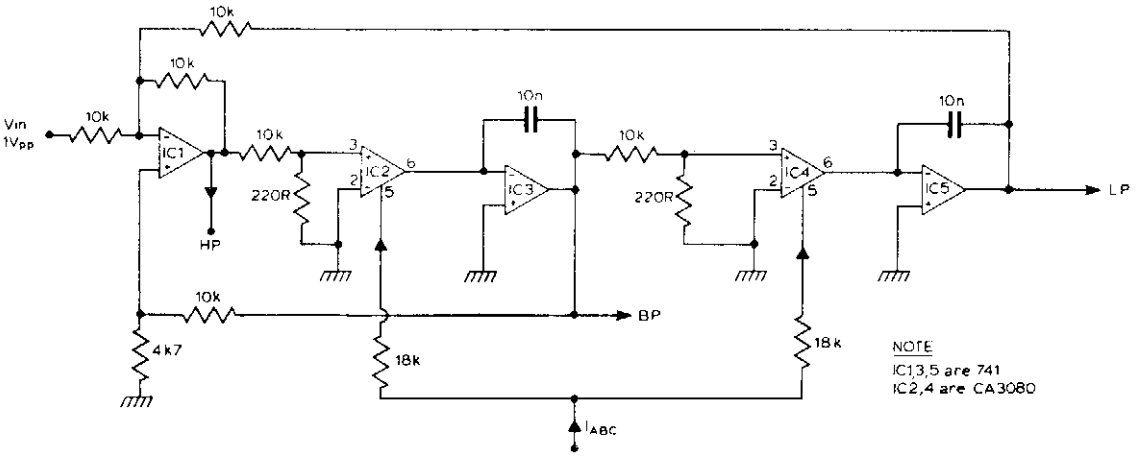
where g is the transconductance at room temperature.

Interchanging the capacitor C with the resistor R at the input of the circuit high-pass operation is obtained. A low-pass filter is obtained by applying two parallel connections of R and C as shown in Fig. 2.

The low-pass operation may be much improved with the circuit as given in Fig. 3. Here the gain and Q may be set up separately with respect to the cut-off frequency according to the equations

$$Q = 1/fB = 1 + R_2/R_1,$$
$$A = Q^2 \text{ and } \omega_o = g\ fB/C.$$

STATE VARIABLE FILTER



ELECTRONIC TODAY INTERNATIONAL

Fig. 33-2

Circuit Notes

The filter produces three outputs: high-pass, bandpass, and low-pass. Frequency is linearly proportional to the gain of the two integrators. Two CA3080's, (IC2, 4) provide the variable gain, the resonant frequency being proportional to the current I_{ABC}. Using 741 op amps for IC3 a control range of 100 to 1, (resonant frequency) can be obtained. If CA3140's are used instead of 741's then this range can be extended to nearly 10,000 to 1.

WIDEBAND TWO-POLE HIGH-PASS FILTER

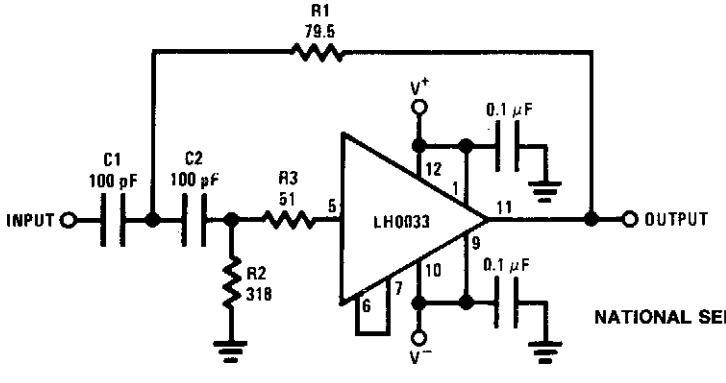


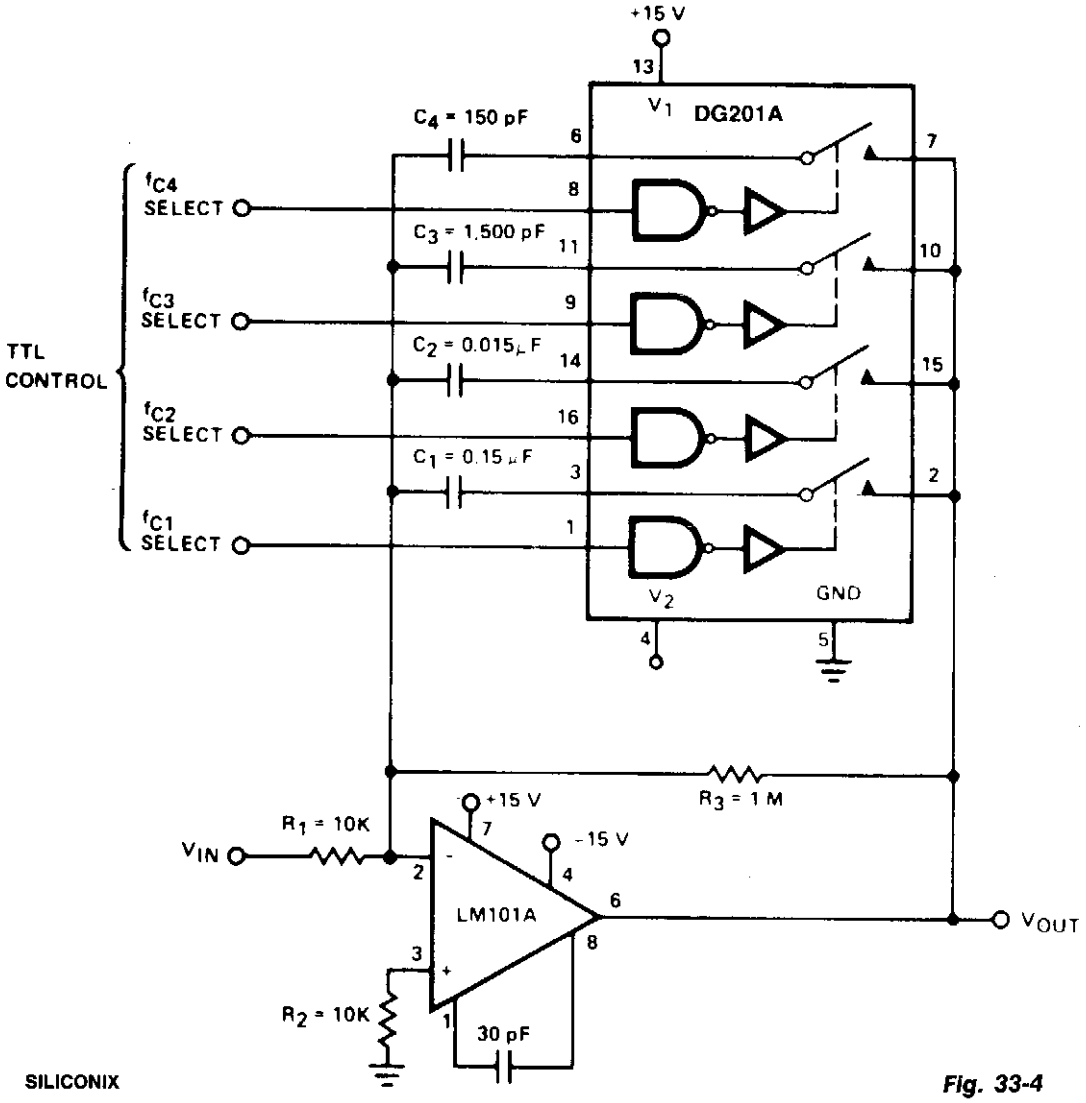
Fig. 33-3

NATIONAL SEMICONDUCTOR CORP.

Circuit Notes

The circuit provides a 10MHz cutoff frequency. Resistor R3 ensures that the input capacitance of the amplifier does not interact with the filter response at the frequency of interest. An equivalent low pass filter is similarly obtained by capacitance and resistance transformation.

ACTIVE LOW-PASS FILTER WITH DIGITALLY SELECTED BREAK FREQUENCY



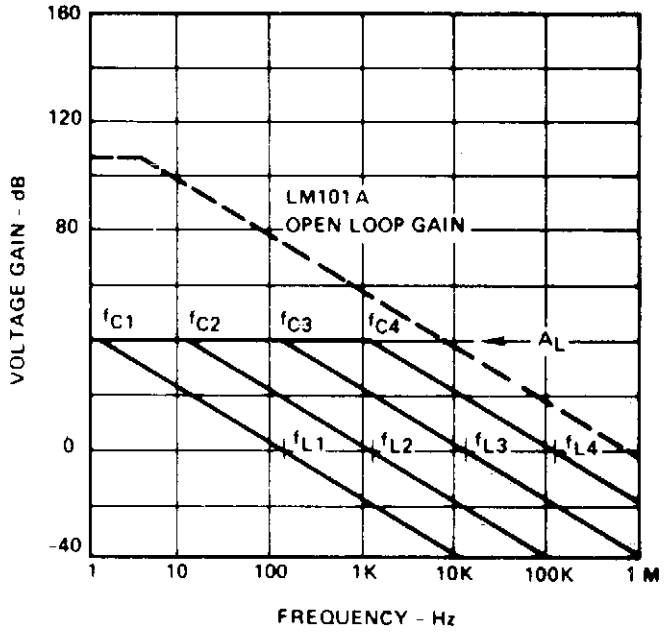
SILICONIX

Fig. 33-4

Circuit Notes

Variable low-pass filter has break frequencies at 1, 10, 100 Hz and 1 kHz. The break frequency is

$$1.\quad fc = \frac{1}{2\,\pi\,R_3\,C_X}$$



A_L (VOLTAGE GAIN BELOW BREAK FREQUENCY)
$= \frac{R_3}{R_1} = 100\ (40\ dB)$

f_c (BREAK FREQUENCY) $= \frac{1}{2\pi R_3 C_X}$

f_L (UNITY GAIN FREQUENCY) $= \frac{1}{2\pi R_1 C_X}$

MAX ATTENUATION $= \frac{r_{DS(on)}}{10K} = -40\ dB$

The low frequency gain is

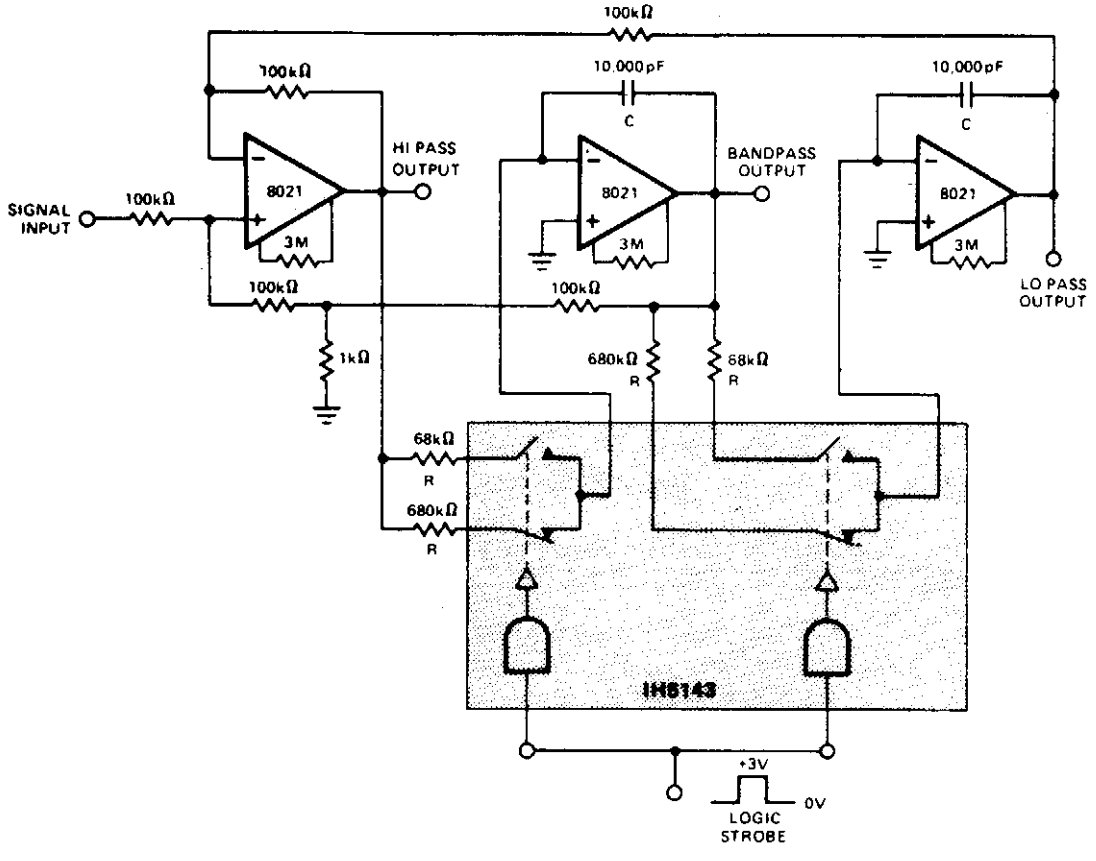
$$2. \quad A_L = \frac{R_3}{R_3} = 100\ (40\ dB)$$

A second break frequency (a zero) is introduced by $r_{DS(on)}$ of the DG201A, causing the minimum gain to be

$$3. \quad A_{MIN} = \frac{r_{DS(on)}}{R_1} \approx \frac{100}{10K} = .01,$$

a maximum attenuation of 40 dB (80 dB relative to the low frequency gain).

DIGITALLY TUNED LOW POWER ACTIVE FILTER



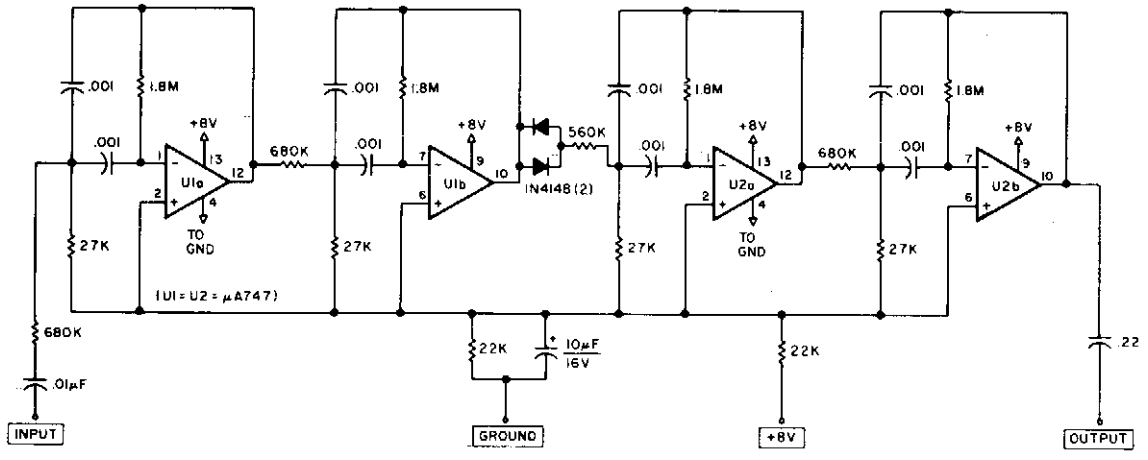
INTERSIL

Fig. 33-5

Circuit Notes

This constant gain, constant Q, variable frequency filter provides simultaneous low-pass, bandpass, and high-pass outputs with the component values shown, the center frequency will be 235 Hz and 23.5 Hz for high and low logic inputs respectively, Q = 100, and gain = 100.

RAZOR-SHARP CW FILTER



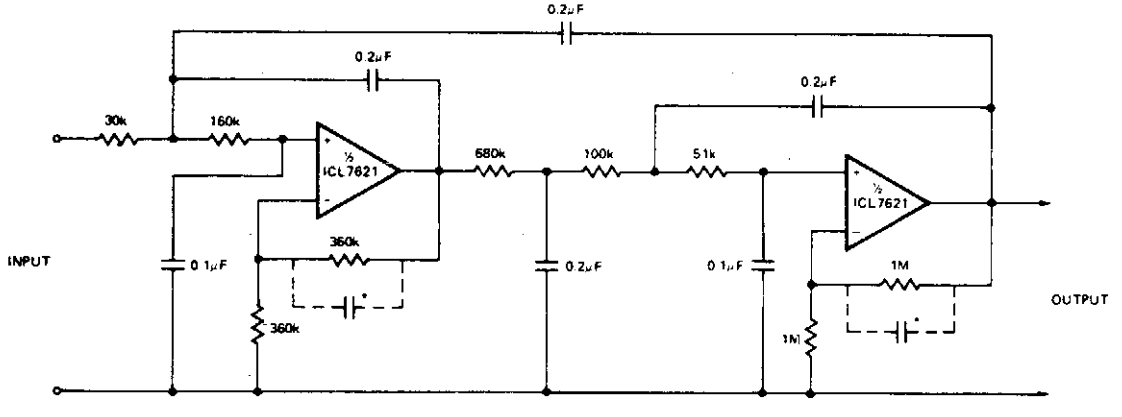
73 MAGAZINE

Fig. 33-6

Circuit Notes

The circuit consists of four stages of active bandpass filtering provided by two type-μA747 integrated-circuit dual op amps and includes a simple threshold detector (diodes D1 and D2) between stages 2 and 3 to reduce low-level background noise. Each of the four filter stages acts as a narrow bandpass filter with an audio bandpass centered at 750 Hz. The actual measured 3-dB bandwidth is only 80 Hz wide.

FIFTH ORDER CHEBYSHEV MULTIPLE FEEDBACK LOW PASS FILTER



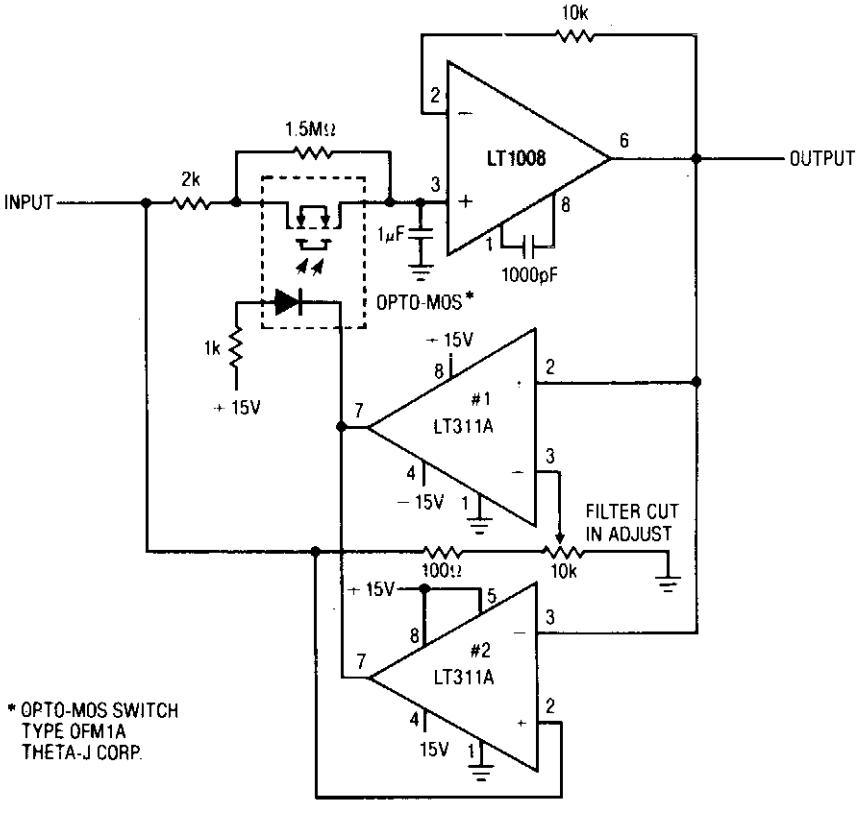
INTERSIL

Fig. 33-7

Circuit Notes

The low bias currents permit high resistance and low capacitance values to be used to achieve low frequency cutoff. f_c = 10 Hz, A_{VCL} = 4, Passband ripple = 0.1 dB. Note that small capacitors (25-50 pF) may be needed for stability in some cases.

PRECISION, FAST SETTLING, LOW-PASS FILTER



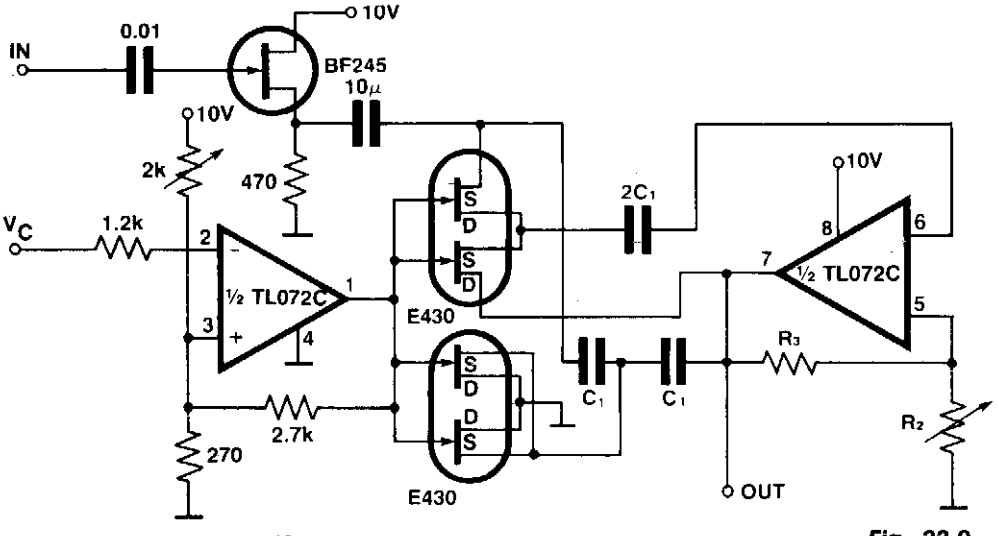
LINEAR TECHNOLOGY

Fig. 33-8

Circuit Notes

This circuit is useful where fast signal acquisition and high precision are required, as in electronic scales. The filter's time constant is set by the 2 K ohm resistor and the 1 μF capacitor until comparator No. 1 switches. The time constant is then set by the 1.5 M ohm resistor and the 1 μF capacitor. Comparator No. 2 provides a quick reset. The circuit settles to a final value three times as fast as a simple 1.5 M ohm—1 μF filter, with almost no dc error.

PROGRAMMABLE BANDPASS USING TWIN-T BRIDGE



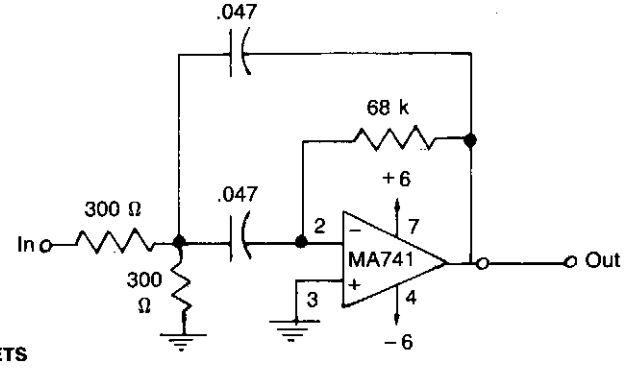
ELECTRONIC ENGINEERING

Fig. 33-9

Circuit Notes

The circuit gives a programmable bandpass where both the cut-over frequency and the gain, A, are controlled independently. In the twin-T bridge the resistors R and R/2 are replaced by two double FETs, E 430, the channel resistance of the first one in the series, the channel resistances of the second one are in parallel as to stimulate the resistance R/2. Both these resistors are controlled by V_c which ranges from 0 V to about 1 V. The gain of the circuit is set by means of the resistors R2 and R3.

ACTIVE BANDPASS FILTER (f0 = 1000 Hz)



WILLIAM SHEETS

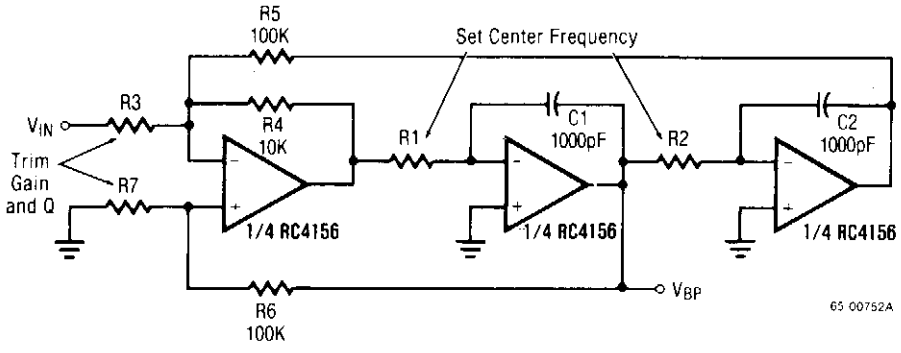
Fig. 33-10

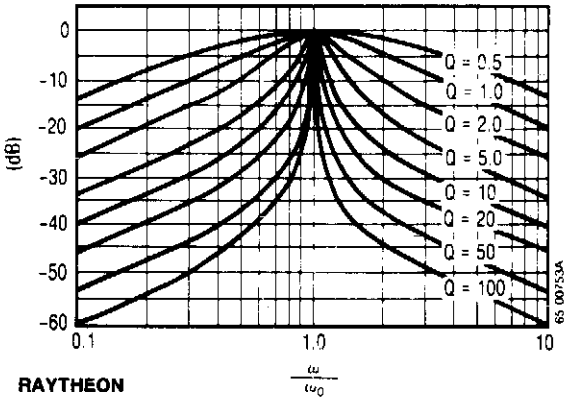
Circuit Notes

This filter has a bandpass centered around 1kHz, for applications such as bridge amplifiers, null detectors, etc.

The circuit uses a μA741 IC and standard 5% tolerance components.

BANDPASS FILTER





$$\frac{V_{BP}}{V_{IN}} = \frac{\frac{\omega}{\omega_0}\ \frac{1}{Q}}{\sqrt{\left[1-\left(\frac{\omega}{\omega_0}\right)^2\right]^2+\left(\frac{1}{Q}\ \frac{\omega}{\omega_0}\right)^2}}$$

RAYTHEON

Fig. 33-11

Circuit Notes

The input signal is applied through R3 to the inverting input of the summing amplifier and the output is taken from the first integrator. The summing amplifier will maintain equal voltage at the inverting and non-inverting inputs. Defining 1/R1C1 as ω_1 and 1/R2C2 as ω_2, this is now a convenient form to look at the center-frequency ω_0 and filter Q.

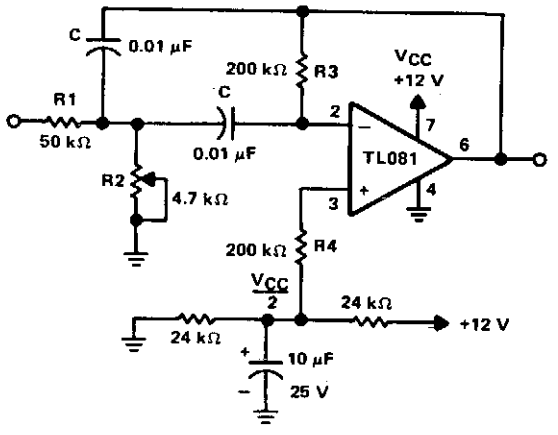
$$\omega_0 = \sqrt{0.1\ \omega_1\ \omega_2}$$

$$= 10^{-9}\sqrt{0.1R1R2}$$

$$\text{and } Q = \left[\frac{1+\frac{10^5}{R7}}{1.1+\frac{10^4}{R3}}\right]\omega_0$$

The frequency response for various values of Q is shown.

ACTIVE BANDPASS FILTER



TEXAS INSTRUMENTS

Fig. 33-12

Circuit Notes

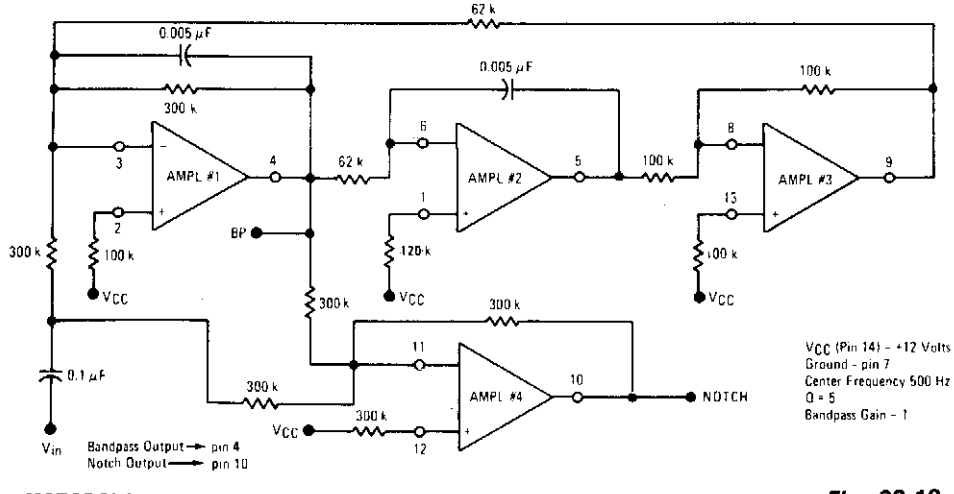
The circuit is a two-pole active filter using a TL081 op amp. This type of circuit is usable only for Qs less than 10. The component values for this filter are calculated from the following equations.

$$R1 = \frac{Q}{2\ fGC} \qquad R3 = \frac{2Q}{2fC}$$

$$R2 = \frac{Q}{(2Q^2 - G)2fC} \qquad R4 = R3$$

The values shown are for a center frequency of 800 Hz.

BANDPASS AND NOTCH FILTER



MOTOROLA

Fig. 33-13

Circuit Notes

The Quad op amp MC4301 is used to configure a filter that will notch out a given frequency and produce that notched-out frequency at the BP terminal, useful in communications or measurement setups. By proper component selection any frequency filter up to a few tens of kilohertz can be obtained.

MULTIPLE-FEEDBACK BANDPASS FILTER

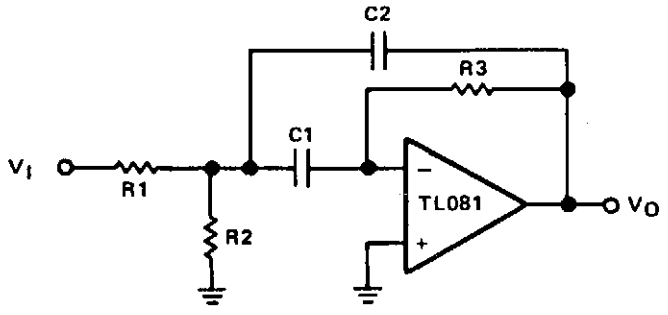


Fig. 33-14

TEXAS INSTRUMENTS

Circuit Notes

The op amp is connected in the inverting mode. Resistor R3 from the output to the inverting input sets the gain and current through the frequency-determining capacitor, C1. Capacitor C2 provides feedback from the output to the junction of R1 and R2. C1 and C2 are always equal in value. Resistor R2 may be made adjustable in order to adjust the center frequency which is determined from:

$$fo = \frac{1}{2\pi C} \quad \frac{1}{R3} \times \frac{R1 + R2}{R1R2}$$

When designing a filter of this type it is best to select a value for C1 and C2, keeping them equal. Typical audio filters have capacitor values from 0.01 μF to 0.1 μF which will result in reasonable values for the resistors.

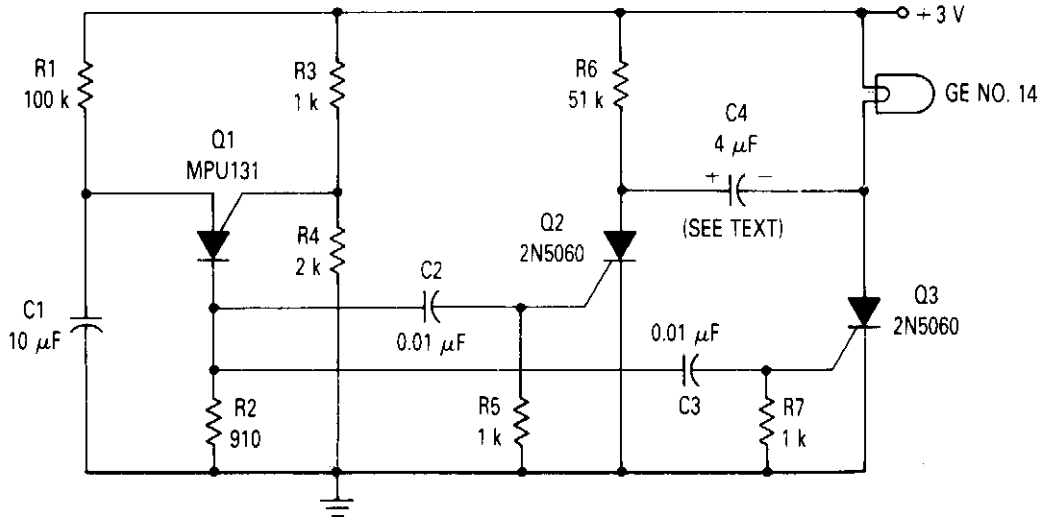
34

Flashers and Blinkers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low Voltage Lamp Flasher
Miniature Transistorized Light Flasher
Alternating Flasher
Electronic Light Flasher
High-Power Battery-Operated Flasher
Series SCR Lamp Flasher Handles a Wide Range of Loads
SCR Relaxation Flasher
Low Current Consumption Lamp Flasher
LED Flasher Uses PUT
2 kW Flasher with Photoelectric Control
Sequential Flasher
1 kW Flip-Flop Flasher Circuit
Low Frequency Oscillator Flasher
High Drive Oscillator/Flasher
Transistorized Flashers
Sequential ac Flasher
Astable Flip-Flop with Starter
LED Flasher Uses UJT

LOW VOLTAGE LAMP FLASHER



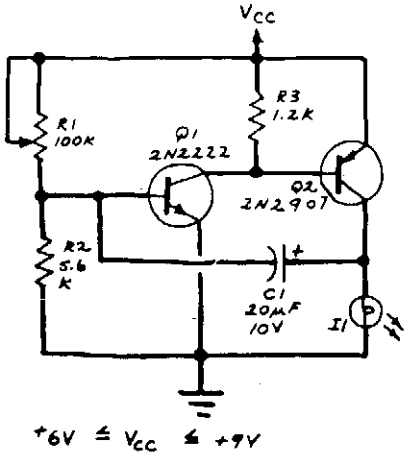
MOTOROLA

Fig. 34-1

Circuit Notes

The circuit is composed of a relaxation oscillator formed by Q1 and an SCR flip-flop formed by Q2 and Q3. With the supply voltage applied to the circuit, the timing capacitor C1 charges to the firing point of the PUT, 2 volts plus a diode drop. The output of the PUT is coupled through two 0.02 μF capacitors to the gate of Q2 and Q3. To clarify operation, assume that Q3 is on and capacitor C4 is charged plus to minus as shown in the figure. The next pulse from the PUT oscillator turns Q2 on. This places the voltage on C4 across Q3 which momentarily reverse biases Q3. This reverse voltage turns Q3 off. After discharging, C4 then charges with its polarity reversed to that shown. The next pulse from Q1 turns Q3 on and Q2 off. Note that C4 is a non-polarized capacitor. For the component values shown, the lamp is on for about ½ second and off the same amount of time.

MINIATURE TRANSISTORIZED LIGHT FLASHER



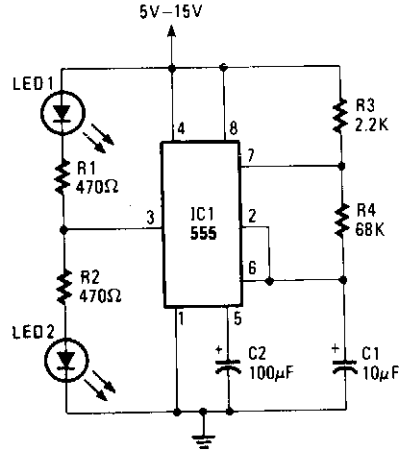
Circuit Notes

R1 adjusts the flash rate. The lamp should be a No. 122, No. 222 or other similar, miniature incandescent lamp.

POPULAR ELECTRONICS

Fig. 34-2

ALTERNATING FLASHER



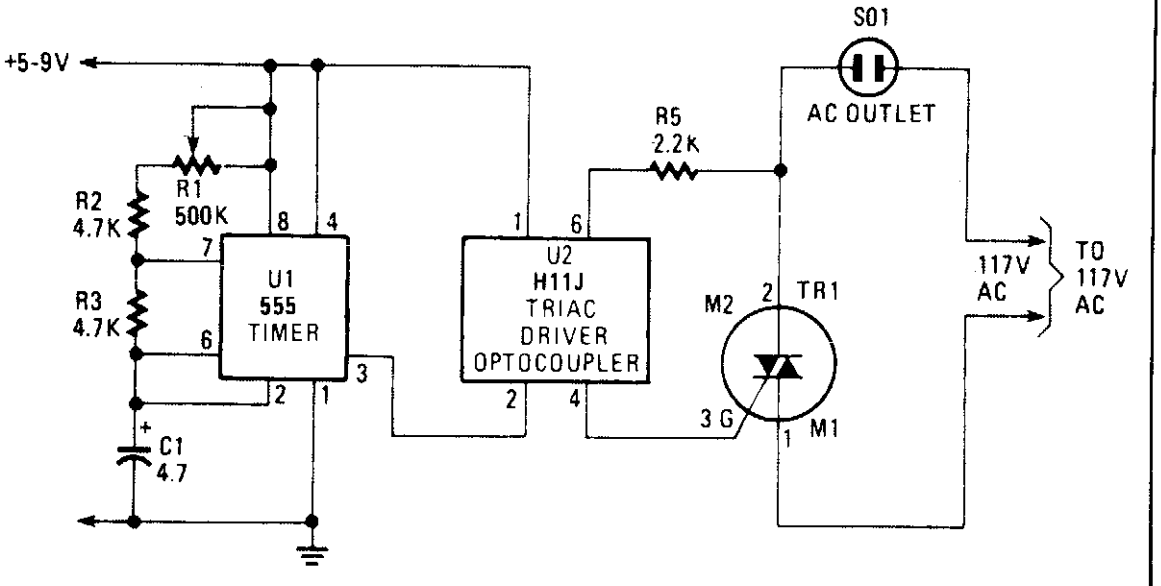
Circuit Notes

The LED's flash alternately. The flash rate is determined by C1 and R4.

HANDS-ON ELECTRONICS

Fig. 34-3

ELECTRONIC LIGHT FLASHER



HANDS-ON ELECTRONICS

Fig. 34-4

Circuit Notes

The blinking or flashing rate is determined by U1, a 555 timer integrated circuit. Its output, at pin 3, feeds U2, a HIIJ triac driver. That driver consists of an infrared LED that is coupled internally to a light-activated silicon bilateral switch (DIAC). When the LED internal to U2 is turned on by the timer, U1, its light triggers the DIAC; effectively closing the circuit between pins 4 and 6, and fires the Triac, TR1 through its gate circuit. When the Triac is firing, it acts as a closed circuit that turns on the light (or other device it may be controlling via S01). When the timer turns off, the LED, the DIAC and Triac stop conducting and the light turns off. The sequence then repeats. The flashing rate can be varied by means of R1, a 500,000 ohm potentiometer.

HIGH-POWER BATTERY-OPERATED FLASHER

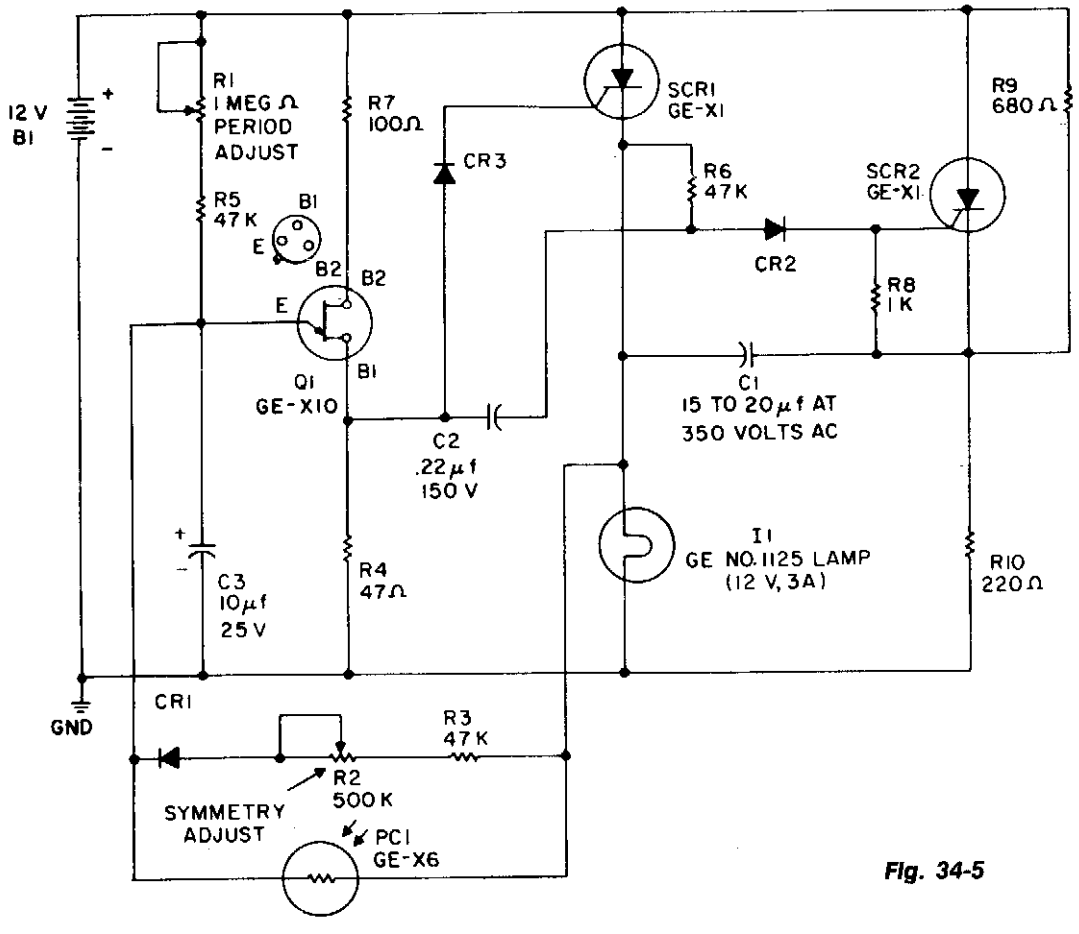


Fig. 34-5

GENERAL ELECTRIC

Circuit Notes

This flasher operates from a 12-volt car or boat battery. It offers 36 to 40-watts output, variable flash rate (up to 60 flashes per minute), independent control of both on and off cycles and photoelectric night and day control that turns the flasher on at night and shuts it off during the day for automatic operation. SCR1 and SCR2 form a basic dc flip-flop. The lamp load is the cathode leg of one SCR so that the other side of the load may be at ground (negative) potential (required in some applications). The flip-flop timing is controlled by a conventional UJT oscillator arrangement (Q1, R1, C3, etc.). Potentiometer R2 and diode CR1 provide on/off timing independence. Photoconductor PC1 locks out the UJT firing circuit during the daylight hours.

SERIES SCR LAMP FLASHER HANDLES A WIDE RANGE OF LOADS

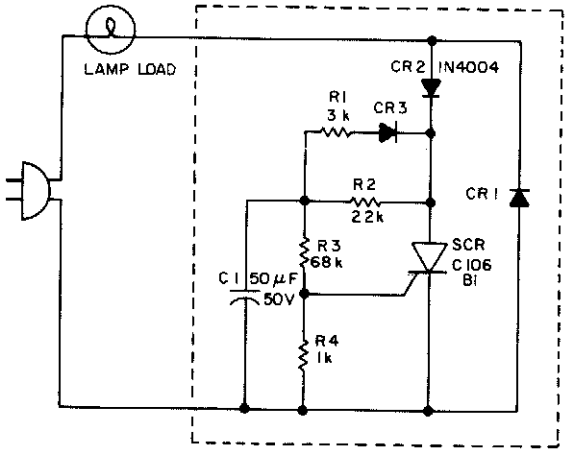


Fig. 34-6

ELECTRONIC DESIGN

Circuit Notes

Brief full-power flashes are obtained when the SCR conducts during positive half cycles of the line voltage. The SCR fires when the voltage at the divider, R3 and R4, reaches the gate-firing level. Diode D1 conducts during the reverse cycle of the SCR and provides preheating current to the lamp filaments.

SCR RELAXATION FLASHER

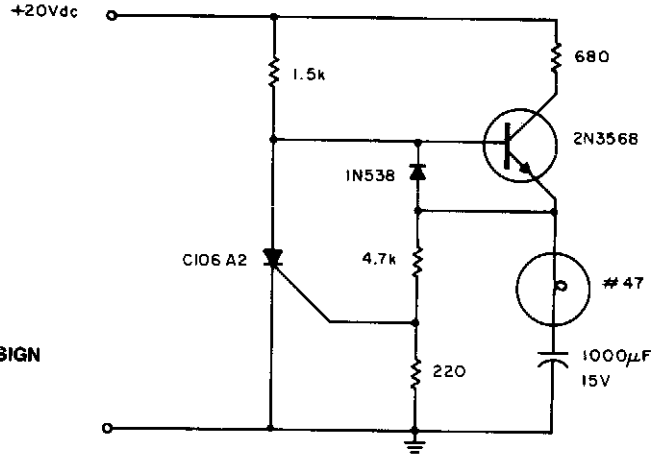


Fig. 34-7

ELECTRONIC DESIGN

Circuit Notes

Flashing occurs each time the capacitor discharges through the turned-on SCR. When the discharge current falls below the SCR holding current, the SCR turns off, and the capacitor begins charging for another cycle. The circuit will maintain a slower but good flashing capability even after considerable battery degradation.

LOW CURRENT CONSUMPTION LAMP FLASHER

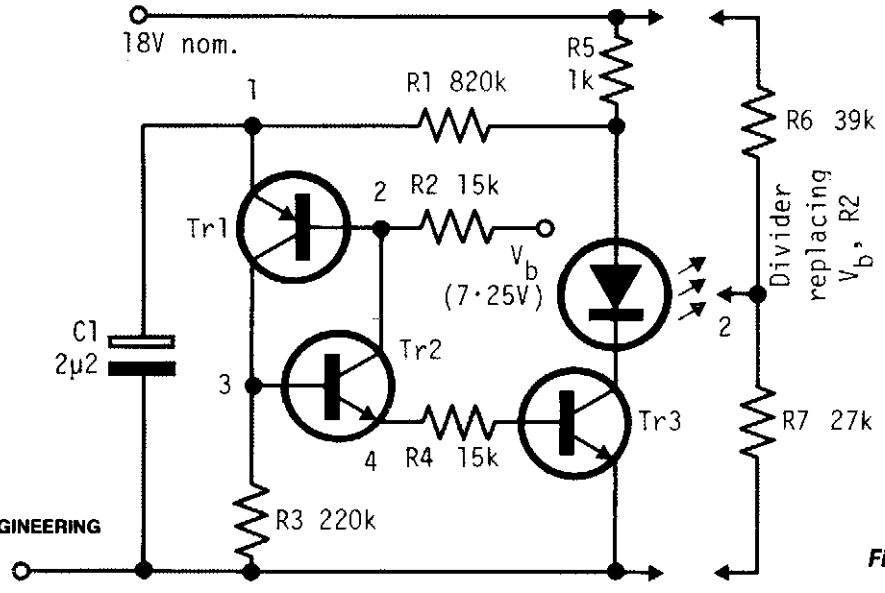


Fig. 34-8

Circuit Notes

The circuit is economical in components, and will work with virtually any transistors and is reliably self-starting. The voltage Vb can be taken from a divider, as shown at the right. If taken from a fixed source, flashing becomes slower as battery voltage falls. The lowest drive current into the base of Tr3 is about (Vb—0,6 V)/(R2 + R4). Resistor R4 limits the initial current from C1 and, as shown, R2 and R4 can be roughly equal when a divider is used for Vb. Resistor R2 equals R6R7/(R6 + R7). With the voltages shown, and with R2 = R4, the on-time is about 1.1 C1R2 and the off-time about 0.28 C1R1. Using the component values shown the period is about 0.55 sec. with a duty cycle of about 7% and a mean battery current including the Vb divider, about 1.5 mA.

LED FLASHER USES UJT

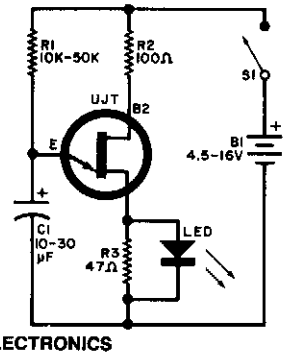
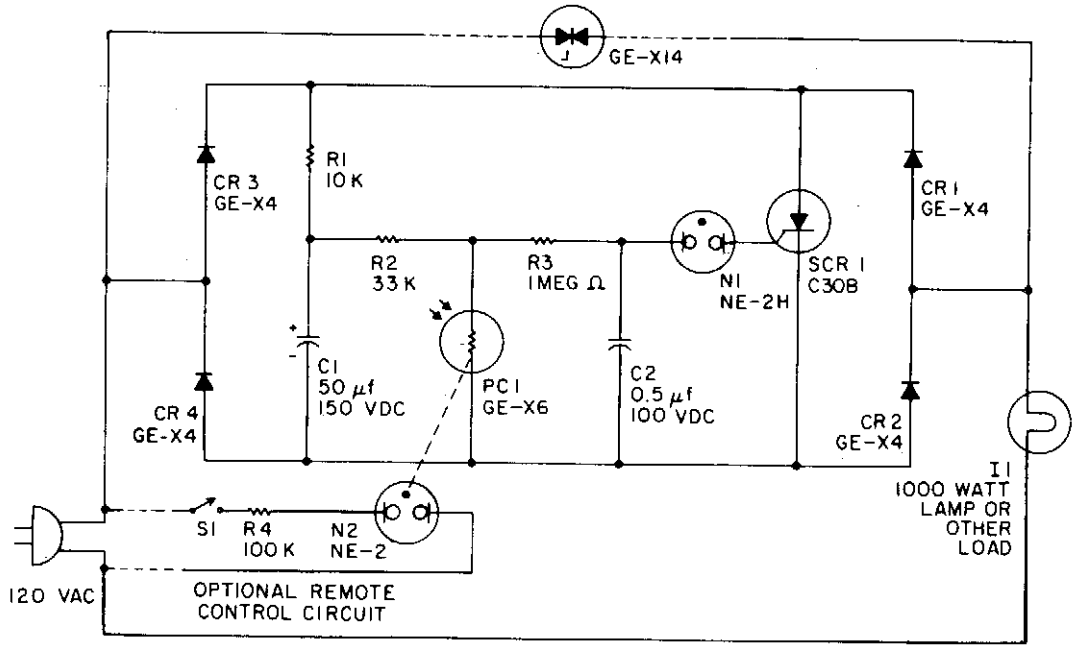


Fig. 34-9

Circuit Notes

A relaxation oscillator is used to flash an LED in the base circuit. C1 is charged slowly through R1 by the power source, then discharged periodically through R3 and the LED by the UJT. Flashing rate is determined by the supply voltage and by R1-C1's time constant. UJT = 2N4871

2 kW FLASHER WITH PHOTOELECTRIC CONTROL



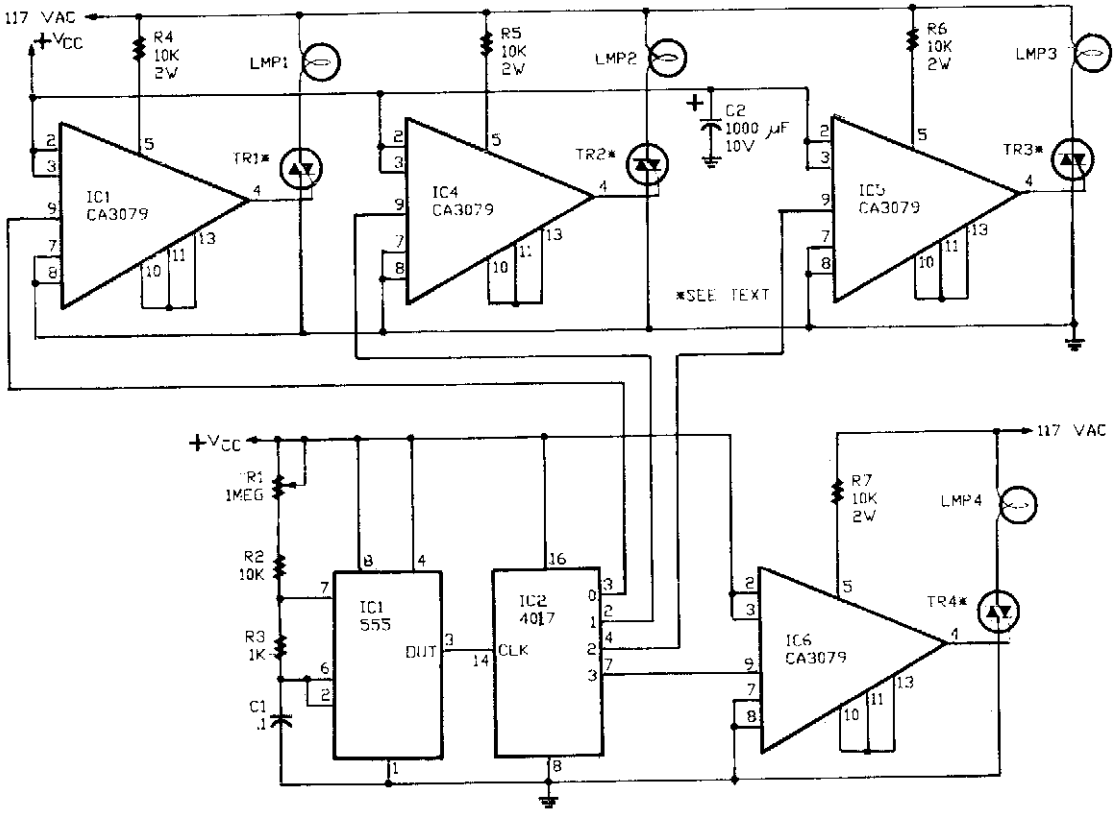
GENERAL ELECTRIC

Fig. 34-10

Circuit Notes

CR1, CR2, CR3, and CR4 form a bridge circuit with the SCR across the dc legs. With light on the photoconductor PC1, C1 charges through R1 to about 150 Vdc. The resistance of PC1 is low when illuminated, so very little voltage appears across it or C2. At about 90 volts C1 starts discharging through R1 and the SCR, but the SCR cannot turn off until C1 is almost completely discharged. When the SCR turns off during the interval line voltage is near zero, the full supply voltage again appears across the bridge, and C1 charges again to a high voltage. The voltage on C2 also starts rising until the neon lamp fires and the cycle repeats. An alternative remote control can be made by adding a second neon lamp, N2, and masking the photocell so it sees only N2. A very sensitive remote control is thus obtained that is completely isolated from the load circuit. For low-voltage remote control a flashlight lamp may be used instead of N2 and operated at about ½ its normal voltage thus giving exceptionally long life. Performance of the photoelectric control may be inverted (flash when the photoconductor is illuminated) by interchanging PC1, and R2. Sensitivity in either the normal or inverted modes can be decreased by partially masking PC1, and can be increased by increasing resistor R2 to about 470 K. To increase on time, increase C1; to increase off time, increase R3.

SEQUENTIAL FLASHER



RADIO-ELECTRONICS

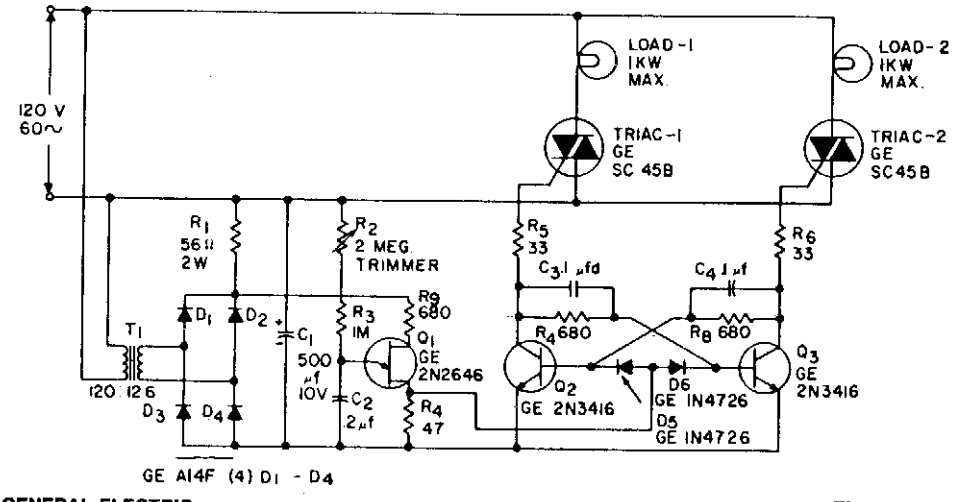
Fig. 34-11

Circuit Notes

A 555 timer, IC1, drives a 4017 CMOS decade counter. Each of the 4017's first four outputs drives a CA3079 zero-voltage switch. Pin 9 of the CA3079 is used to inhibit output from pin 4, thereby disabling the string of pulses that IC normally delivers. Those pulses occur every 8.3 ms, i.e., at a rate of 120 Hz. Each pulse has a width of 120 µs. Due to the action of the CA3079, the lamps connected to the TRI-AC's turn on and off near the zero crossing of the ac waveform. Switching at that point increases lamp life by reducing the inrush of current that would happen if the lamp were turned on near the high point of the ac waveform. In addition, switching at the zero crossing reduces Radio-Frequency Interference (RFI) considerably.

CAUTION: The CA3079's are driven directly from the 117-volt ac power line, so use care.

1 kW FLIP-FLOP FLASHER CIRCUIT



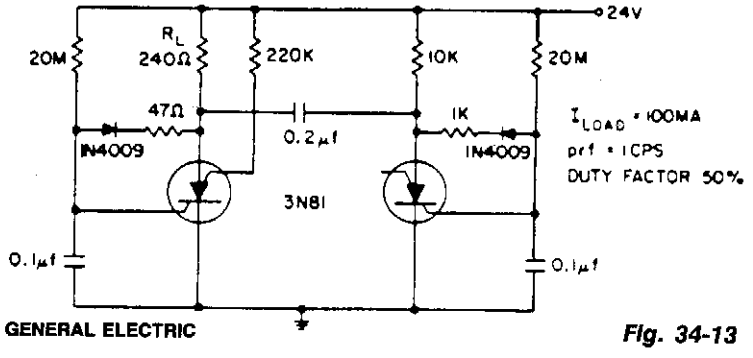
GENERAL ELECTRIC

Fig. 34-12

Circuit Notes

This is an application of the static switch circuit where the control logic is a flip-flop which is controlled by the unijunction transistor. The flashing rate can be adjusted from about 0.1 second to a 10 second cycle time.

LOW FREQUENCY OSCILLATOR-FLASHER



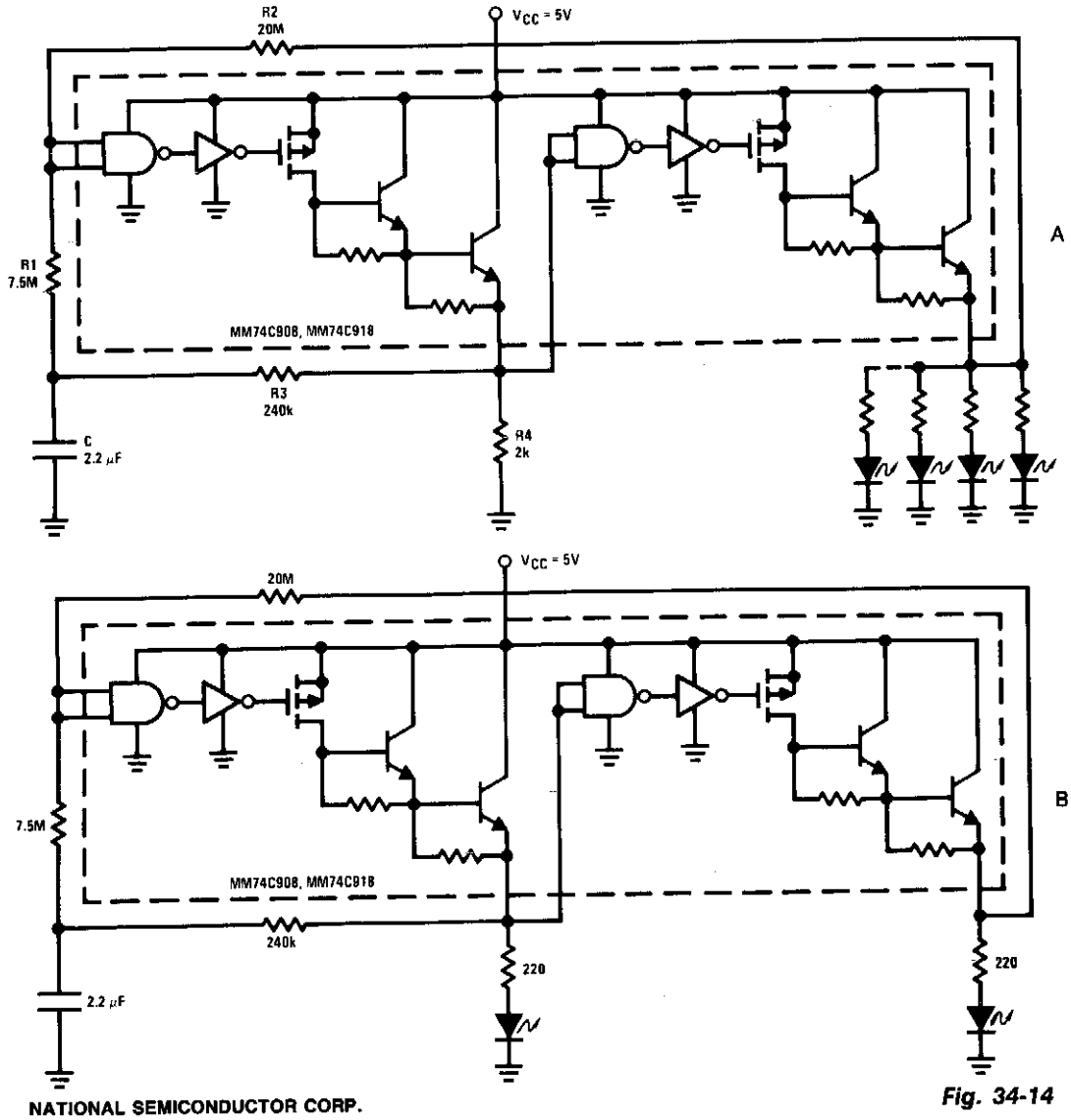
GENERAL ELECTRIC

Fig. 34-13

Circuit Notes

Electrolytic capacitors are unnecessary to generate a 1 cps frequency. As an scs triggers on, the 0.2 μF commutating capacitor turns off the other one and charges its gate capacitor to a negative potential. The gate capacitor charges towards 24 volts through 20 M retriggering its scs. Battery power is delivered to the load with 88% efficiency. The 20 M resistors can be varied to change prf or duty factor.

HIGH DRIVE OSCILLATOR/FLASHER



NATIONAL SEMICONDUCTOR CORP.

Fig. 34-14

Circuit Notes

The driver in the package is connected as a Schmitt trigger oscillator (A) where R1 and R2 are used to generate hysteresis. R3 and C are the inverting feedback timing elements and R4 is the pull-down load for the first driver. Because of its current capability, the circuit can be used to drive an array of LEDs or lamps. If resistor R4 is replaced by an LED (plus a current limiting resistor), the circuit becomes a double flasher with the 2 LEDs flashing out of phase (B).

TRANSISTORIZED FLASHERS

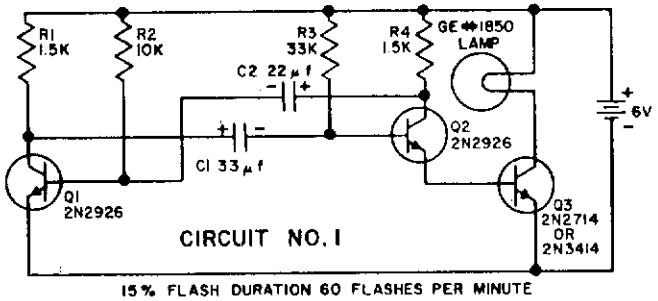
Flasher circuit performance as a function of temperature using limit sample transistors.

			AMBIENT TEMPERATURE -40°F	77°F	212°F
CIRUIT #1	Flashes per Minute	min.	58.0	60.0	59.6
		max.	56.7	59.6	58.7
	Flash Duration in %	min.	14.5	14.6	14.3
		max.	15.1	14.9	15.3
CIRCUIT #2	Flashes per Minute	min.	58.6	60.0	59.0
		max.	55.6	58.1	56.4
	Flash Duration in %	min.	26.3	27.5	22.2
		max.	28.8	29.1	29.6
CIRCUIT #3	Flashes per Minute	min.	59.0	60.0	61.5
		max.	55.4	57.7	55.2
	Flash Duration in %	min.	45.2	46.0	48.2
		max.	45.6	46.2	47.0

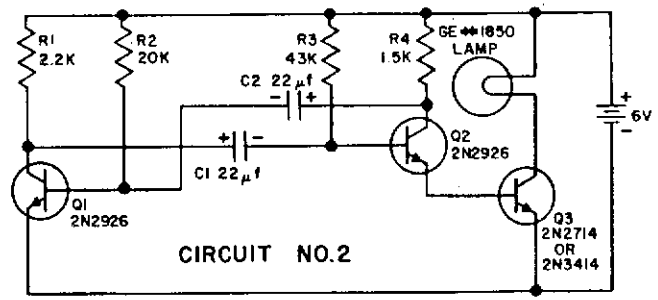
Circuit Notes

Transistors Q1 and Q2 are connected as a free running multivibrator. The output, at the emitter of Q2, drives the base of the common emitter amplifier Q3, which controls the lamp. This circuit configuration permits the flash duration, the interval between flashes, and the lamp type to be varied independently. Flash duration is proportional to the product of R2C2 while the off interval is proportional to the product of R3C1. Consequently, when the flash timing must be accurately maintained, these component tolerances will have to be held to similar limits.

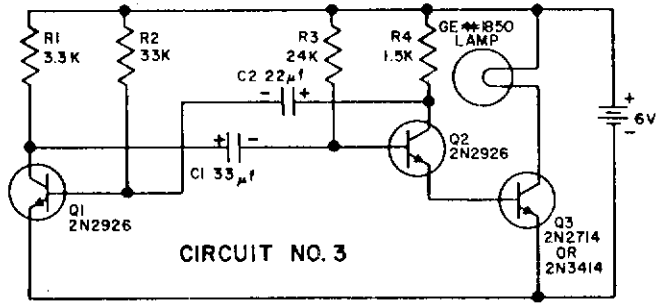
All three circuits described are designed for barricade warning flasher lights such as used in highway construction. They differ only in flash duration which normally is 15%, 25%, or 50% of the flash rate. Performance has been checked at ambient temperatures of −40°F, 77°F, and 212°F. A GE 5 volt, 90 milliampere type No. 1850 lamp is used.



15% FLASH DURATION 60 FLASHES PER MINUTE



25% FLASH DURATION 60 FLASHES PER MINUTE



50% FLASH DURATION 60 FLASHES PER MINUTE

Fig. 34-15

GENERAL ELECTRIC

SEQUENTIAL AC FLASHER

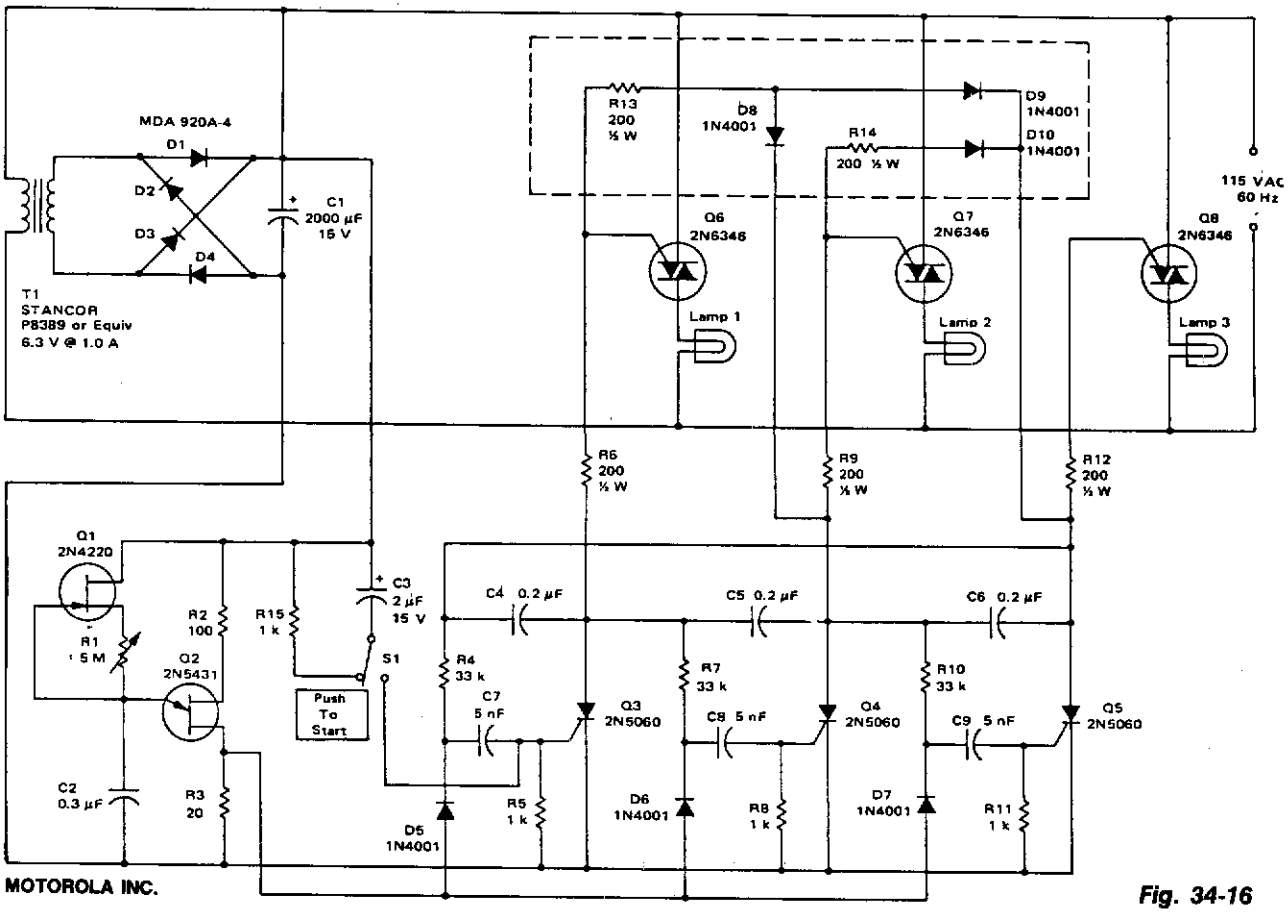
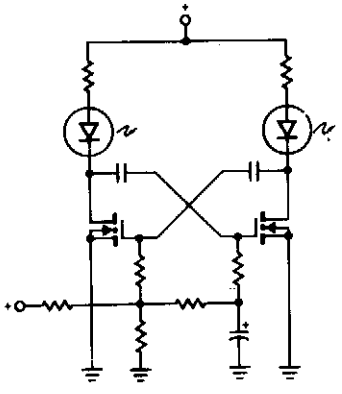


Fig. 34-16

Circuit Notes

This circuit uses a ring counter made up of Q3, Q4 and Q5. Q2 acts as an oscillator which triggers the SCRs Q3, Q4 and Q5 in sequence. When an SCR fires, a 0.2µF capacitor provides commutation, shutting off the previous stage. Triacs Q6, Q7 and Q8 control lamps 1, 2 and 3 respectively.

ASTABLE FLIP-FLOP WITH STARTER

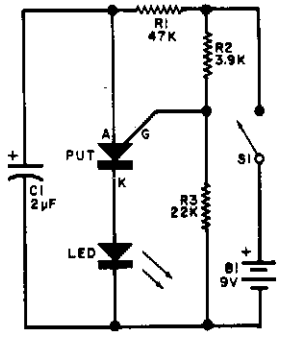


SILICONIX, INC. *Fig. 34-17*

Circuit Notes

A pair of non-zenered MOSPOWER transistors, a pair of LEDs and a simple RC circuit make an easy sequential flasher with almost unlimited sequencing time—from momentary to several seconds. The infinite input resistance of the MOSFET gate allows for very long sequencing times that are impossible when using bipolars. One precaution, though, don't wire your circuit using phenolic or printed circuit boards when looking for slow sequencing (they exhibit too much leakage!).

LED FLASHER USES PUT



POPULAR ELECTRONICS *Fig. 34-18*

Circuit Notes

This flasher circuit operates as a relaxation oscillator with C1 discharged periodically through the LED as the PUT switches on. The flashing rate is about 100/minute with the component values listed.

35

Flow Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Thermally Based Anemometer (Air Flowmeter)
Air Flow Detector

THERMALLY BASED ANEMOMETER (AIR FLOWMETER)

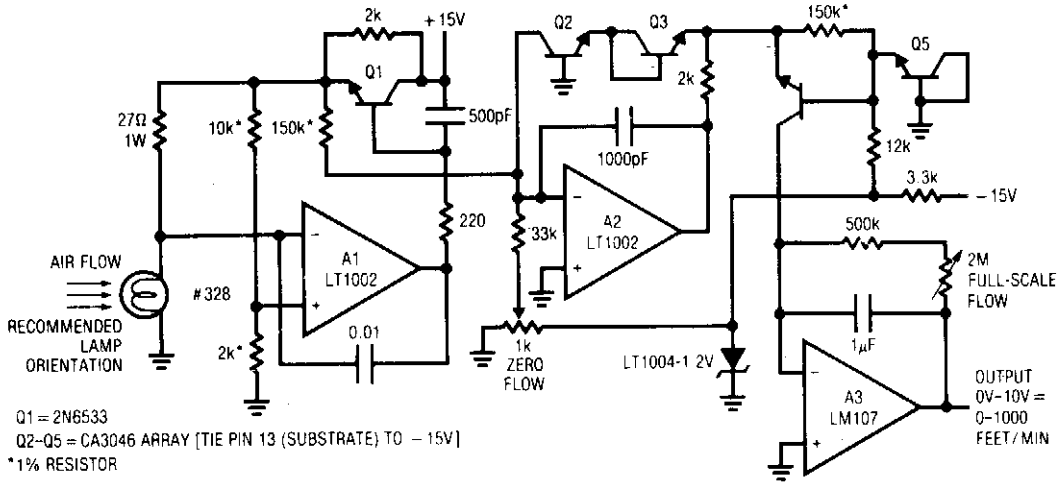


Fig. 35-1

LINEAR TECHNOLOGY CORPORATION

Circuit Notes

This design used to measure air or gas flow works by measuring the energy required to maintain a heated resistance wire at constant temperature. The positive temperature coefficient of a small lamp, in combination with its ready availability, makes it a good sensor. A type 328 lamp is modified for this circuit by removing its glass envelope. The lamp is placed in a bridge which is monitored by A1. A1's output is current amplified by Q1 and fed back to drive the bridge. When power is applied, the lamp is at a low resistance and Q1's emitter tries to come full on. As current flows through the lamp, its temperature quickly rises, forcing its resistance to increase. This action increases A1's negative input potential. Q1's emitter voltage decreases and the circuit finds a stable operating point. To keep the bridge balanced, A1 acts to force the lamp's resistance, hence its temperature, constant. The 20 k - 2 k bridge values have been chosen so that the lamp operates just below the incandescence point.

To use this circuit, place the lamp in the air flow so that its filament is at a 90° angle to the flow. Next, either shut off the air flow or shield the lamp from it and adjust the zero flow potentiometer for a circuit output of 0 V. Then, expose the lamp to air flow of 1000 feet/minute and trim the full flow potentiometer for 10 V output. Repeat these adjustments until both points are fixed. With this procedure completed, the air flowmeter is accurate within 3% over the entire 0-1000 foot/minute range.

AIR FLOW DETECTOR

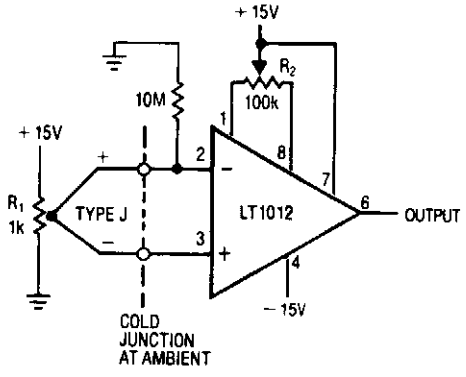


Fig. 35-2

Mount R_1 in airflow.
Adjust R_2 so output goes high when airflow stops.

LINEAR TECHNOLOGY CORP.

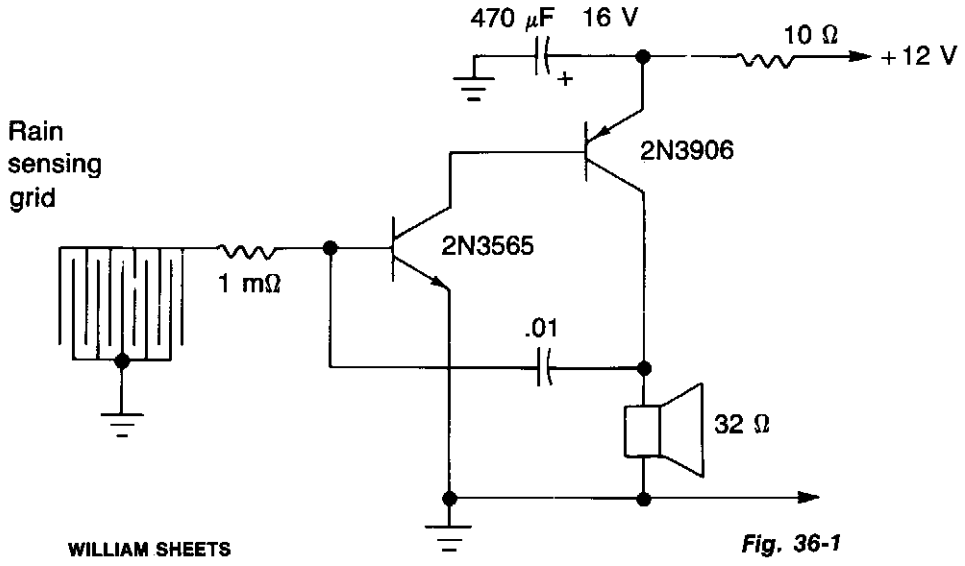
36

Fluid and Moisture Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Rain Warning Bleeper
Water-Level Indicator
Acid Rain Monitor
Plant-Water Monitor
Water-Level Sensing and Control
Single Chip Pump Controller
Plant-Water Gauge
Liquid Flowmeter

RAIN WARNING BLEEPER



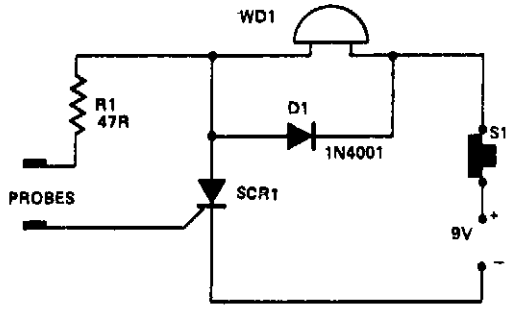
WILLIAM SHEETS

Fig. 36-1

Circuit Notes

One small spot of rain on the sense pad of this bleeper will start this audio warning. It can also be operated by rising water. The circuit has two transistors, with feedback via capacitor C1, but Trl cannot operate as long as the moisture sense pad is dry. When the pad conducts, Tr1 and Tr2 form an audio oscillatory circuit, the pitch depends somewhat on the resistance.

WATER-LEVEL INDICATOR



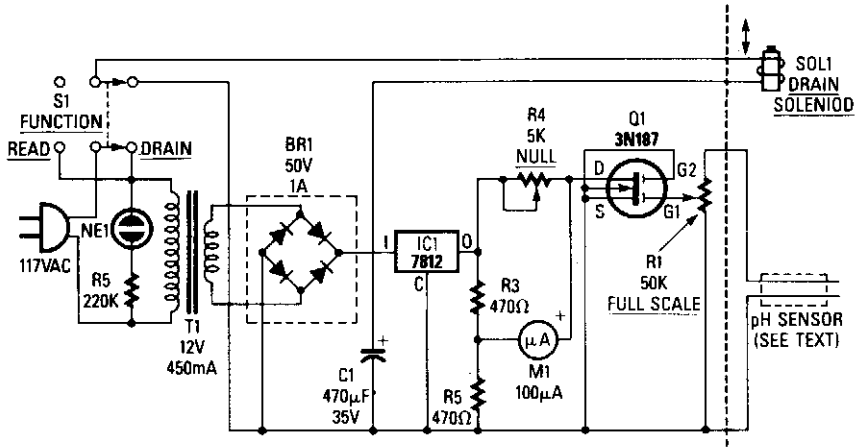
ELECTRONICS TODAY INTERNATIONAL

Fig. 36-2

Circuit Notes

In this a warning device WD1 is in series with SCR1. When the liquid level causes a conductive path between the probes, the SCR conducts sounding WD1. The warning device may be a Sonalert (TM), a lamp or a buzzer. D1 acts as a transient suppressor. Press S1 to reset the circuit.

ACID RAIN MONITOR



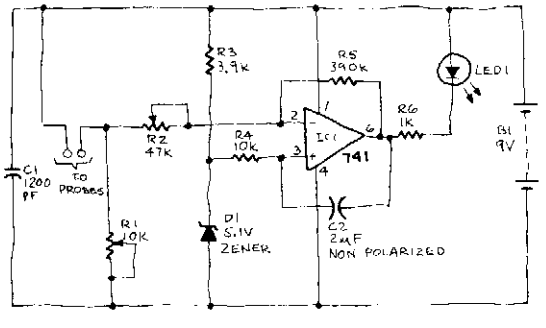
RADIO-ELECTRONICS

Fig. 36-3

Circuit Notes

A bridge rectifier and 12-volt regulator powers the MOSFET sensing circuit. The unregulated output of the bridge rectifier operates the drain solenoid via switch S1. The sensor itself is built from two electrodes, one made of copper, the other of lead. In combination with the liquid trapped by the sensor, they form a miniature lead-acid cell whose output is amplified by MOSFET Q1. The maximum output produced by our prototype cell was about 50 µA. MOSFET Q1 serves as the fourth leg of a Wheatstone bridge. When acidity causes the sensor to generate a voltage, Q1 turns on slightly, so its drain-to-source resistance decreases. That resistance variation causes an imbalance in the bridge, and that imbalance is indicated by meter M1.

PLANT-WATER MONITOR



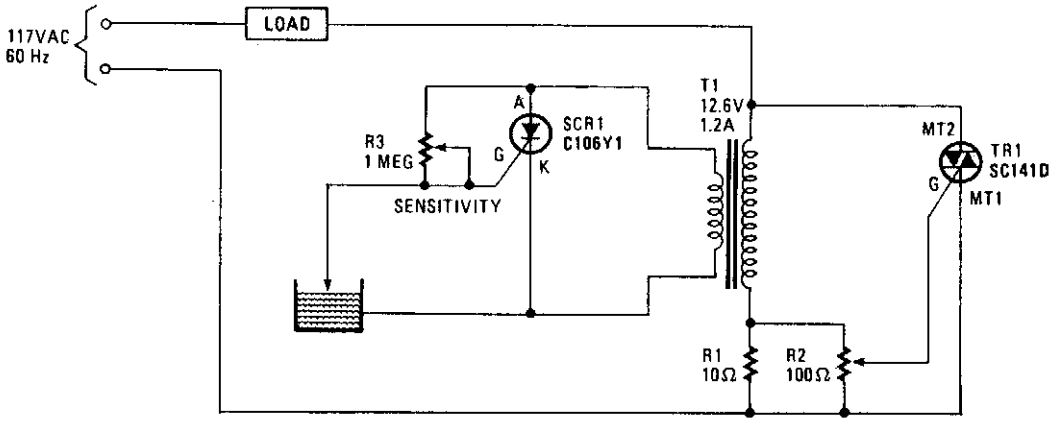
Circuit Notes

When the soil is moist, the LED glows. If the moisture falls below a certain predetermined level, the LED begins to flash. If there is still less moisture, the LED turns off. To calibrate, connect the battery and insert the probe into a container of dry soil. Set R1 to its maximum value then reduce that resistance until the LED begins to flash. The range over which the LED flashes before going out is adjusted using R2.

RADIO-ELECTRONICS

Fig. 36-4

WATER-LEVEL SENSING AND CONTROL



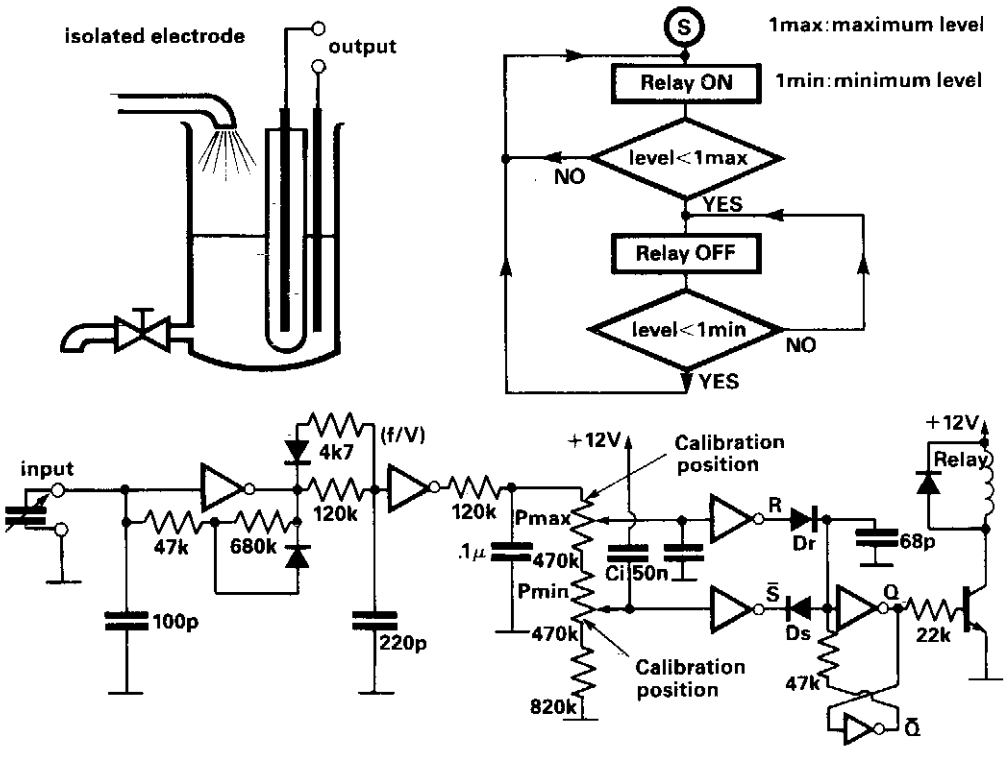
HANDS-ON ELECTRONICS

Fig. 36-5

Circuit Notes

The operation of the circuit is based on the difference in the primary impedance of a transformer when its secondary is loaded and when it is open-circuit. The impedance of the primary of T1 and resistor R1 are in series with the load. The triac's gate-control voltage is developed across parallel resistors R1 and R2. When the water level is low, the probe is out of the water and SCR1 is triggered on. It conducts and imposes a heavy load on transformer T1's secondary winding. That load is reflected back into the primary, gating triac TR1 on, which energizes the load. If the load is an electric value in the water-supply line, it will open and remain open until the water rises and touches the probe, which shorts SCR1's gate and cathode, thereby turning off the SCR1, which effectively open-circuits the secondary. The open-circuit condition—when reflected back to the primary winding—removes the triac's trigger signal, thereby turning the water off. The load may range from a water valve, a relay controlling a pump supplying water for irrigation, or a solenoid valve controlling the water level in a garden lily pond.

SINGLE CHIP PUMP CONTROLLER



ELECTRONIC ENGINEERING

Fig. 36-6

Circuit Notes

This circuit controls the level of a tank using a bang-bang controlled electrical pump. The actual level of liquid is measured by a capacitive level-meter. The first inverter performs as a capacitance to frequency converter. It is a Schmitt oscillator and its frequency output decreases as the capacitance increases. The second inverter is a monostable which performs as a frequency to voltage converter (f/V). Its output is applied to the maximum and minimum level comparator inputs. Maximum and minimum liquid levels may be set by the potentiometers. The maximum level (1 max) may be preset between the limits: 65 pF less than C (1 max) less than 120 pF. The minimum level is presetable and the limits are: 0 less than C (1 min) less than 25 pF.

PLANT WATER GAUGE

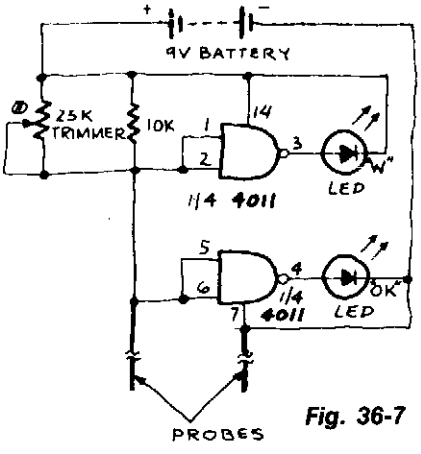
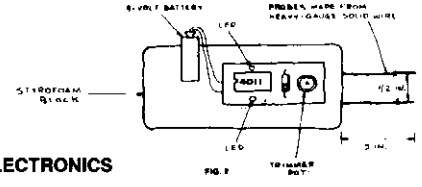


Fig. 36-7



RADIO-ELECTRONICS

Circuit Notes

To calibrate the gauge, connect the battery and press the probes gently into a pot containing a plant that is just on the verge of needing water (stick it in so that only an inch of the probe is left visible at the top). Turn the potentiometer until the "OK" LED lights and then turn it back to the point where that LED goes out and the "W", or "Water", LED just comes on. The device should now be properly adjusted.

LIQUID FLOWMETER

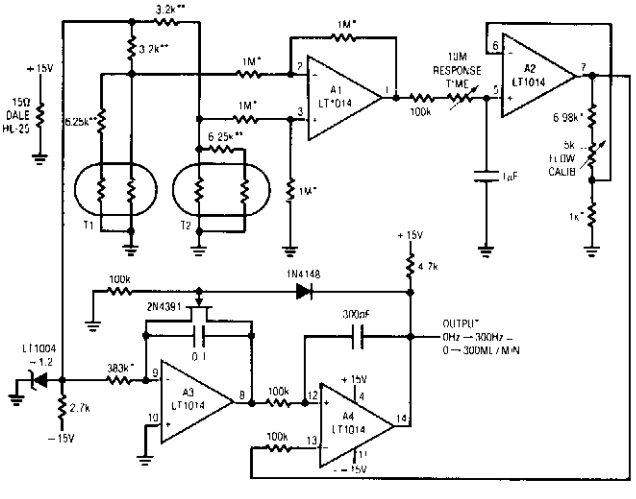
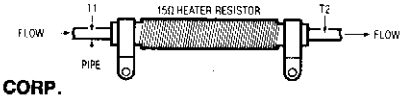


Fig. 36-8



*1% FILM RESISTOR
**SUPPLIED WITH YSI THERMISTOR NETWORK
T1, T2 YSI THERMISTOR NETWORK = #44201
FLOW IN PIPE IS INVERSELY PROPORTIONAL TO RESISTANCE OF T1–T2 TEMPERATURE DIFFERENCE.
A1–A2 PROVIDE GAIN. A3, A4 PROVIDE LINEARIZED FREQUENCY OUTPUT.

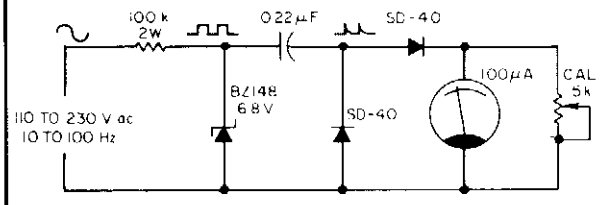
LINEAR TECHNOLOGY CORP.

37
Frequency Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Power Frequency Meter
Low Cost Frequency Indicator

POWER-FREQUENCY METER



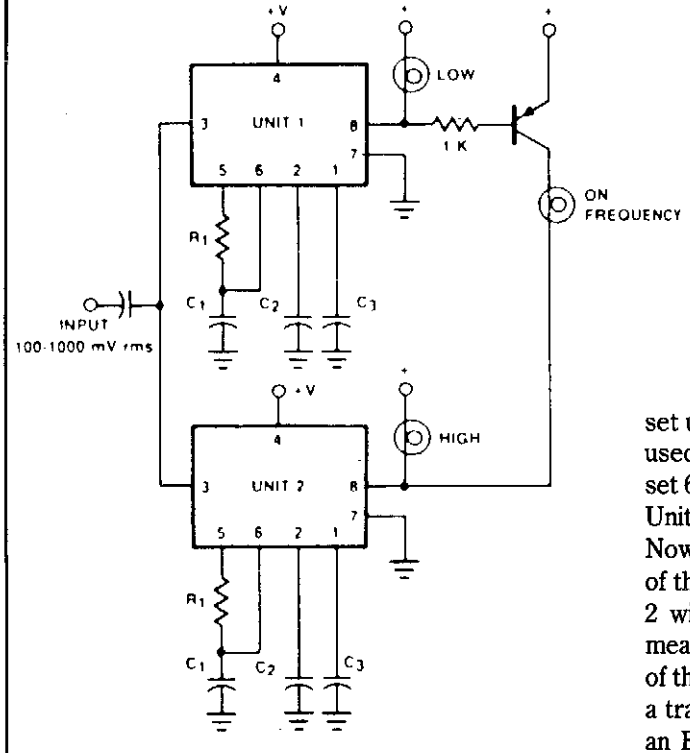
ELECTRONIC DESIGN

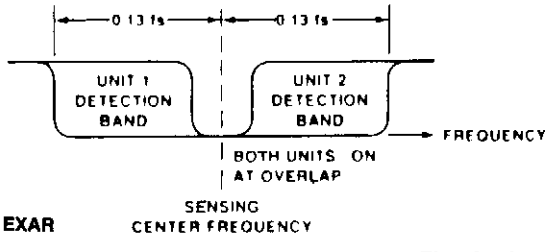
Fig. 37-1

Circuit Notes

The meter uses a zener diode to form square waves from input sine waves. After calibration with the 5 k ohm potentiometer, the 100 μA meter reads directly in hertz.

LOW COST FREQUENCY INDICATOR





EXAR

Fig. 37-2

Circuit Notes

The circuit shows how two tone decoders set up with overlapping detection bands can be used for a go/no go frequency meter. Unit 1 is set 6% above the desired sensing frequency and Unit 2 is set 6% below the desired frequency. Now, if the incoming frequency is within 13% of the desired frequency, either Unit 1 or Unit 2 will give an output. If both units are on, it means that the incoming frequency is within 1% of the desired frequency. Three light bulbs and a transistor allow low cost read-out. The IC is an EXAR 567.

38

Frequency Multiplier and Divider Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Nonselective Frequency Tripler Uses Transistor Saturation Characteristics
Frequency Doubler Works to 1 MHz
Low Frequency Divider
Frequency Divider

NONSELECTIVE FREQUENCY TRIPLER USES TRANSISTOR SATURATION CHARACTERISTICS

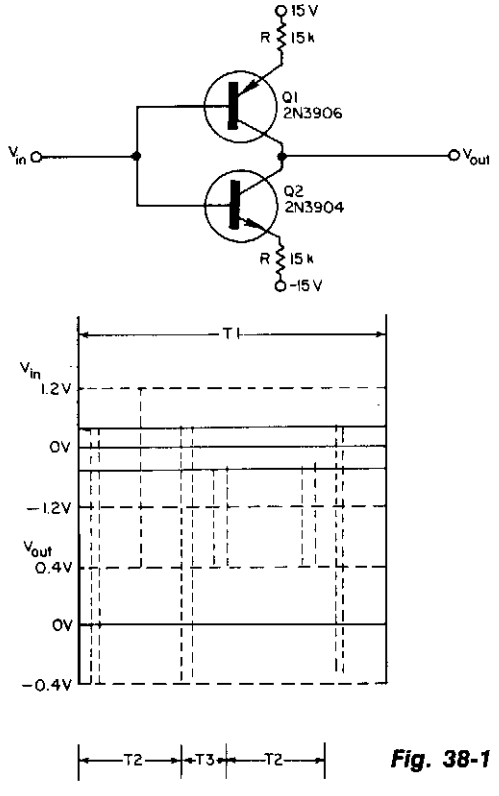


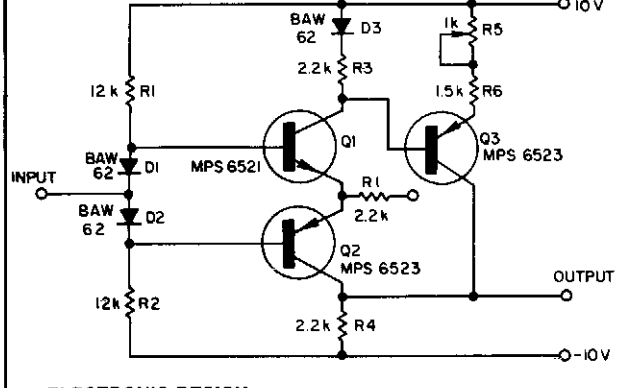
Fig. 38-1

ELECTRONIC DESIGN

Circuit Notes

The turn-on and turn-off characteristics of two complementary transistors can be combined to attain nonselective frequency tripling. The resulting circuit handles any periodic waveform with nonvertical sides. Each input signal peak produces three output signal peaks. The additional peaks occur where the input signal causes saturation of one of the two transistors. The circuit operates over a frequency range from dc to the upper limits of the complementary transistor pair. About the only disadvantage of the circuit is the lack of symmetry of the output signal peaks.

FREQUENCY DOUBLER WORKS TO 1 MHz



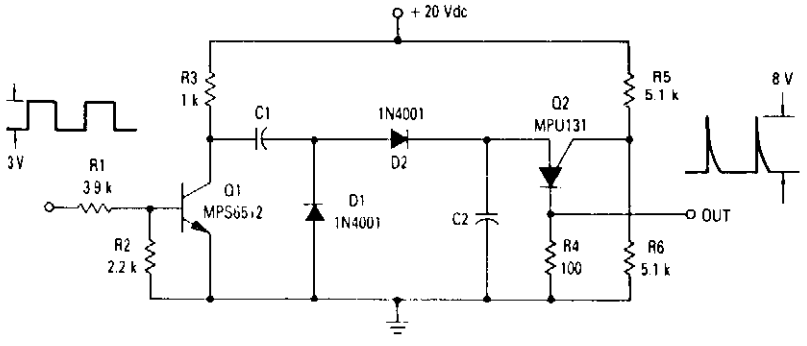
ELECTRONIC DESIGN

Fig. 38-2

Circuit Notes

Adding components Q3, D3, and resistors R3 through R6 to a conventional complementary symmetry class AB buffer can double the frequency of an input sine wave.

LOW FREQUENCY DIVIDER



TABLE

C_1	C_2	Division
0.01 μF	0.01 μF	2
0.01 μF	0.02 μF	3
0.01 μF	0.03 μF	4
0.01 μF	0.04 μF	5
0.01 μF	0.05 μF	6
0.01 μF	0.06 μF	7
0.01 μF	0.07 μF	8
0.01 μF	0.08 μF	9
0.01 μF	0.09 μF	10
0.01 μF	0.1 μF	11

MOTOROLA

Fig. 38-3

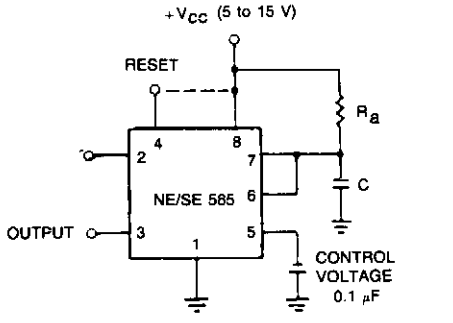
Circuit Notes

The ratio of capacitors C1 and C2 determines division. With a positive pulse applied to the base of Q1, assume that C1 = C2 and that C1 and C2 are discharged. When Q1 turns off, both C1 and C2 charge to 10 volts each through R3. On the next pulse to the base of Q1, C1 is again discharged but C2 remains charged to 10 volts. As Q1 turns off this time, C1 and C2 again charge. This time C2 charges to the peak point firing voltage of the PUT causing it to fire. This discharges capacitor C2 and allows capacitor C1 to charge to the line voltage. As soon as C2 discharges and C1 charges, the PUT turns off. The next cycle begins with another positive pulse on the base of Q1 which again discharges C1. The input and output frequency can be approximated by the equation

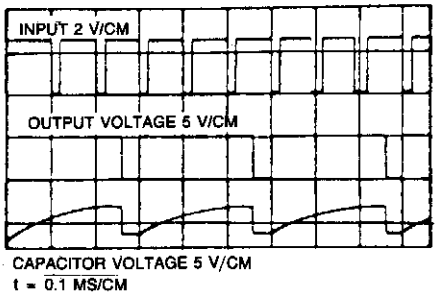
$$f_{in} = \frac{(C1 = C2)}{C1} f_{out}$$

For a 10 kHz input frequency with an amplitude of 3 volts, the table shows the values for C1 and C2 needed to divide by 2 to 11.

FREQUENCY DIVIDER



Schematic Diagram



Expected Waveforms

SIGNETICS

Fig. 38-4

Circuit Notes

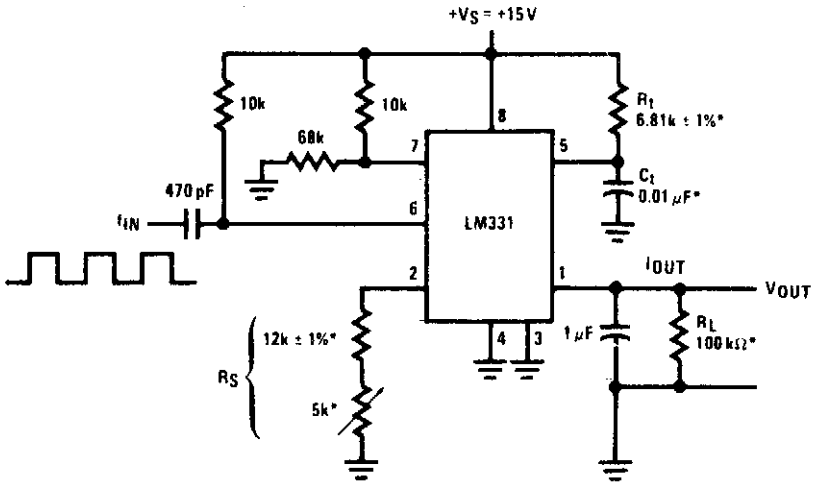
If the input frequency is known, the timer can easily be used as a frequency divider by adjusting the length of the timeing cycle. Figure shows the waveforms of the timer when used as a divide-by-three circuit. This application makes use of the fact that this circuit cannot be retriggered during the timing cycle.

39

Frequency-to-Voltage Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Frequency-to-Voltage Converters



TL/H/5680–7

$$V_{OUT} = f_{IN} \times 2.09V \times \frac{R_L}{R_S} \times (R_tC_t)$$

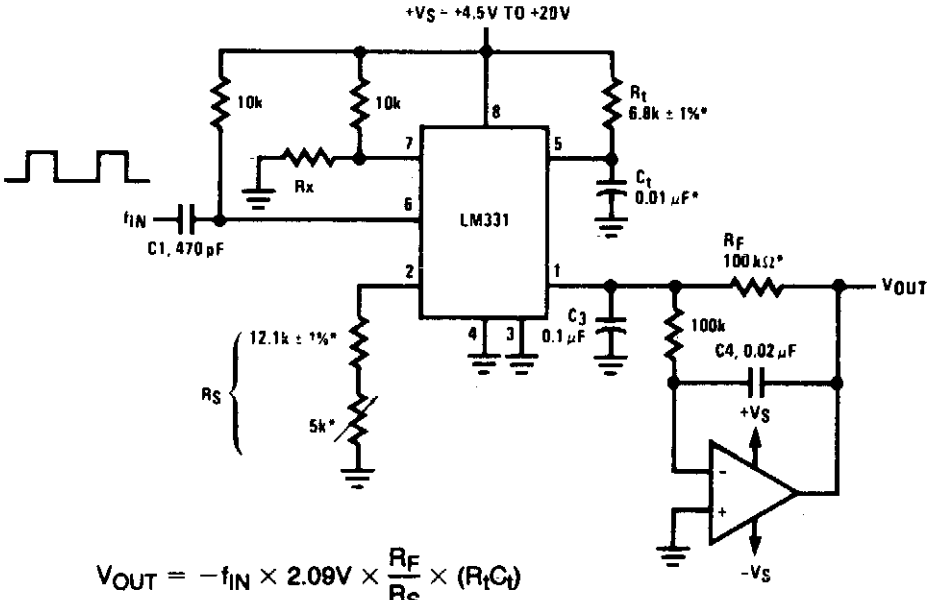
*Use stable components with low temperature coefficients.

Simple Frequency-to-Voltage Converter, 10 kHz Full-Scale, ±0.06% Non-Linearity

NATIONAL SEMICONDUCTOR CORP. *Fig. 39-1*

Circuit Notes

In these applications, a pulse input at f_{IN} is differentiated by a C-R network and the negative-going edge at pin 6 causes the input comparator to trigger the timer circuit. Just as with a V-to-F converter, the average current flowing out of pin 1 is $I_{AVERAGE} = i \times (1.1\ R_1C_1) \times f$. In this simple circuit, this current is filtered in the network R_L = 100 k ohm and 1 μF. The ripple will be less than 10 mV peak, but the response will be slow, with a 0.1 second time constant, and settling of 0.7 second to 0.1%



$$V_{OUT} = -f_{IN} \times 2.09V \times \frac{R_F}{R_S} \times (R_tC_t)$$

TL/H/5680–8

$$\text{SELECT Rx} = \frac{(V_S - 2V)}{0.2\ mA}$$

*Use stable components with low temperature coefficients.

Precision Frequency-to-Voltage Converter, 10 kHz Full-Scale with 2-Pole Filter, ±0.01% Non-Linearity Maximum

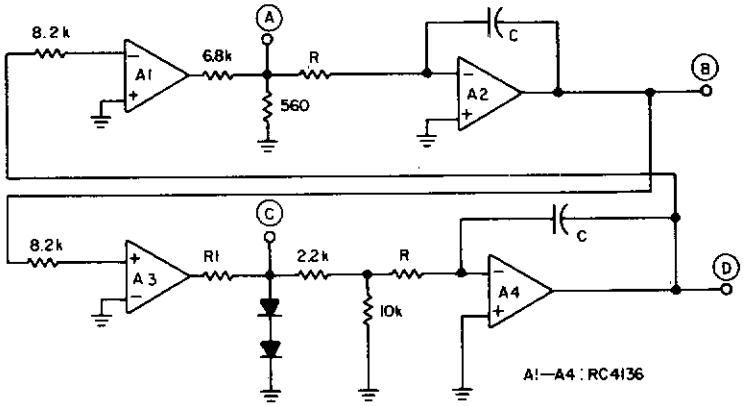
accuracy. In the precision circuit, an operational amplifier provides a buffered output and also acts as a 2-pole filter. The ripple will be less than 5 mV peak for all frequencies above 1 kHz, and the response time will be much quicker than in Part 1. However, for input frequencies below 200 Hz, this circuit will have worse ripple than the figure. The engineering of the filter time-constants to get adequate response and small enough ripple simply requires a study of the compromises to be made. Inherently, V-to-F converter response can be fast, but F-to-V response cannot.

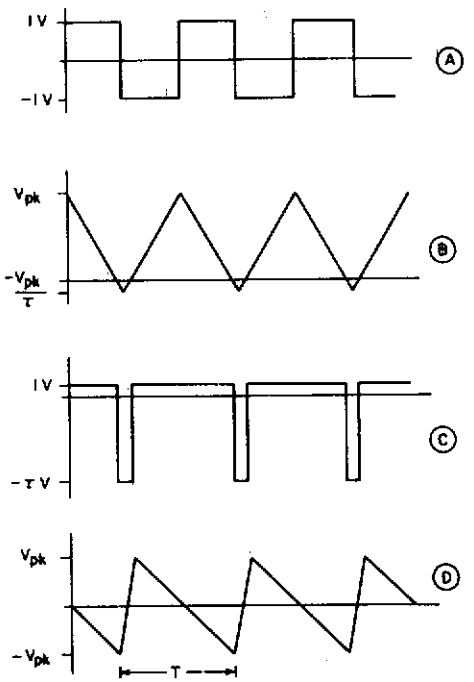
40
Function Generator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Quad Op Amp Generates Four Different Synchronized Waveforms Simultaneously
A Sine/Cosine Generator for 0.1-10 kHz
Oscillator or Amplifier with Wide Frequency Range
Linear Triangle/Square Wave VCO
Circuit for Multiplying Pulse Widths
Programmable Voltage Controlled Frequency Synthesizer
Emitter-Coupled RC Oscillator
Voltage Controlled High Speed One Shot
Ramp Generator with Variable Reset Level
555 Astable with Low Duty Cycle
Monostable Using Video Amplifier and Comparator
UJT Monostable Circuit Insensitive to Change in Bias Voltage
Astable Multivibrator
Waveform Generator
Linear Ramp Generator
Function Generator
Waveform Generator
Single Supply Function Generator
Precise Wave Generator

QUAD OP AMP GENERATES FOUR DIFFERENT SYNCHRONIZED WAVEFORMS SIMULTANEOUSLY





R1(kΩ)	τ	Duty cycle (%)
0.3	10.4	8.8
0.5	10.2	8.9
1	9.85	9.2
4	8.0	11.1
10	5.86	14.6
20	4.04	19.84
50	2.09	32.4
100	1.16	46.30
117.8	1	50

Fig. 40-1

ELECTRONIC DESIGN

Circuit Notes

A quad op amp can simultaneously generate four synchronized waveforms. The two comparators (A1 and A3) produce square and pulse waves, while the two integrators (A2 and A4) give triangular and sawtooth waves. Resistor R1 sets the duty cycle and the frequency, along with resistors R and capacitors C.

A SINE/COSINE GENERATOR FOR 0.1 - 10 kHz

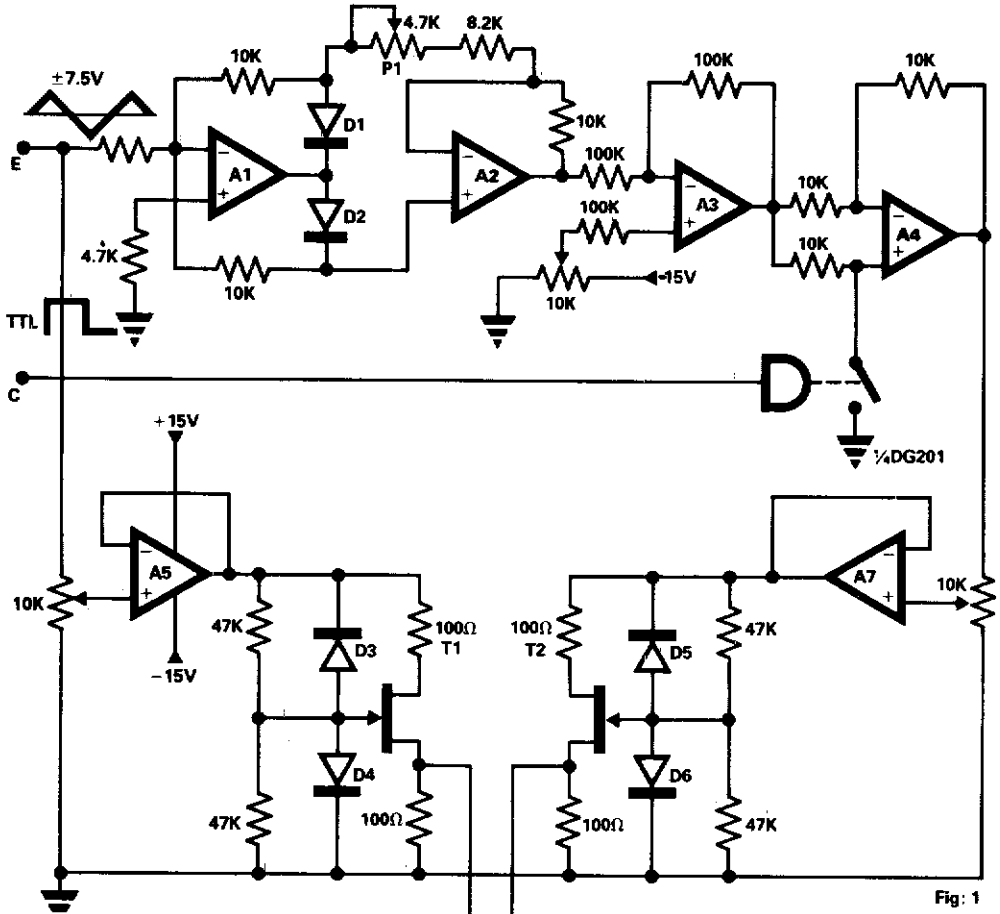


Fig: 1

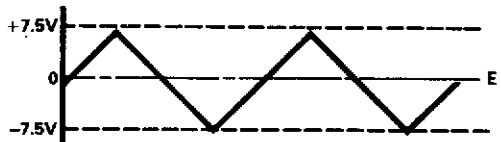


Fig. 40-2

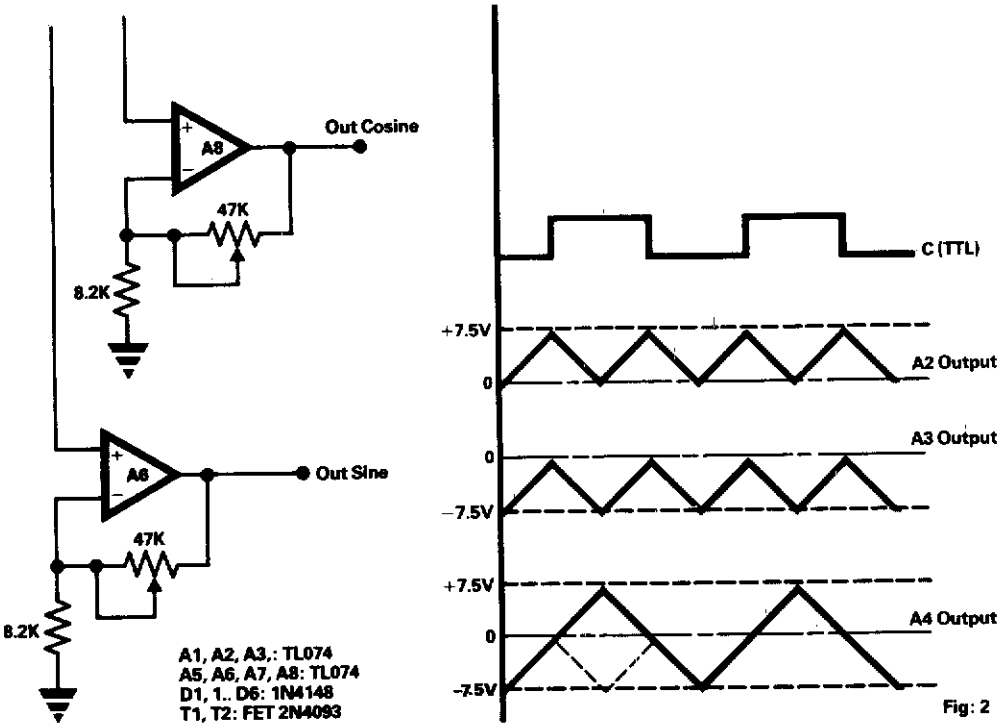


Fig: 2

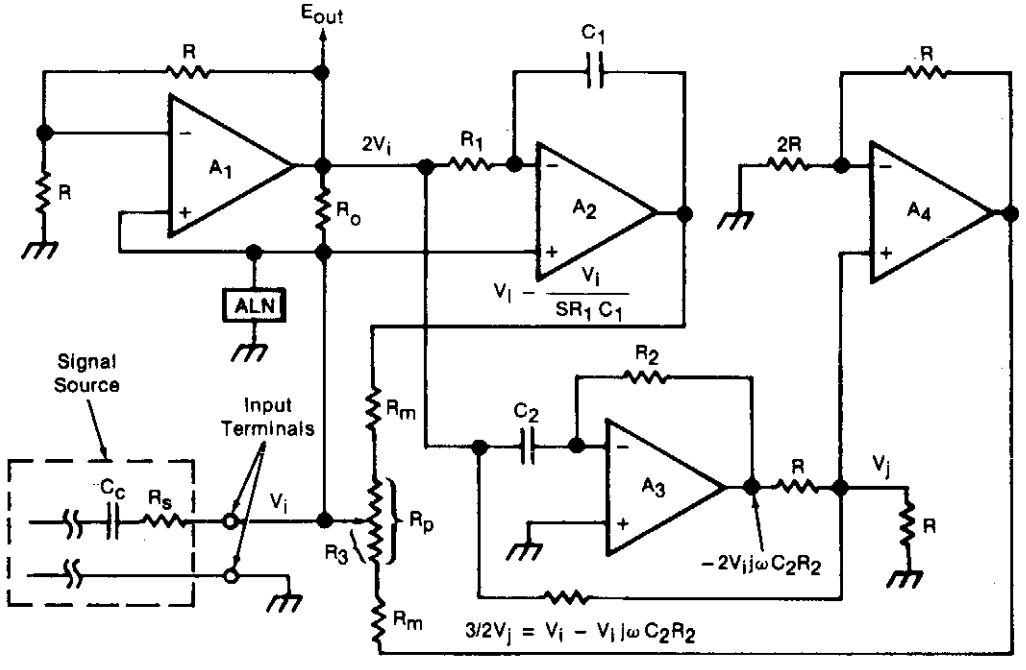
Circuit Notes

The scheme presented delivers waveforms from any function generator producing a triangular output and a synchronized TTL square wave. A1 and A2 act as a two-phase current rectifier by inverting the negative voltage appearing at the input of A1.

Positive input: Both A1 and A2 work as unity gain followers, D1 and D2 being in the off-state.

Negative input: A1 has a $-\frac{2}{3}$ gain (D1 off and D2 on), A2 has a $+\frac{1}{2}$ gain and the total voltage transfer is -1 between output and input. P1 allows a fine trimming of the -1 gain for the negative input signals. A3 adds a continuous voltage to the rectified positive signal in order to attack A4 which acts as a $\pm$ multiplier commanded by the TTL input through the analog switch. The signal polarity is reconstructed and the output of A4 delivers a triangular waveform shifted by 90° with respect to the input signal, Fig. 2. The original and the shifted voltages are fed into the triangle to sine converters through A5 and A7 working as impedance converters. Over the frequency dynamic ranges from 0.1 Hz to 10 kHz, the phase shift is constant and the distortion on the sine voltage is less than 1%.

OSCILLATOR OR AMPLIFIER WITH WIDE FREQUENCY RANGE



NOTES: 1. A_1, A_2, A_3, and A_4 are operational amplifiers
2. ALN = Amplitude-Limiting Network

NASA

Fig. 40-3

Circuit Notes

An oscillator/amplifier is resistively tunable over a wide frequency range. Feedback circuits containing operational amplifiers, resistors, and capacitors synthesize the electrical effects of an inductance and capacitance in parallel between the input terminals. The synthetic inductance and capacitance, and, therefore, the resonant frequency of the input admittance, are adjusted by changing a potentiometer setting. The input signal is introduced in parallel to the noninverting input terminals of operational amplifiers A_1 and A_2 and to the potentiometer cursor. The voltages produced by the feedback circuits in response to input voltage V_1 are indicated at the various circuit nodes.

LINEAR TRIANGLE/SQUARE WAVE VCO

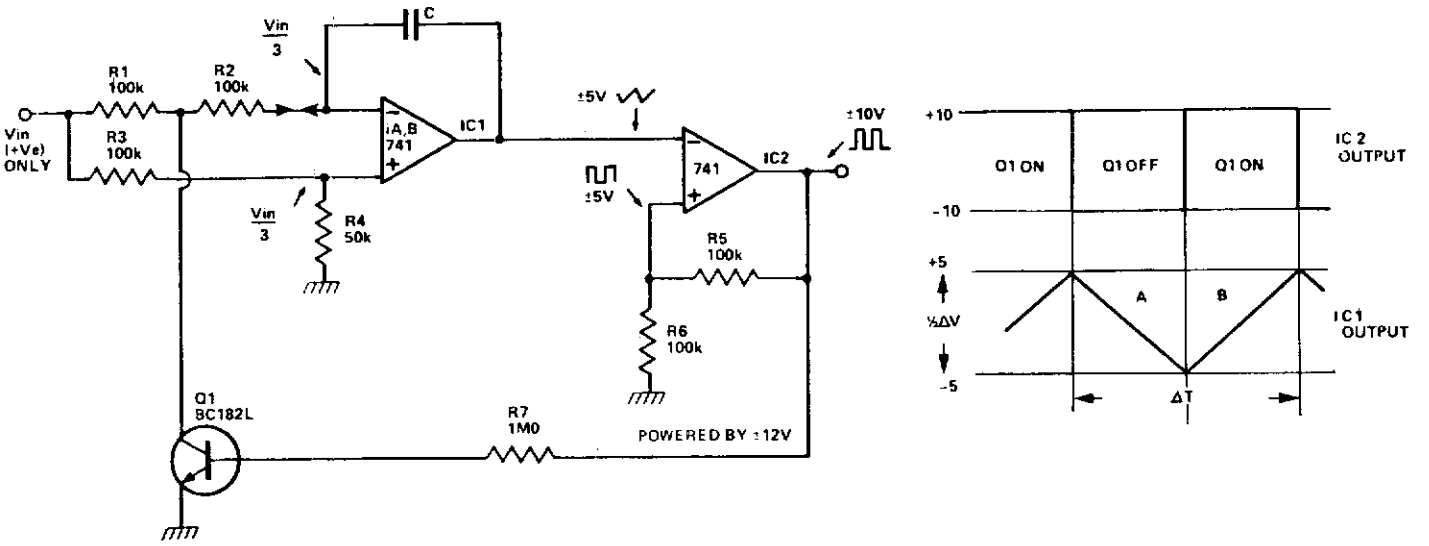


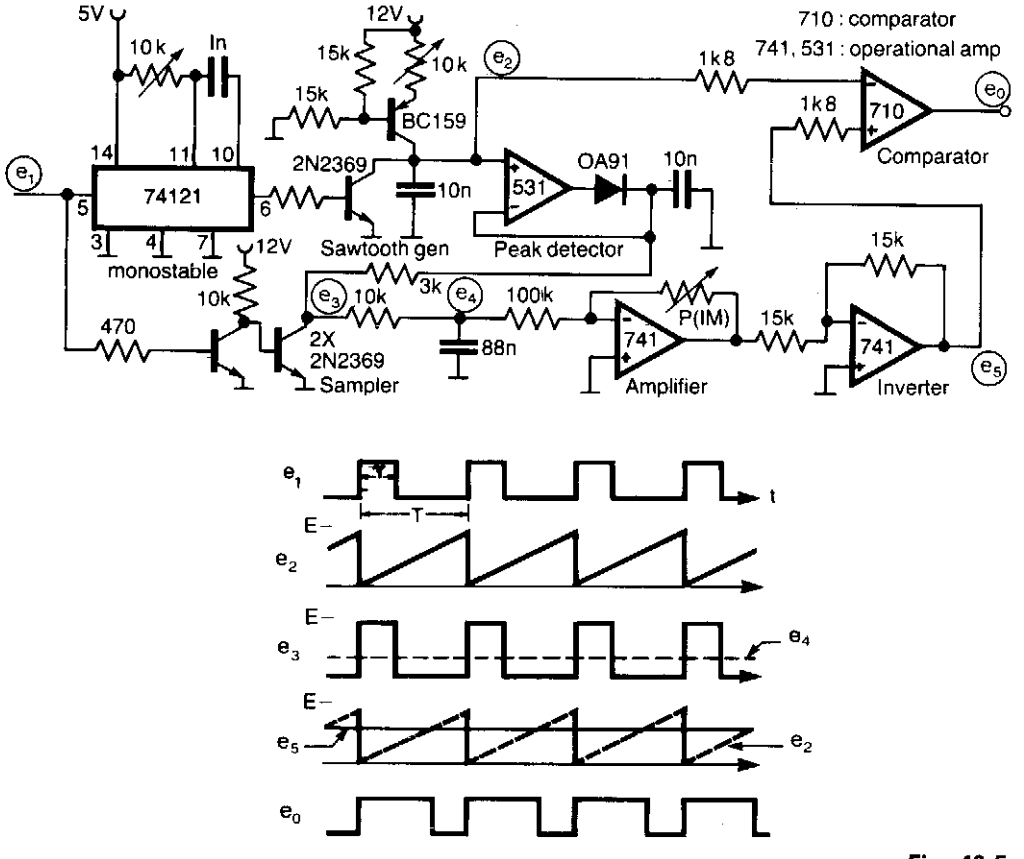
Fig. 40-4

ELECTRONICS TODAY INTERNATIONAL

Circuit Notes

The VCO has two buffered outputs; a triangle wave and a square wave. Frequency is dependent on the output voltage swing of the Schmitt trigger, IC2. Superior performance can be obtained by replacing Q1 with a switching FET. Fast FET op amps will improve high frequency performance.

CIRCUIT FOR MULTIPLYING PULSE WIDTHS



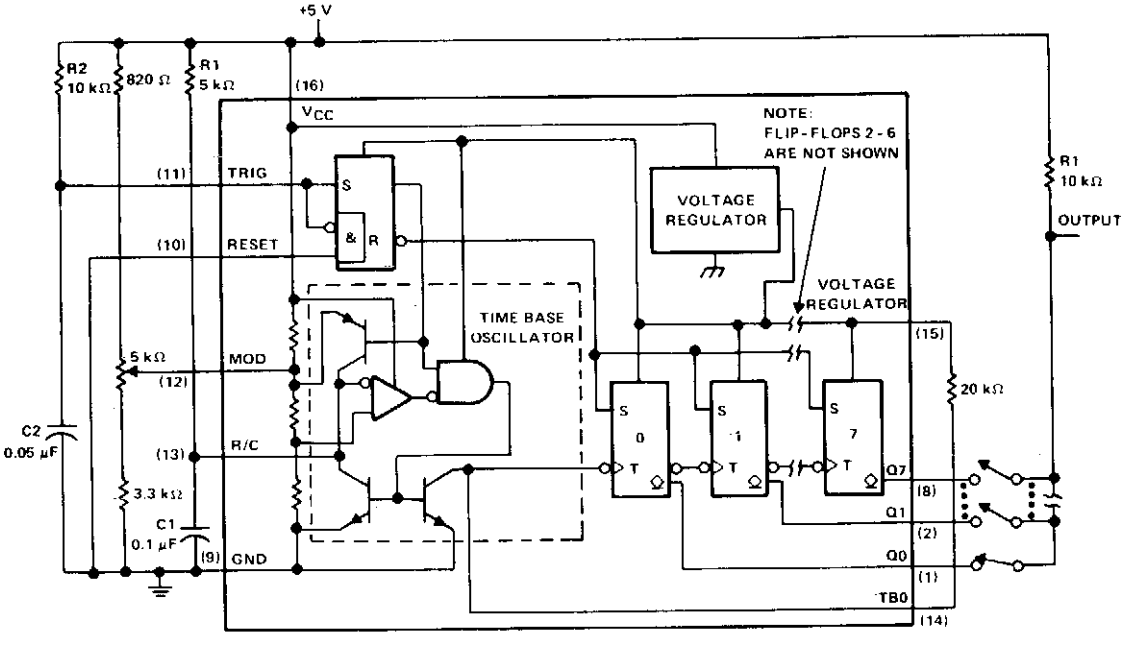
ELECTRONIC ENGINEERING

Fig. 40-5

Circuit Notes

A circuit for multiplying the width of incoming pulses by a factor greater or less than unity is simple to build and has the feature that the multiplying factor can be selected by adjusting one potentiometer only. The multiplying factor is determined by setting the potentiometer P in the feedback of a 741 amplifier. The input pulses e_1 of width τ and repetition period T is used to trigger a sawtooth generator at its rising edges to produce the waveform e_2 having a peak value of (E) volt. This peak value is then sampled by the input pulses to generate the pulse train e_3 having an average value of $e_4 (= E\ E\tau/T)$ which is proportional to τ and independent on T. The dc voltage e_4 is amplified by a factor k and compared with sawtooth waveform e_2 giving output pulses of duration k τ. The circuit is capable of operating over the frequency range 10 kHz - 100 kHz.

PROGRAMMABLE VOLTAGE CONTROLLED FREQUENCY SYNTHESIZER



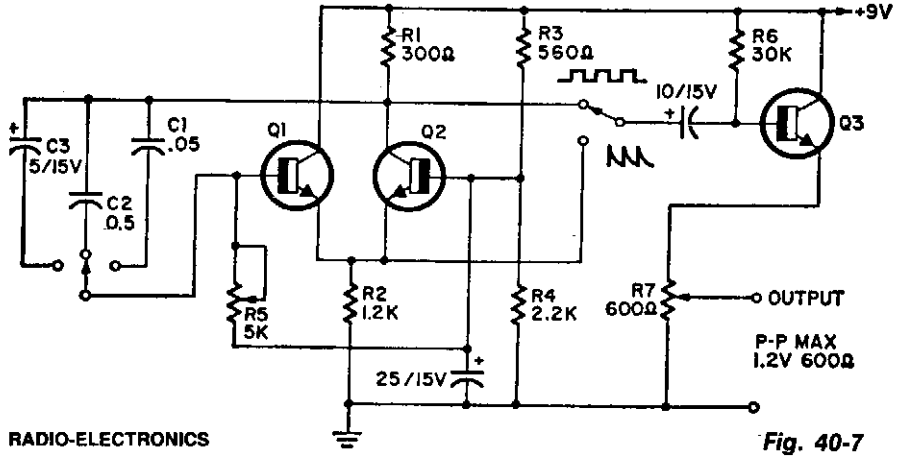
TEXAS INSTRUMENTS

Fig. 40-6

Circuit Notes

The μA2240 consists of four basic circuit elements: (1) a time-base oscillator, (2) an eight-bit counter, (3) a control flip-flop, and (4) a voltage regulator. The basic frequency of the time-base oscillator (TBO) is set by the external time constant determined by the values of R1 and C1 (1?R1C1 = 2 kHz). The open-collector output of the TBO is connected to the regulator output via a 20 k ohm pull-up resistor, and drives the input to the eight-bit counter. At power-up, a positive trigger pulse is detected across C2 which starts the TBO and sets all counter outputs to a low state. Once the μA2240 is initially triggered, any further trigger inputs are ignored until it is reset. In this astable operation, the μA2240 will free-run from the time it is triggered until it receives an external reset signal. Up to 255 discrete frequencies can be synthesized by connecting different counter outputs.

EMITTER-COUPLED RC OSCILLATOR



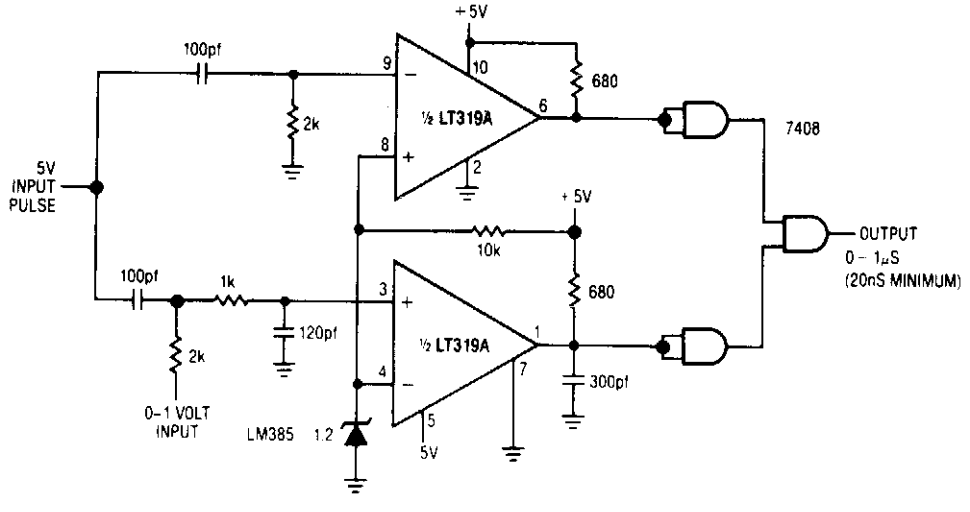
RADIO-ELECTRONICS

Fig. 40-7

Circuit Notes

The circuit covers 15 Hz-30 kHz and is useful as a function generator. The 2N2926 or equivalent transistors can be used.

VOLTAGE CONTROLLED HIGH SPEED ONE SHOT



LINEAR TECHNOLOGY CORP.

Fig. 40-8

RAMP GENERATOR WITH VARIABLE RESET LEVEL

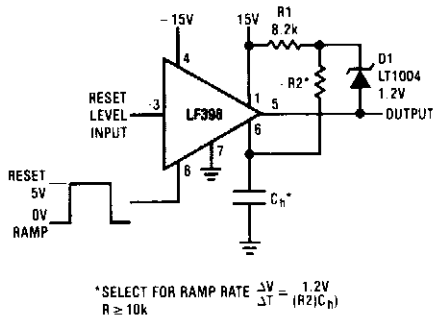
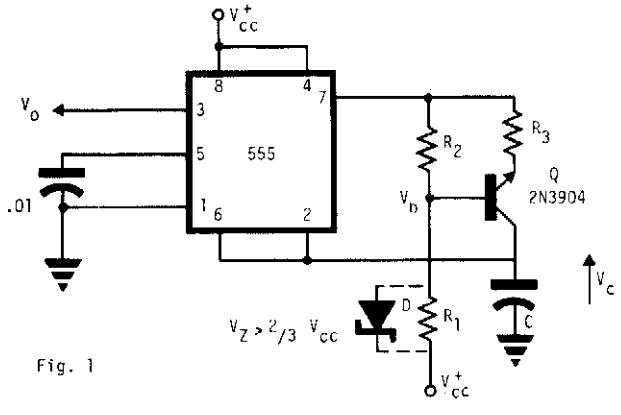
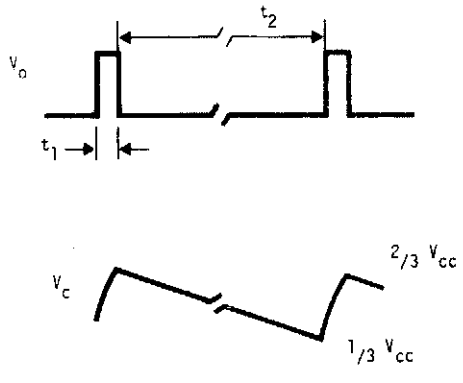


Fig. 40-9

LINEAR TECHNOLOGY CORP.

555 ASTABLE WITH LOW DUTY CYCLE





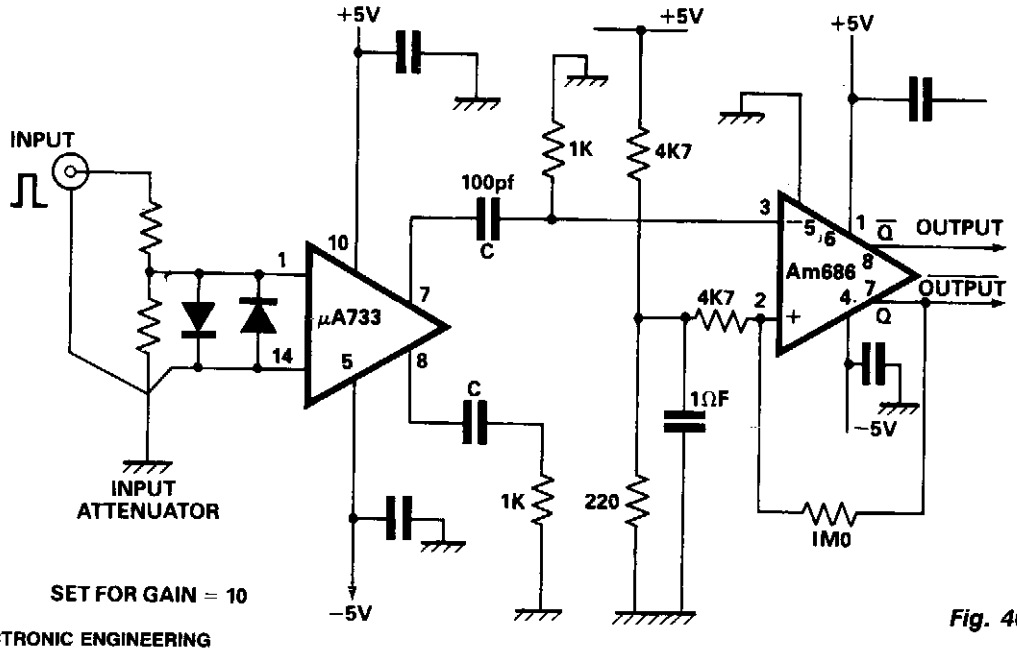
ELECTRONIC ENGINEERING

Fig. 40-10

Circuit Notes

This free-running multivibrator uses an external current sink to discharge the timing capacitor, C. Therefore, interval t_2 may easily be 1000 × the pulse duration, t_1, which defines a positive output. Capacitor voltage, V_C, is a negative going ramp with exponential rise during the pulse output periods.

MONOSTABLE USING VIDEO AMPLIFIER AND COMPARATOR



ELECTRONIC ENGINEERING

Fig. 40-11

Circuit Notes

The output of a video amplifier is differentiated before being fed to a Schottky comparator. The propagation delay is reduced to typically 10ns. The output pulse width is set by the value of C, 100pf giving a pulse of about 90ns duration.

UJT MONOSTABLE CIRCUIT INSENSITIVE TO CHANGE IN BIAS VOLTAGE

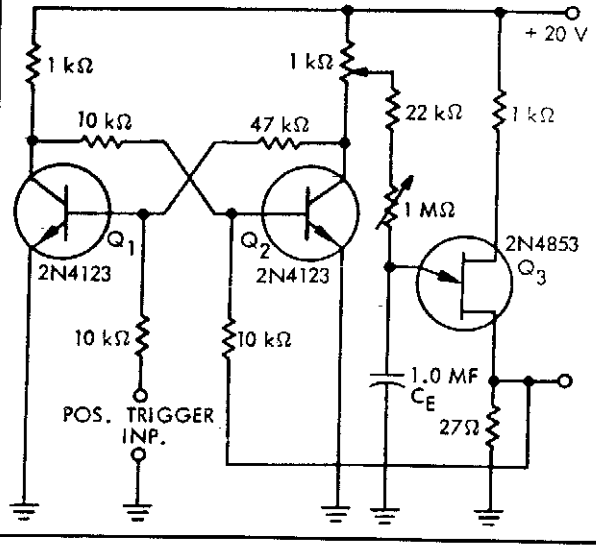
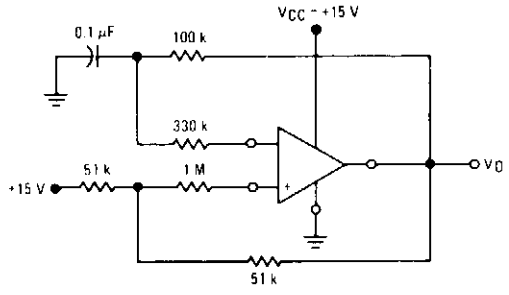


Fig. 40-12

MOTOROLA INC.

ASTABLE MULTIVIBRATOR

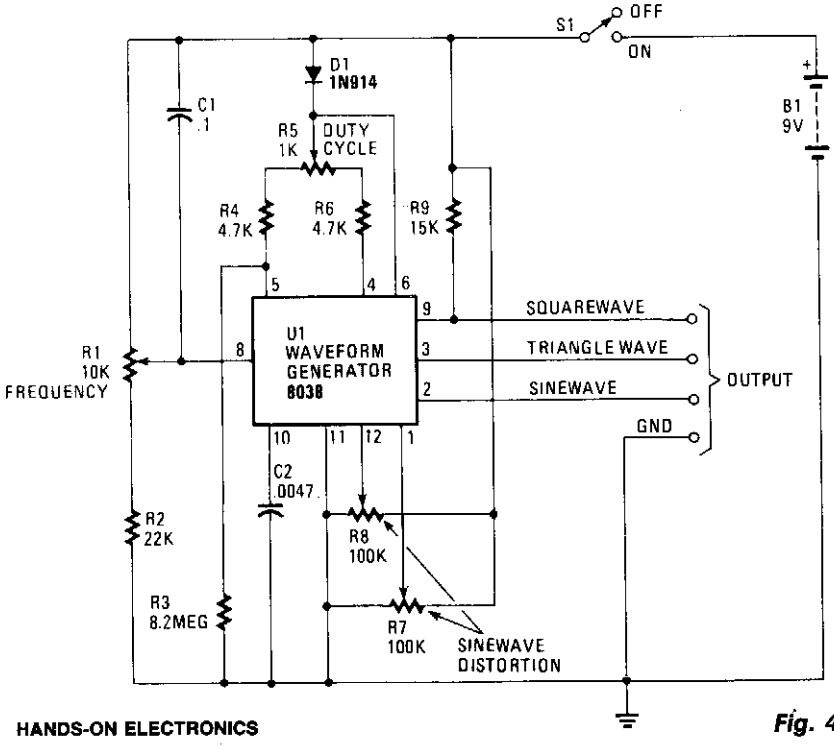


MOTOROLA INC.

Fig. 40-13

IC = MC3301

WAVEFORM GENERATOR



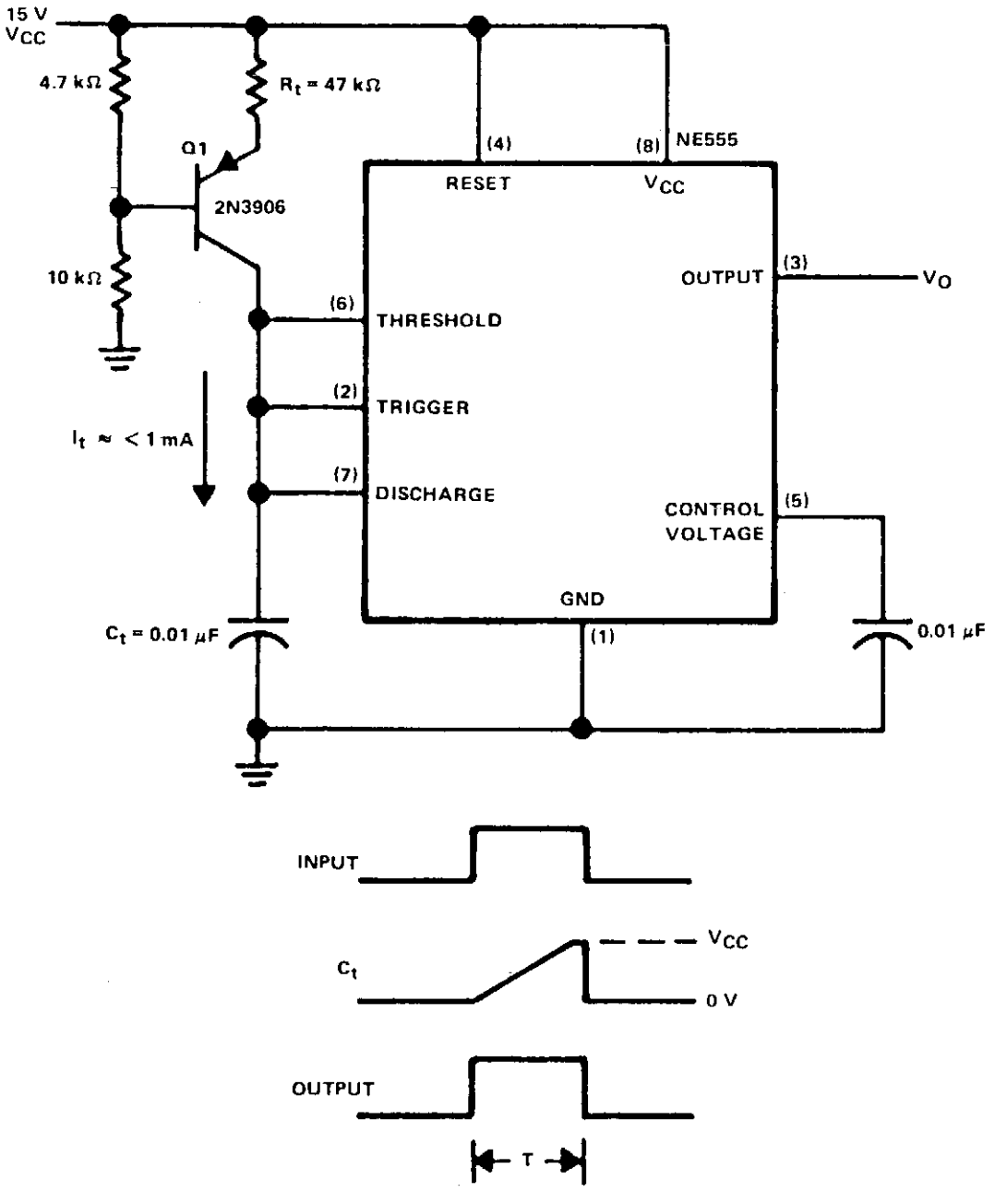
HANDS-ON ELECTRONICS

Fig. 40-14

Circuit Notes

The circuit is designed around the Intersil 8038CC. Frequency range is approximately 20 Hz to 20 kHz—a tuning range of 1000:1 with a single control. The output frequency depends on the value of C2 and on the setting of potentiometer R1. Other values of C2 change the frequency range. Increase the value of C2 to lower the frequency. The lowest possible frequency is around .001 Hz and the highest is around 300 kHz.

LINEAR RAMP GENERATOR



TEXAS INSTRUMENTS

Fig. 40-15

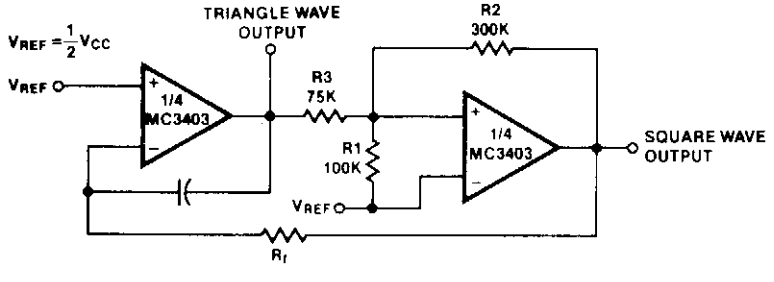
LINEAR RAMP GENERATOR, Continued.

Circuit Notes

The linear charging ramp is most useful where linear control of voltage is required. Some possible applications are a long period voltage controlled timer, a voltage to pulse width converter, or a linear pulse width modulator. Q1 is the current source transistor, supplying constant current to the timing capacitor C_t. When the timer is triggered, the clamp on C_t is removed and C_t charges linearly toward V_{CC} by virtue of the constant current supplied by Q1. The threshold at pin 6 is ⅔ V_{CC}; here, it is termed V_C. When the voltage across C_t reaches V_C volts, the timing cycle ends. The timing expression for output pulse with T is:

In general, I_t should be 1 mA value compatible with the NE555.

FUNCTION GENERATOR

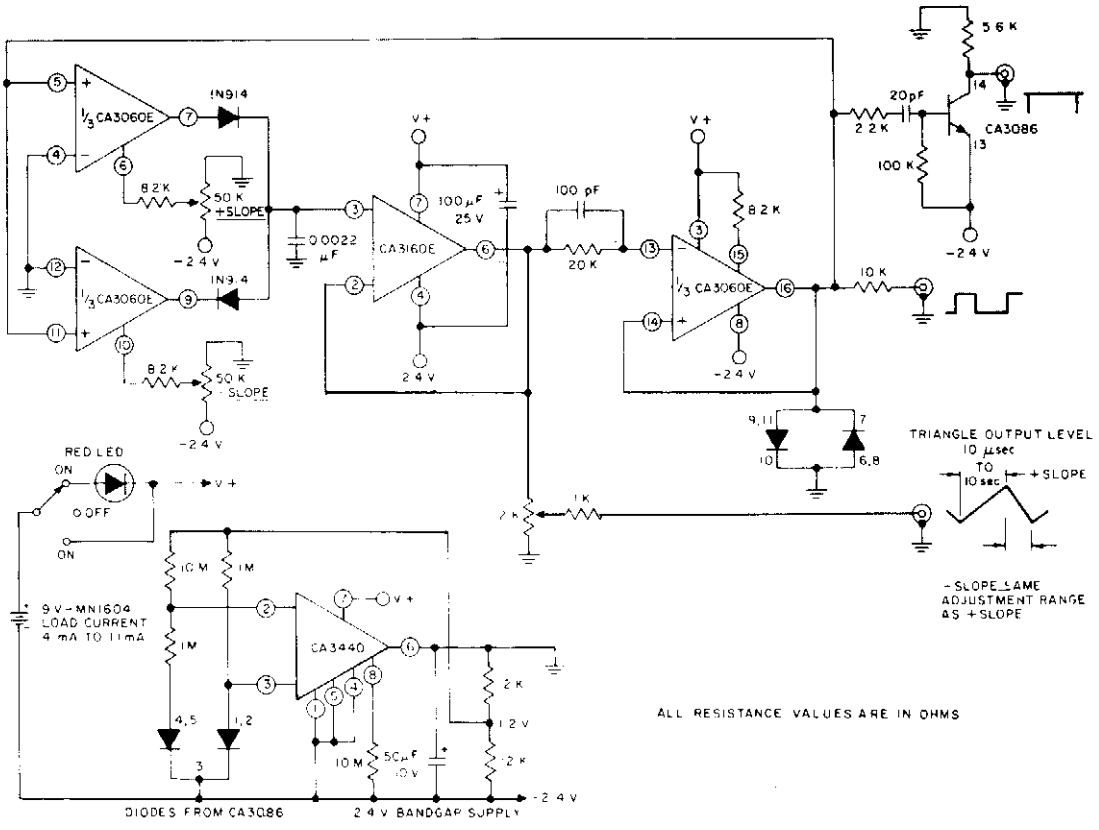


$$f = \frac{R1 + R_C}{4 C R_f R1} \quad \text{IF } R3 = \frac{R2\ R1}{R2 + R1}$$

SIGNETICS

Fig. 40-16

WAVEFORM GENERATOR



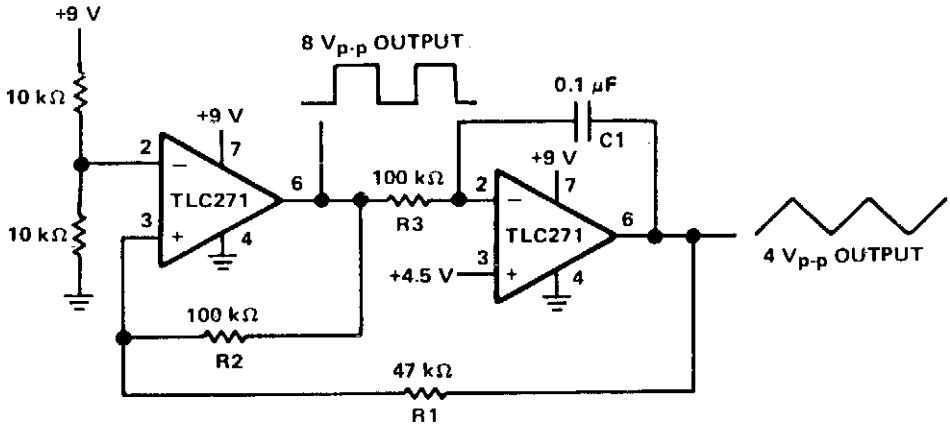
GENERAL ELECTRIC/RCA

Fig. 40-17

Circuit Notes

The circuit uses a CA3060 triple OTA (two units serve as switched current generators controlled by a third amplifier). A CA3160 BiMOS op amp serves as a voltage follower to buffer the 0.0022 μF integrating capacitor. The circuit has an adjustment range of 1,000,000:1 and a timing range of 20 μs to 20 sec. The "ON-OFF" switch actuates an LED that serves as both a pilot light and a low-battery indicator. The LED extends battery life, since it drops battery voltage to the circuit by approximately 1.2 volts, thus reducing supply current.

SINGLE SUPPLY FUNCTION GENERATOR



TEXAS INSTRUMENTS

Fig. 40-18

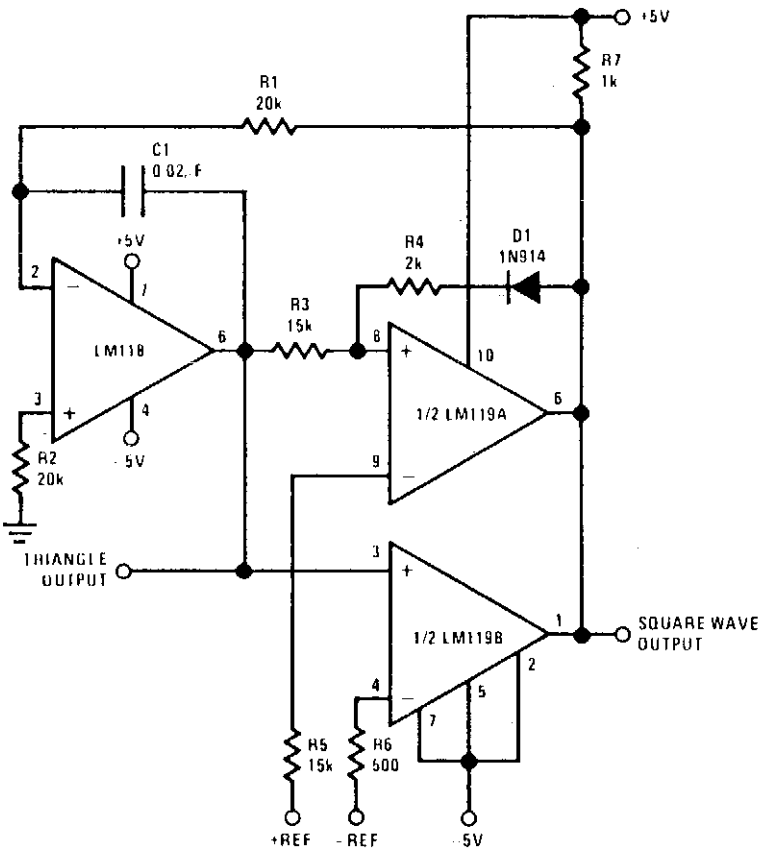
Circuit Notes

The circuit has both square-wave and triangle-wave output. The left section is similar in function to a comparator circuit that uses positive feedback for hysteresis. The inverting input is biased at one-half the V_{CC} voltage by resistors R4 and R5. The output is fed back to the non-inverting input of the first stage to control the frequency. The amplitude of the square wave is the output swing of the first stage, which is 8 V peak-to-peak. The second stage is basically an op amp integrator. The resistor R3 is the input element and capacitor C1 is the feedback element. The ratio R1/R2 sets the amplitude of the triangle wave, as referenced to the square-wave output. For both waveforms, the frequency of oscillation can be determined by the equation:

$$fo = \frac{1}{4R3C1} \quad \frac{R2}{R1}$$

The output frequency is approximately 50 Hz with the given components.

PRECISE WAVE GENERATOR



NATIONAL SEMICONDUCTOR

Fig. 40-19

Circuit Notes

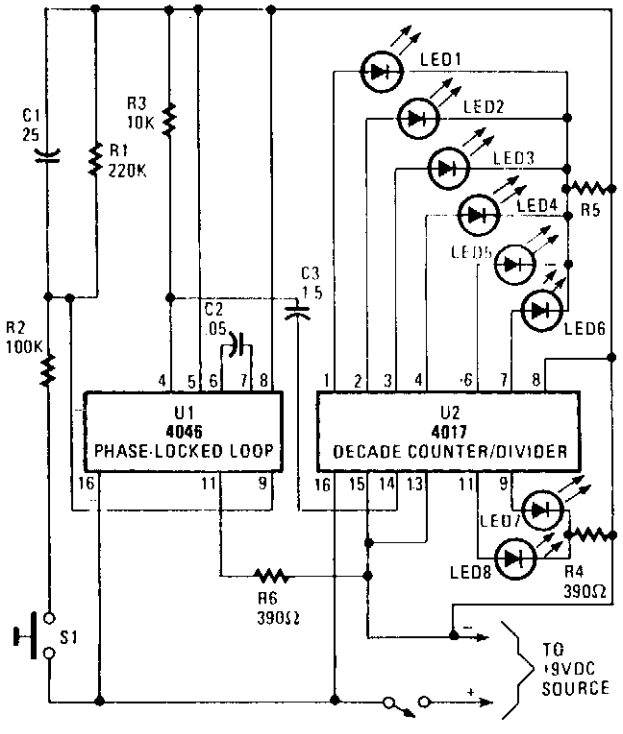
The positive and negative peak amplitude is controllable to an accuracy of about ±0.01 V by a dc input. Also, the output frequency and symmetry are easily adjustable. The oscillator consists of an integrator and two comparators—one comparator sets the positive peak and the other the negative peak of the triangle wave. If R1 is replaced by a potentiometer, the frequency can be varied over at least a 10 to 1 range without affecting amplitude. Symmetry is also adjustable by connecting a 50 kΩ resistor from the inverting input of the LM118 to the arm of the 1 kΩ potentiometer. The ends of the potentiometer are connected across the supplies. Current for the resistor either adds or subtracts from the current through R1, changing the ramp time.

41
Games

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Electronic Roulette
Lie Detector

ELECTRONIC ROULETTE



TAB BOOKS, INC.

Fig. 41-1

Circuit Notes

U1 (a 4046 PLL containing a voltage controlled oscillator or VCO, two phase comparators, a source follower, and a Zener diode) is used to produce a low-frequency, pulsed output of about 40 Hz. The VCO's frequency range is determined by R6 and C2, which can be altered by varying the voltage at pin 9. The rising voltage causes the frequency to rise from zero to threshold and remain at that frequency as long as S1 is closed. When S1 is opened, C1 discharges slowly through R1 to ground and the voltage falls toward zero. That produces a decreasing pulse rate. The output of U1 at pin 4 is connected to the clock input of U2 (a 4017 decade decoder/driver) at pin 14 via C3. U2 sequentially advances through each of its ten outputs (0 to 9)—pins 1 to 7, and 9 to 11—with each input pulse. As each output goes high, its associated LED is lighted, and extinguished when it returns to the low state. Only eight outputs are used in the circuit, giving two numbers to the spinner of the house. The circuit can be set up so that the LED's lights sequence or you can use some staggered combination; the LEDs grouped in a straight line or a circle.

LIE DETECTOR

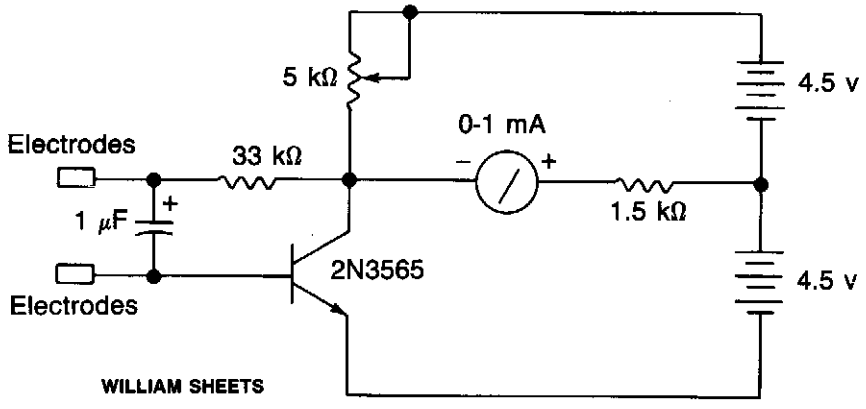


Fig. 41-2

Circuit Notes

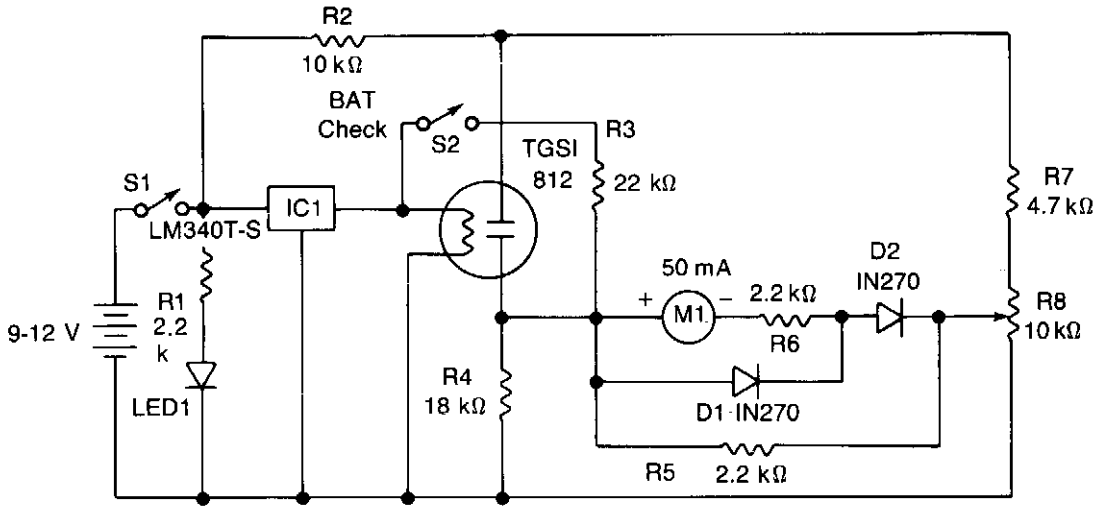
The two probes shown are held in the hands and the skin resistance applies bias to the transistor. The 5 k ohm pot is set for zero deflection on the meter. When the "subject" is embarrassed or lies, sweating on the hands takes place, increasing the bias to the transistor and upsetting the bridge balance.

42
Gas and Smoke Detectors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Gas and Vapor Detector
Toxic Gas Detector
Gas Analyzer

GAS AND VAPOR DETECTOR



WILLIAM SHEETS

Fig. 42-1

Circuit Notes

The power drain is approximately 150 mA. IC1 provides a regulated 5-volt supply for the filament heater of the sensor. The gas sensitive element is connected as one arm of a resistance bridge consisting of R4, R7, R8 and the meter M1 with its associated resistors R5 and R6. The bridge can be balanced by adjusting R8 so that no current flows through the meter. A change in the sensor's resistance, caused by detection of noxious gases, will unbalance the circuit and deflect the meter. Diodes D1, D2 and resistor R5 protect the meter from overload while R6 determines overall sensitivity. R2 limits the current through the sensor; R1 and LED1 indicate that the circuit is working, so that you do not drain the battery leaving the unit on inadvertently; R3 and S2 give a battery level check.

TOXIC GAS DETECTOR

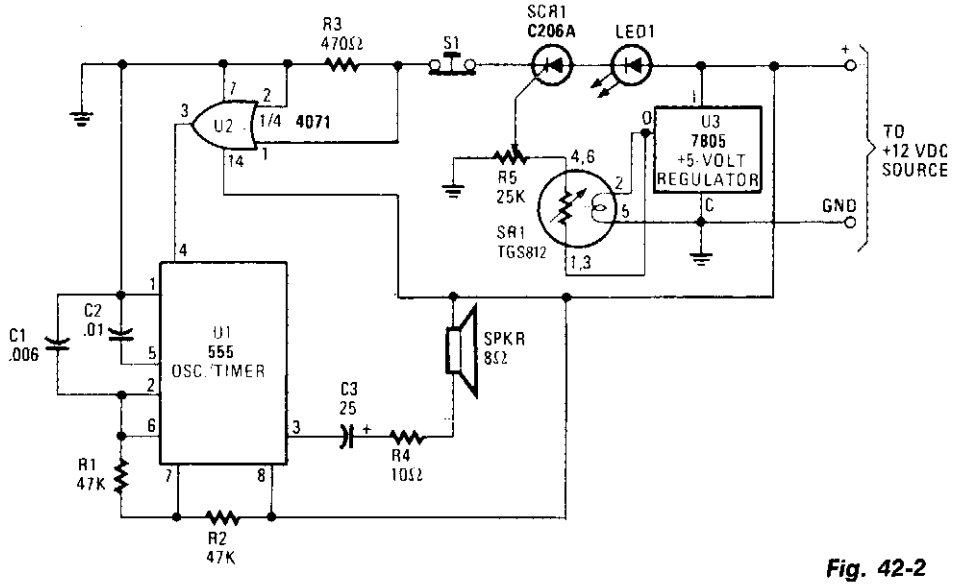


Fig. 42-2

Circuit Notes

The major device in the circuit is SR1 (a TGS812 toxic-gas sensor manufactured by Figaro Engineering Inc.) The gas-sensitive semiconductor (acting like a variable resistor in the presence of toxic gas) decreases in electrical resistance when gaseous toxins are absorbed from the sensor surface. A 25,000 ohm potentiometer (R5) connected to the sensor serves as a load, voltage-dividing network, and sensitivity control and has its center tap connected to the gate of SCR1. When toxic fumes come in contact with the sensor, decreasing its electrical resistance, current flows through the load (potentiometer R5). The voltage developed across the wiper of R5, which is connected to the gate of SCR1, triggers the SCR into conduction. With SCR1 now conducting, pin 1-volt supply for the semiconductor elements of the TGS812 in spite of the suggested 10 volts, thus reducing the standby current. A 7805 regulator is used to meet the 5-volt requirement for the heater and semiconductor elements.

GAS ANALYZER

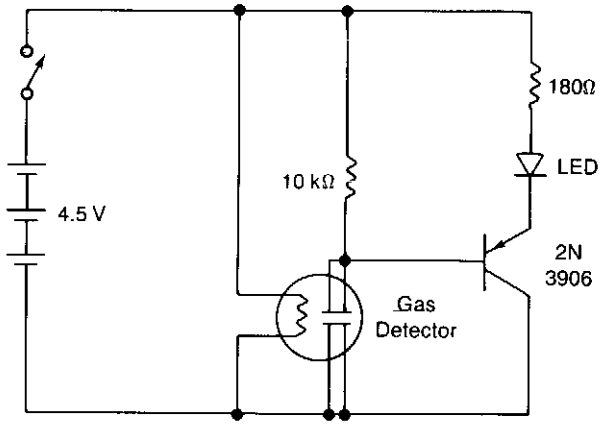


Fig. 42-3

WILLIAM SHEETS

Circuit Notes

The circuit shows a simple yes/no gas detector. Three 1.5-V D cells are used as a power supply, with S1 acting as an on/off switch. The heater is energized directly from the battery, while the electrodes are in series with a 10 k resistor. The voltage across this resistor is monitored by a pnp transistor. When the sensor is in clean air, the resistance between the electrodes is about 40 k, so that only about 0.9 V is dropped across the 10 k resistor. This is insufficient to turn on the transistor, because of the extra 1.6 V required to forward bias the light emitting diode (LED) in series with the emitter. When the sensor comes in contact with contaminated air, the resistance starts to fall, increasing the voltage dropped across the 10 k resistor. When the sensor resistance falls to about 10 k or less, the transistor starts to turn on, current passes through the LED, causing it to emit. The 180 ohm resistor limits the current through the LED to a safe value.

43
Hall Effect Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Angle of Rotation Detector
Door Open Alarm

ANGLE OF ROTATION DETECTOR

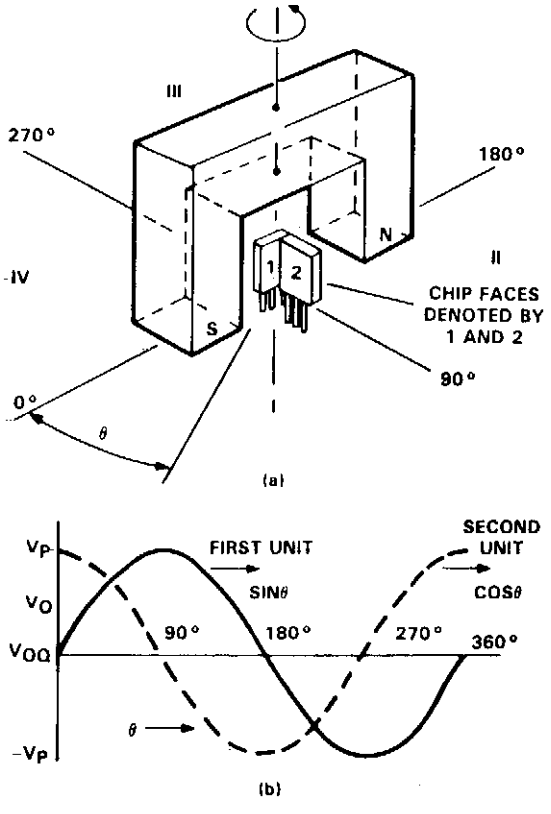


Fig. 43-1

TEXAS INSTRUMENTS

Circuit Notes

The figure shows two TL3103 linear Hall-effect devices used for detecting the angle of rotation. The TL3103s are centered in the gap of a U-shaped permanent magnet. The angle that the south pole makes with the chip face of unit #1 is defined as angle 0. Angle 0 is set to 0° when the chip face of unit #1 is perpendicular to the south pole of the magnet. As the south pole of the magnet sweeps through a 0° to 90° angle, the output of the sensor increases from 0°. Sensor unit #2 decreases from its peak value of +Vp at 0° to a value V_{OQ} at 90°. So, the output of sensor unit #1 is a sine function of 0 and the output of unit #2 is a cosine function of 0 as shown. Thus, the first sensor yields the angle of rotation and the second sensor indicates the quadrant location.

DOOR OPEN ALARM

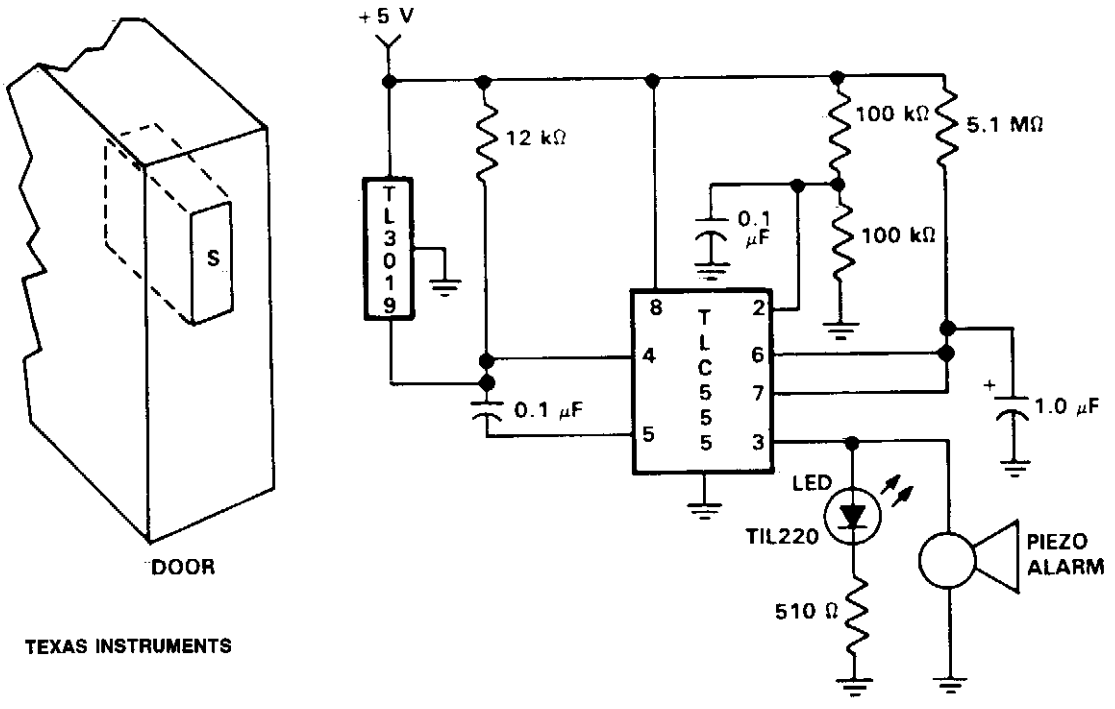


Fig. 43-2

Circuit Notes

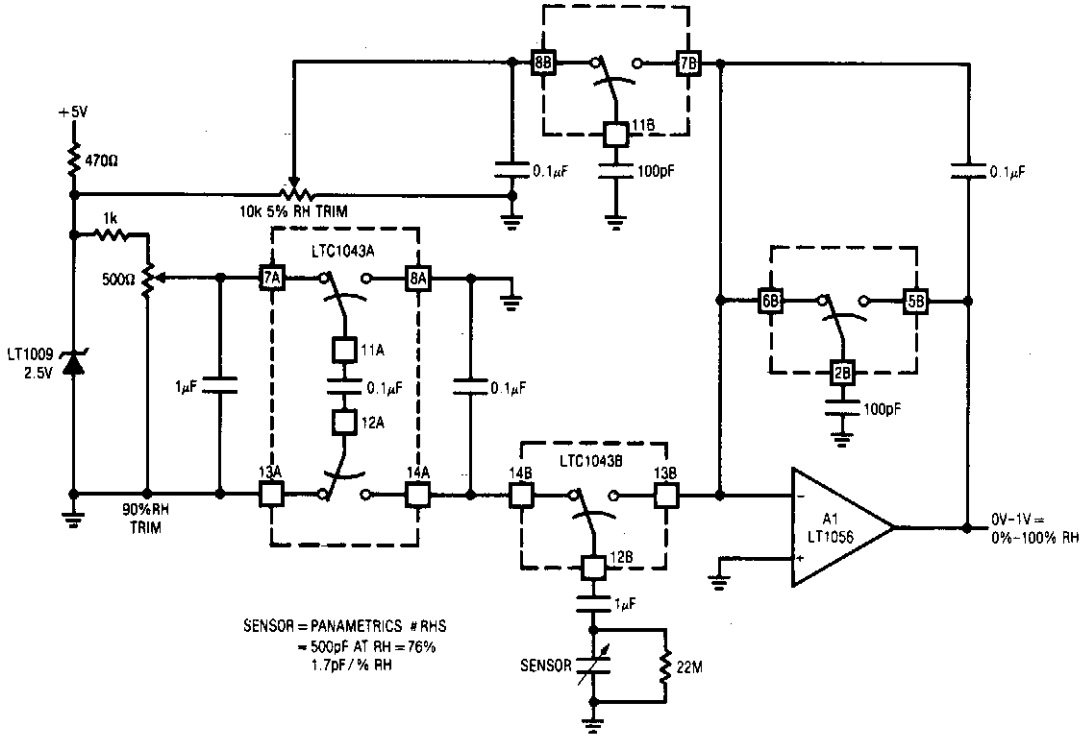
Door open alarms are used chiefly in automotive, industrial, and appliance applications. This type of circuit can sense the opening of a refrigerator door. When the door opens, a triac could be activated to control the inside light. The figure shows a door position alarm. When the door is opened, an LED turns on and the piezo alarm sounds for approximately 5 seconds. This circuit uses a TL3019 Hall-effect device for the door sensor. This normally open switch is located in the door frame. The magnet is mounted in the door. When the door is in the closed position, the TL3019 output goes to logic low, and remains low until the door is opened. This design consists of a TLC555 monostable timer circuit. The 1 μF capacitor and 5.1 M ohm resistor on pins 6 and 7 set the monostable RC time constant. These values allow the LED and piezo alarm to remain on about 5 seconds when triggered. One unusual aspect of this circuit is the method of triggering. Usually a 555 timer circuit is triggered by taking the trigger, pin 2, low which produces a high at the output, pin 3. In this configuration with the door in the closed position, the TL3019 output is held low. The trigger, pin 2, is connected to ½ the supply voltage V_{CC}. When the door opens, a positive high pulse is applied to control pin 5 through a 0.1 μF capacitor and also to reset pin 4. This starts the timing cycle. Both the piezo alarm and the LED visual indicator are activated.

44
Humidity Sensors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Relative Humidity Sensor Signal Conditioner

RELATIVE HUMIDITY SENSOR SIGNAL CONDITIONER

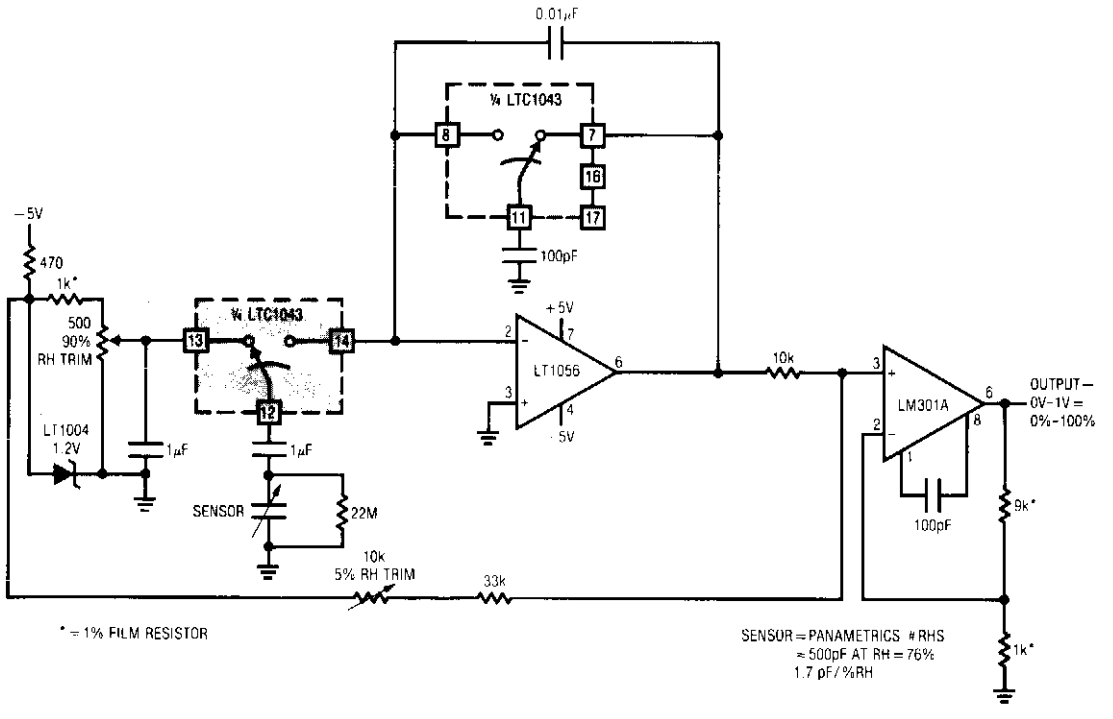


LINEAR TECHNOLOGY CORPORATION

Fig. 44-1

Circuit Notes

This circuit combines two LTC1043s with a based humidity transducer in a simple charge-pump based circuit. The sensor specified has a nominal 400 pF capacitance at RH = 76%, with a slope of 1.7 pF/% RH. The average voltage across this device must be zero. This provision prevents deleterious electrochemical migration in the sensor. The LTC1043A inverts a resistively scaled portion of the LT1009 reference, generating a negative potential at pin 14A. The LTC1043B alternately charges and discharges the humidity sensor via pins 12B, 13B, and 14B. With 14B and 12B connected, the sensor charges via the 1 μF unit to the negative potential at pin 14A. When the 14B-12B pair



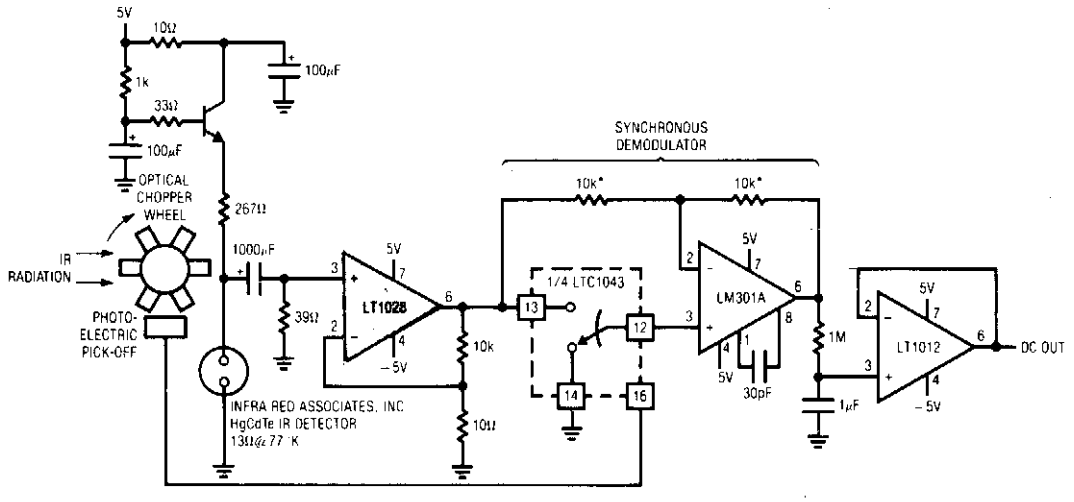
opens, 12B is connected to A1's summing point via 13B. The sensor now discharges into the summing point through the 1 μF capacitor. Since the charge voltage is fixed, the average current into the summing point is determined by the sensor's humidity related value. The 1 μF value ac couples the sensor to the charge-discharge path, maintaining the required zero average voltage across the device. The 22M resistor prevents accumulation of charge, which would stop current flow. The average current into A1's summing point is balanced by packets of charge delivered by the switched-capacitor gives A1 an integrator-like response, and its output is dc.

45
Infrared Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Low Noise Infrared Detector
Infrared Transmitter
Infrared Transmitter
Invisible Infrared Pulsed Laser Rifle
Infrared Receiver
Pulsed Infrared Diode Emitter Drive

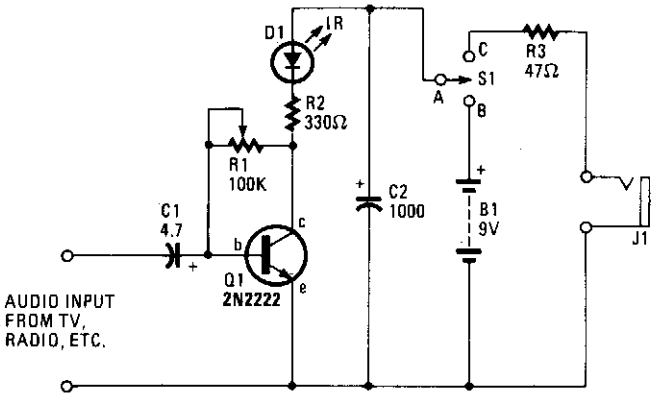
LOW NOISE INFRARED DETECTOR



LINEAR TECHNOLOGY CORP.

Fig. 45-1

INFRARED TRANSMITTER



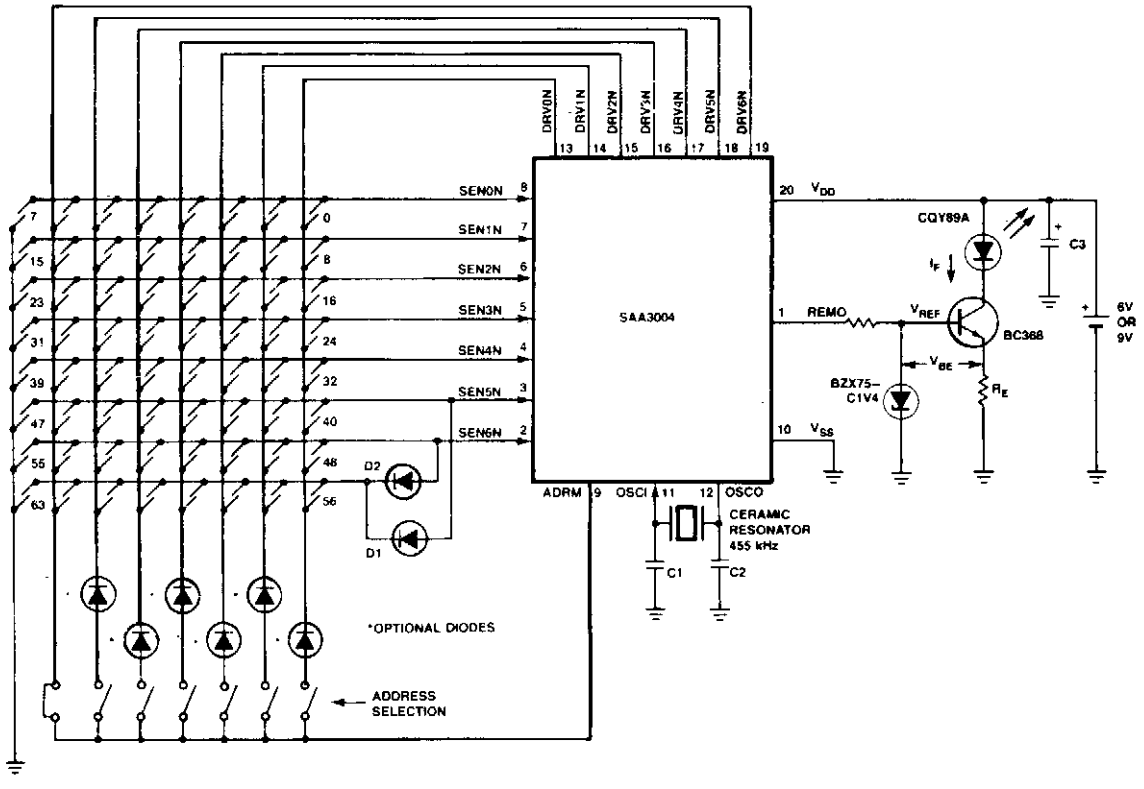
HANDS-ON ELECTRONICS

Fig. 45-2

Circuit Notes

The ultra-simple one-transistor, IR transmitter shown is designed to transmit the sound from any 8 or 16 ohm audio source, such as a TV, radio, or tape recorder on an infrared beam of light.

INFRARED TRANSMITTER



SIGNETICS

Fig. 45-3

Circuit Notes

The transmitter keyboard is arranged as a scanned matrix. The matrix consists of 7 driver outputs and 7 sense inputs. The driver outputs DRVON to DRV6N are open-drain n-channel transistors and they are conductive in the stand-by mode. The 7 sense inputs (SENON to SEN6N) enable the generation of 56 command codes. With 2 external diodes all 64 commands are addressable. The sense inputs have p-channel pull-up transistors, so that they are HIGH until they are pulled LOW by connecting them to an output via a key depression to initiate a code transmission.

INVISIBLE INFRARED PULSED LASER RIFLE

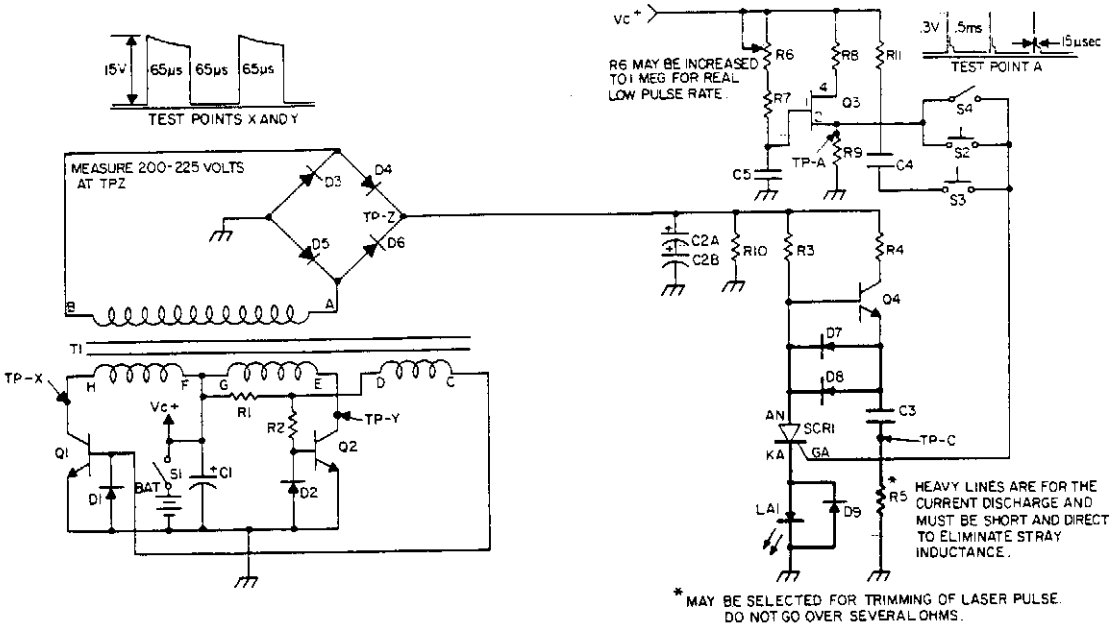


Fig. 45-4

TAB BOOKS INC.

Circuit Notes

The device generates an adjustable frequency of low to medium powered IR pulses of invisible energy and must be treated with care.

The portable battery pack is stepped up to 200 to 300 volts by the inverter circuit consisting of Q1, Q2, and T1. Q1 conducts until saturated, at which time, the base no longer can sustain it in an "on" state and Q1 turns "off," causing the magnetic field in its collector winding to collapse thus producing a voltage or proper phase in the base drive winding that turns on Q2 until saturated, repeating the above sequence of events in an "on/off" action. The diodes connected at the bases provide a return path for the base drive current. The stepped up squarewave voltage on the secondary of T1 is rectified and integrated on C2.

INFRARED RECEIVER

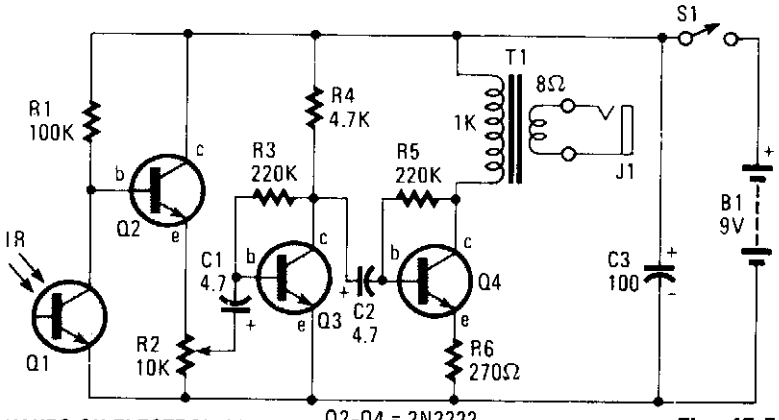


Fig. 45-5

Circuit Notes

The circuit consists of Q1—a phototransistor that responds to an intensity of amplitude-modulated IR light source—and a three-stage, high-gain audio amplifier. Transformer T1 is used to match the output impedance of the receiver to today's popular low-impedance (low-Z) headphones; but if a set of 1000-2000 ohm, magnetic (not crystal), high-impedance (high-Z) phones are to be used, remove T1 and connect the high-Z phones in place of T1's primary winding—the 1000-ohm winding.

PULSED INFRARED DIODE EMITTER DRIVE

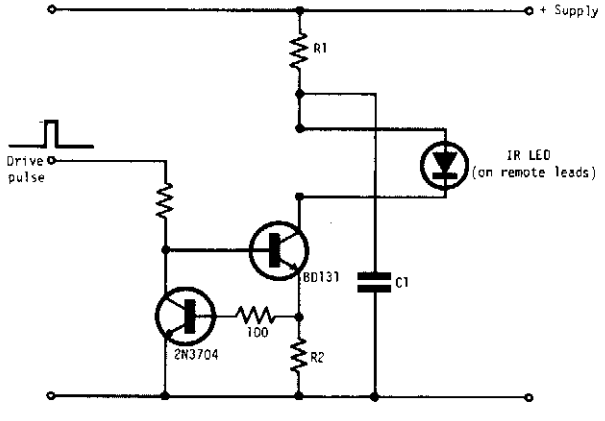


Fig. 45-6

Circuit Notes

Q1 and Q2 form a constant current drive defined by R2. (I approximates to the reciprocal of R2 in the circuit shown for values of I greater than 1 amp). The pulse current is drawn from C1 which is recharged during the time between pulses via R1. The value of C1 is determined from the duration and magnitude of the peak current required, and the time constant R1 C1 is determined from the duration between pulses.

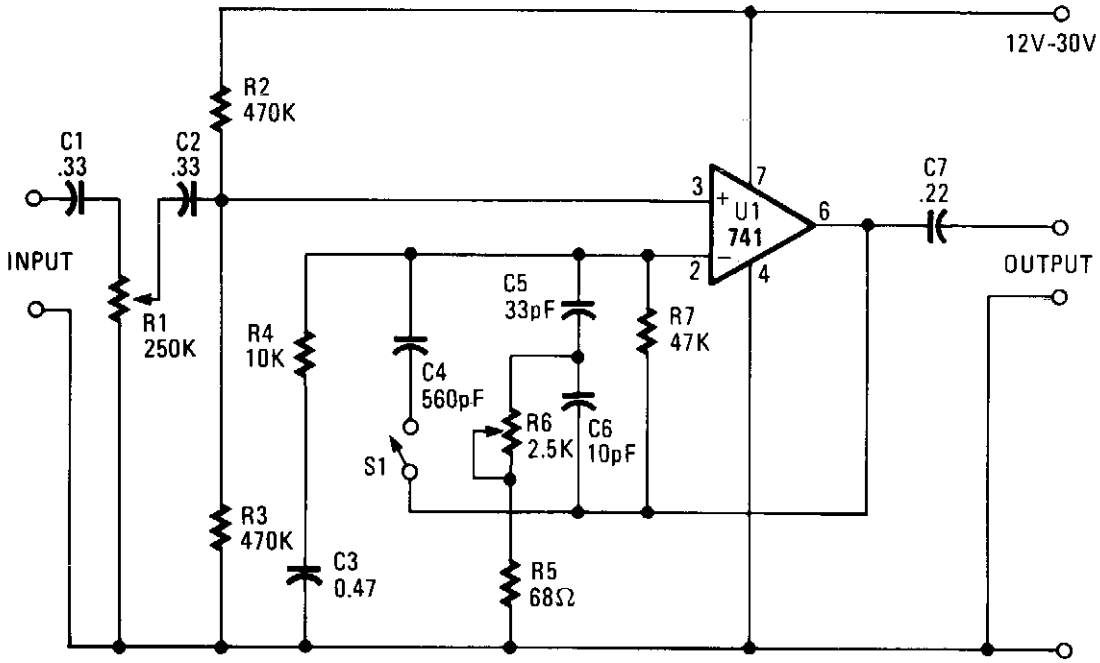
46

Instrumentation Amplifiers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Instrument Preamp
Instrumentation Amplifier with ±100 Volt Common Mode Range
Current-Collector Head-Amplifier
Instrumentation Meter Driver
Saturated Standard Cell Amplifier

INSTRUMENT PREAMP



HANDS-ON ELECTRONICS

Fig. 46-1

Circuit Notes

The input impedance is the value of potentiometer R1. If your instrument has extra-deep bass, change capacitor C1 to 0.5 μF. What appears to be an extra part in the feedback loop is a brightening tone control. The basic feedback from the op amp's output (pin 6) to the inverting input (pin 2) consists of resistor R7, and the series connection of resistor R4 and capacitor C3, which produce a voltage gain of almost 5 (almost 14 dB). That should be more extra oomph than usually needed. If the circuit is somewhat short on bass response, increase the value of capacitor C3 to 1 to 10 μF. Start with 1 μF and increase the value until you get the bass effect you want.

INSTRUMENTATION AMPLIFIER WITH ±100 VOLT COMMON MODE RANGE

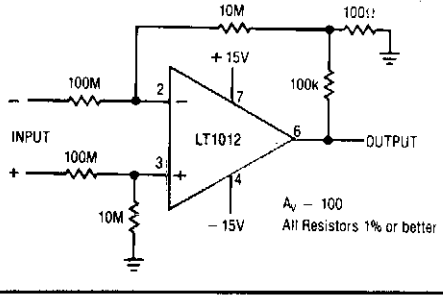
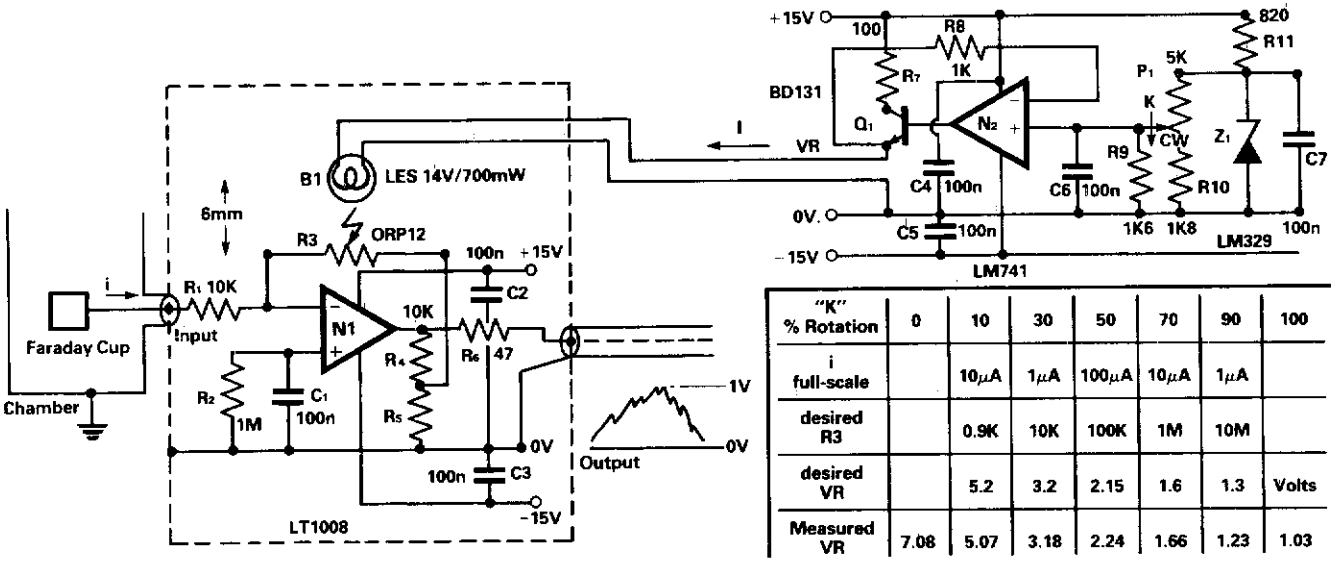


Fig. 46-2

LINEAR TECHNOLOGY CORP.

CURRENT-COLLECTOR HEAD-AMPLIFIER



"K" % Rotation	0	10	30	50	70	90	100	
i full-scale		10μA	1μA	100μA	10μA	1μA		
desired R3		0.9K	10K	100K	1M	10M		
desired VR		5.2	3.2	2.15	1.6	1.3		Volts
Measured VR	7.08	5.07	3.18	2.24	1.66	1.23	1.03	

ELECTRONIC ENGINEERING

Fig. 46-3

Circuit Notes

To amplify small current signals such as from an electron-collector inside a vacuum chamber, it is convenient for reasons of noise and bandwidth to have a "head-amplifier" attached to the chamber. The op amp N1 is a precision bipolar device with extremely low bias current and offset voltage (1) as well as low noise, which allows the 100:1 feedback attenuator R4:R5. The resistance of R3 can be varied from above 10M to below 1R, and so the nominal 0 to 1 V-peak output signal corresponds to input current ranges of 1 nA to 10 μA; this current i enters via the protective resistor R1. Light from the bulb B1 shines on R3, and the filament current I is controlled by the op amp N2.

The reference voltage VR is "shaped" by the resistors R9R10 so as to tailor the bulb and LDR characteristics to the desired current ranges. Thus, rotation of the calibrated knob K gives the appropriate resistance to R3 for the peak-current scale shown.

INSTRUMENTATION METER DRIVER

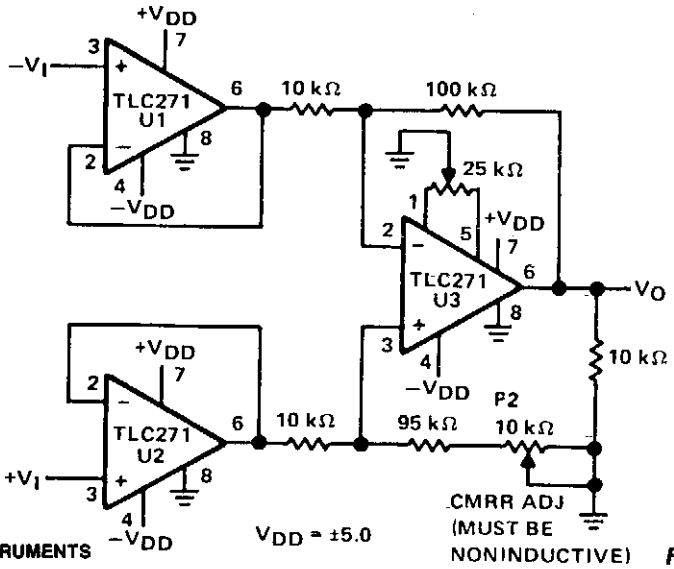


Fig. 46-4

Circuit Notes

Three op amps U1, U2, and U3 are connected in the basic instrumentation amplifier configuration. Operating from ± 5 V, pin 8 of each op amp is connected directly to ground and provides the ac performance desired in this application (high bias mode). P1 is for offset error correction and P2 allows adjustment of the input common mode rejection ratio. The high input impedance allows megohms without loading. The resulting circuit frequency response is 200 kHz at -3 dB and has a slew rate of 4.5 V/μs.

SATURATED STANDARD CELL AMPLIFIER

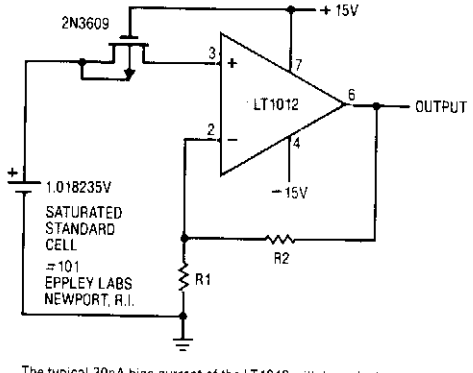


Fig. 46-5

The typical 30pA bias current of the LT1012 will degrade the standard cell by only 1 ppm/year. Noise is a fraction of a ppm. Unprotected gate MOSFET isolates standard cell on power down.

LINEAR TECHNOLOGY CORP.

47

Integrator Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Improved Non-Inverting Integrator
Active Integrator with Inverting Buffer
Long Time Integrator

IMPROVED NON-INVERTING INTEGRATOR

Fig. 1

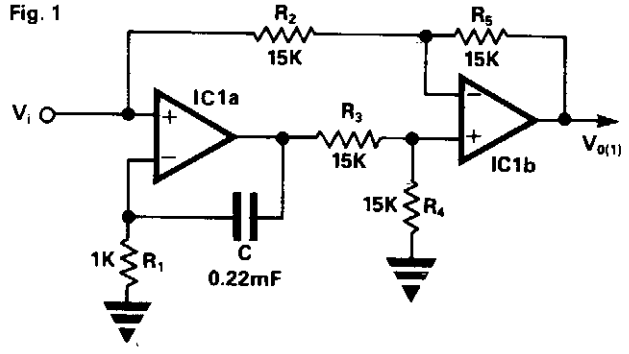


Fig. 2

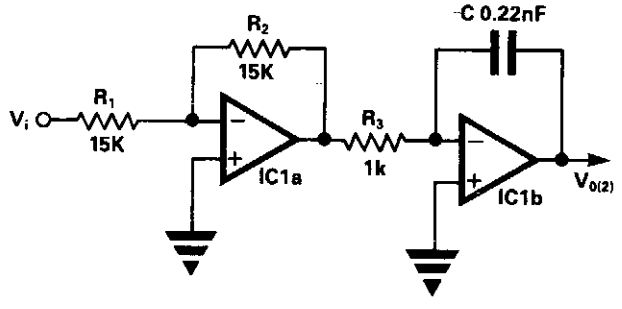


Fig. 3

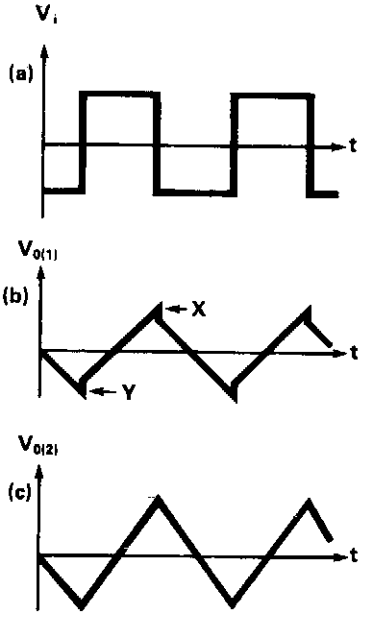


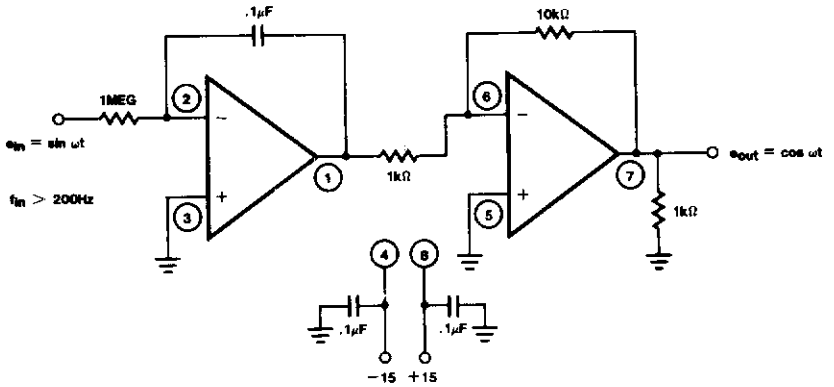
Fig. 47-1

Circuit Notes

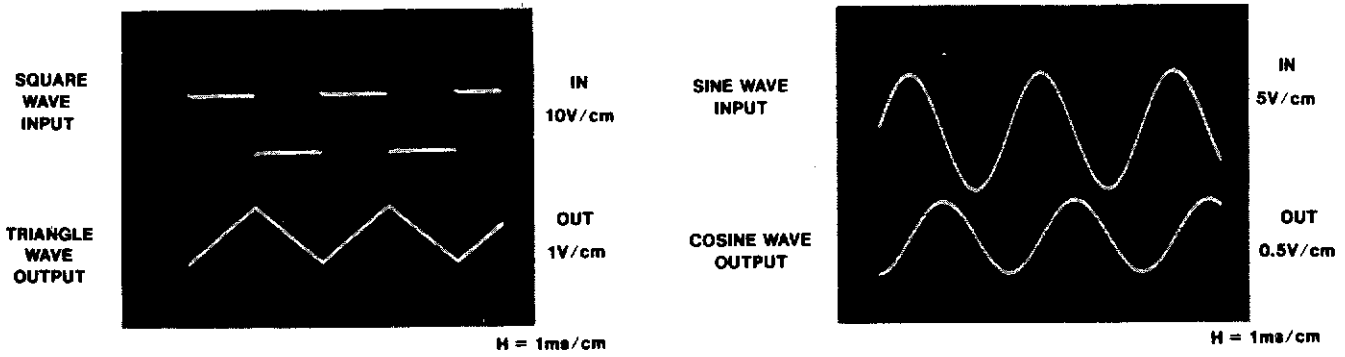
In the circuit in Fig. 1, IC1a produces the integral term required but also has the side effect of producing a proportional term not required, so this term is subtracted by IC1b leaving a pure integral. If the ratio R2/R5 does not exactly match the ratio of R3/R4, the subtraction will not be complete and a small amount of the proportional term will reach the output. The result of this with a squarewave input is shown in Fig. 3a as small steps in the output waveform at points X and Y.

This effect can be completely removed by using the simplified circuit shown in Fig. 2. Here the signal is pre-inverted by IC1a, then fed to a standard inverting integrator IC1b. The result is a non-inverting integrator with the advantage that the unwanted proportional term is never produced, so it does not need to be subtracted.

ACTIVE INTEGRATOR WITH INVERTING BUFFER



INTEGRATOR WAVEFORMS



IC = NE/SE5512

SIGNETICS

Fig. 47-2

LONG TIME INTEGRATOR

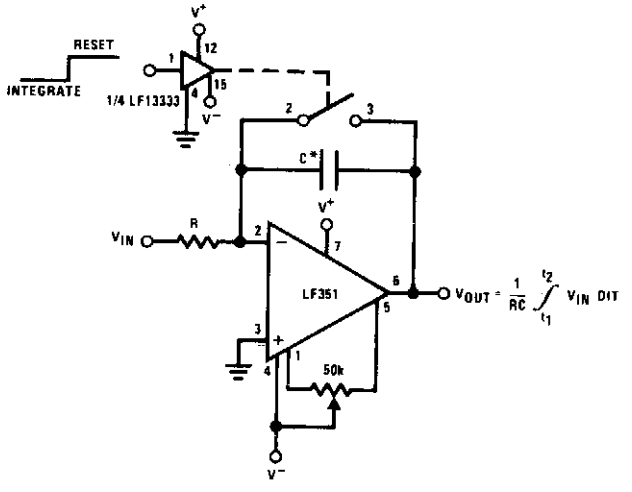


Fig. 47-3

* Low leakage capacitor
• 50k pot used for less sensitive V_{OS} adjust

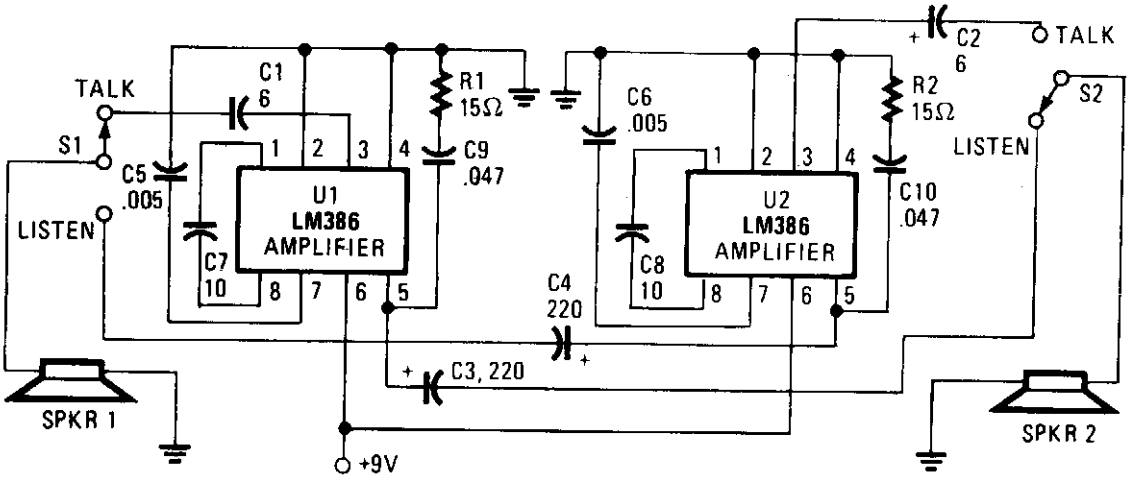
NATIONAL SEMICONDUCTOR CORP.

48
Intercom Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Intercom
Party-Line Intercom

INTERCOM



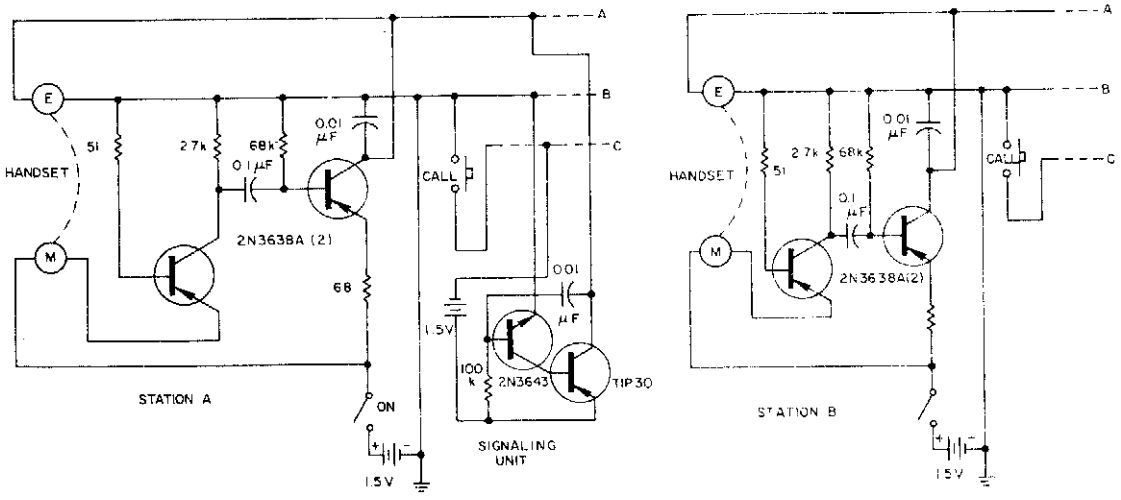
HANDS-ON ELECTRONICS

Fig. 48-1

Circuit Notes

The circuit consists of separate amplifiers—one for each station—rather than a single amplifier and a time sharing arrangement. U1 and U2 are low-voltage audio amplifiers, each of which operates as separate entities with switches at either station controlling which will transmit or receive. With capacitors C7 and C8 included in the circuit, the amplifiers have a gain of 200. Omitting those two components drops the gain to about 20. Other gain levels are available with the addition of a series-connected R/C combination connected between pin 1 and pin 8—for example, a 1000 ohm resistor and 10 μF capacitor for a gain of about 150.

PARTY-LINE INTERCOM



ELECTRONIC DESIGN

Fig. 48-2

Circuit Notes

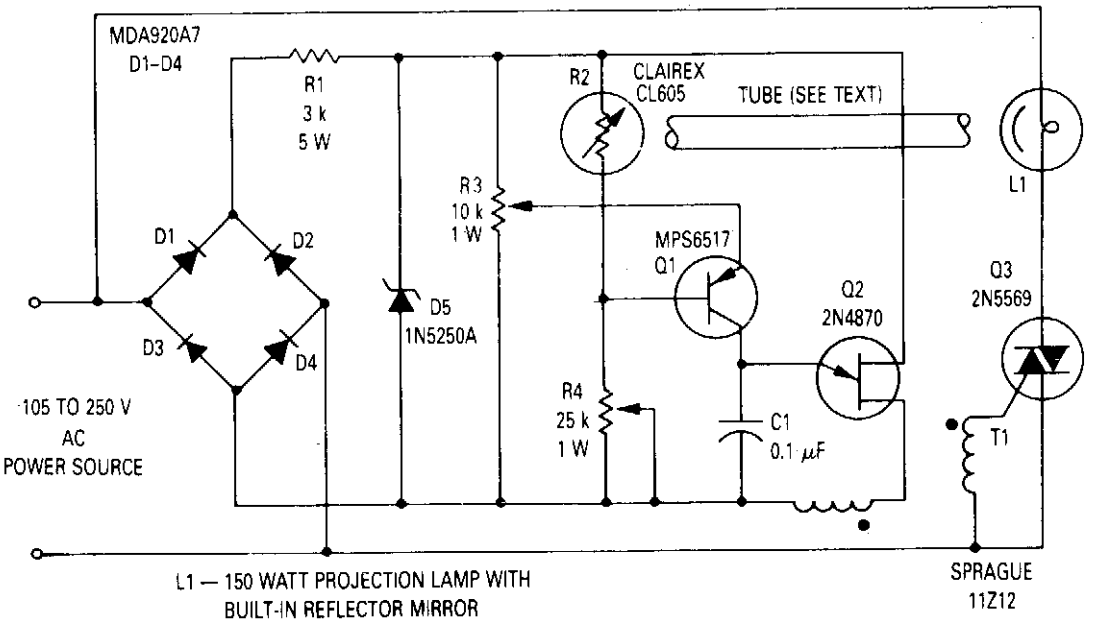
A large number of intercom stations can be tied together. All units are connected in parallel, and the entire system is buzzed by only one signaling circuit. Each unit is powered individually from 1.5-V cells for redundancy. For greater signal volume, 3-V sources can be used for the supplies without changing any other parts of the system. The carbon microphone of a standard telephone handset at each station feeds into a common-base amplifier, and a tandem high-gain common-emitter stage drives the intercom line. All phone earpieces are in parallel across the line. The signaling circuit, also connected across the line, is a simple oscillator that drives all the earpieces.

49
Lamp-Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Voltage Regulator for a Projection Lamp
Machine Vision Illumination Stabilizer
dc Lamp Dimmer
Automatic Light Controller for Carport
800 W Light-Dimmer
Lamp Dimmer
Rugged Lamp Driver is Short-Circuit Proof
TRIAC Lamp Dimmer
TRIAC Zero-Point Switch
Tandem Dimmer (Cross-Fader)

VOLTAGE REGULATOR FOR A PROJECTION LAMP



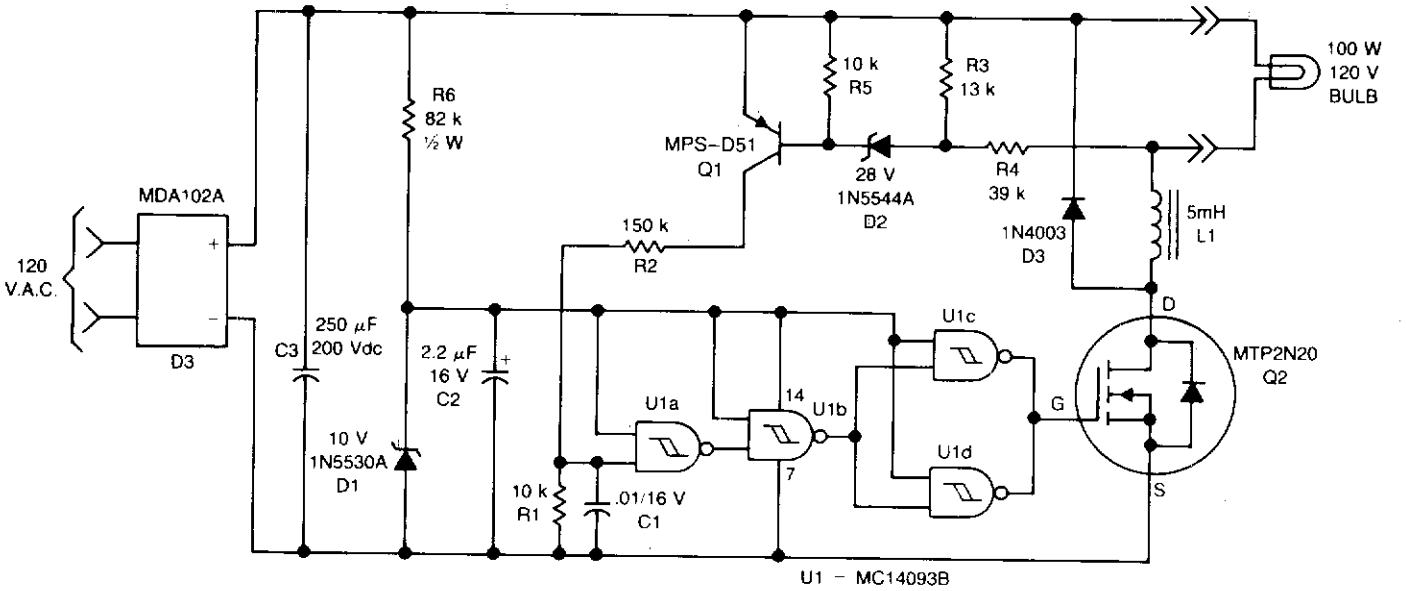
MOTOROLA

Fig. 49-1

Circuit Notes

The circuit will regulate the rms output voltage across the load (a projection lamp) to 100 volts ±2% for an input voltage between 105 and 250 volts ac. This is accomplished by indirectly sensing the light output of lamp L1 and applying this feedback signal to the firing circuit (Q1 and Q2) which controls the conduction angle of TRIAC Q3. The lamp voltage is provided by TRIAC Q3, whose conduction angle is set by the firing circuit for unijunction transistor Q2. The circuit is synchronized with the line through the full-wave bridge rectifier. The voltage to the firing circuit is limited by zener diode D5. Phase control of the supply voltage is set by the charging rate of capacitor C1. Q2 will fire when the voltage on C1 reaches approximately 0.65 times the zener voltage. The charging rate of C1 is set by the conduction of Q1, which is controlled by the resistance of photocell R2. Potentiometers R3 and R4 are used to set the lamp voltage to 100 volts when the line voltage is 105 volts and 250 volts, respectively.

MACHINE VISION ILLUMINATION STABILIZER



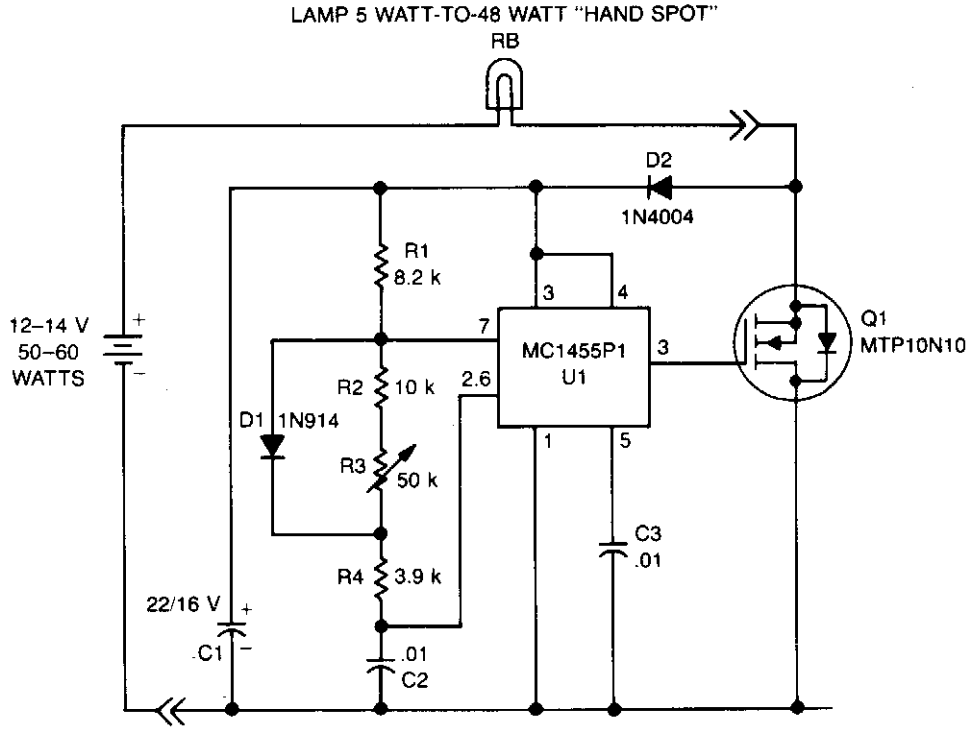
MOTOROLA

Fig. 49-2

Circuit Notes

The combination of Q1, Q2 and U1 form a hysteresis oscillator to stabilize lamp illumination. In operation, full wave bridge D3 operates directly from the ac line to supply unregulated dc to the lamp and also to the 10 V zener that provides power to the quad CMOS Schmitt trigger, U1. When the lamp supply exceeds 115 V, Q1 is turned ON, charging C1 through R2 to raise the input to U1a past the positive-going logic threshold. This drops the output voltage at U1c and U1d, which drives the gate of Q2, turning it OFF. Current then decays through the lamp, L1 and D3 until the lamp voltage falls below 115 V, at which time Q1 turns OFF and the cycle repeats.

DC LAMP DIMMER



MOTOROLA

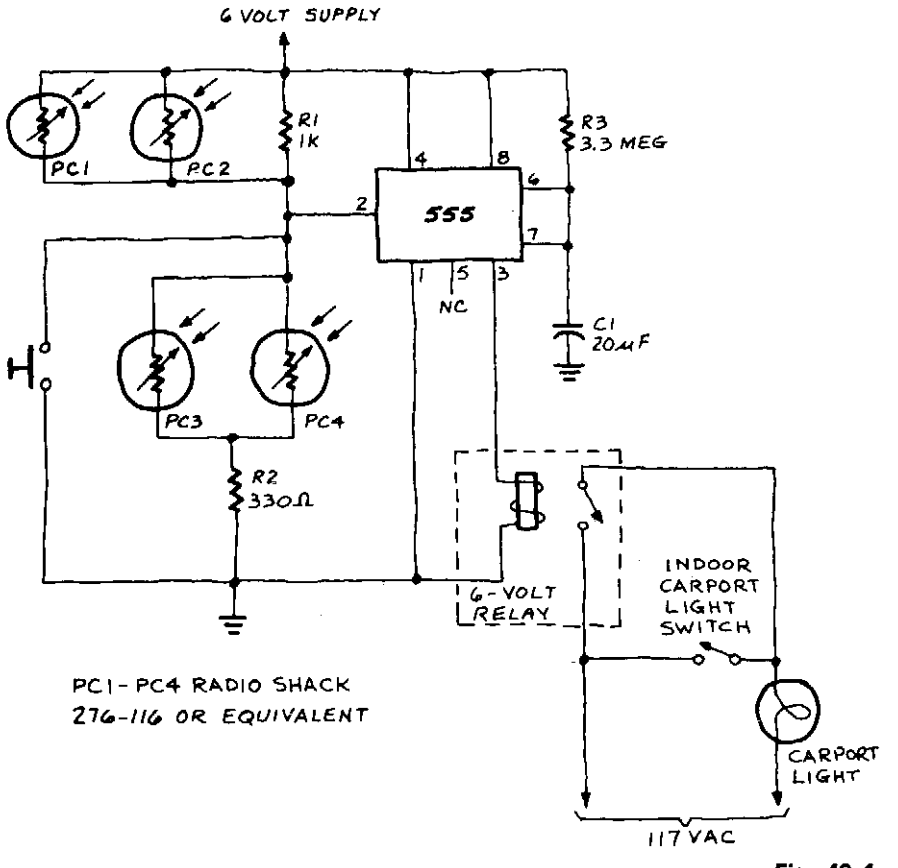
Fig. 49-3

Circuit Notes

A low power, low cost dc lamp dimmer for a two-wire portable "flashlight" can be realized with little or no heatsinking. In addition, a single potentiometer, R3 adjusts lamp brightness.

Battery power is stored in C1 for U1, which is a free-running multivibrator whose frequency is determined by R1, R2, R3, R4, and C2. U1 drives the gate of Q1, turning it and the lamp ON and OFF at a rate proportional to the multivibrator duty cycle.

AUTOMATIC LIGHT CONTROLLER FOR CARPORT



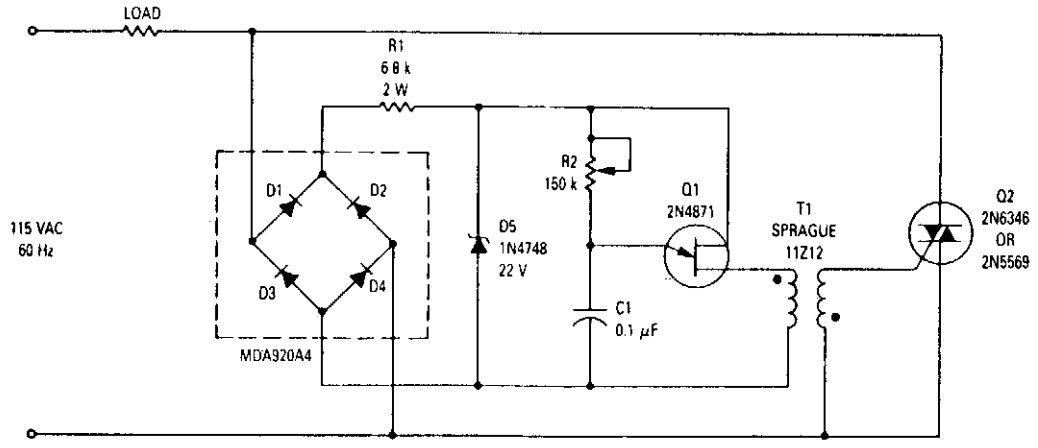
RADIO-ELECTRONICS

Fig. 49-4

Circuit Notes

A 555 timer IC, operating in the one-shot mode, is triggered by light striking photoresistors. These normally have a resistance of several megohms but, in the presence of light, that resistance drops to several hundred ohms, permitting current from the six-volt source to flow in the circuit. The R-C combination shown gives an on-time of about two minutes. Photoresistors PC3 and PC4 are mounted at headlight-height. When headlights illuminate the photoresistor, the timer starts. That actuates a relay, RY1, and the lights are turned on. The lights are automatically turned off when the timer's two minutes are up.

800 W LIGHT-DIMMER



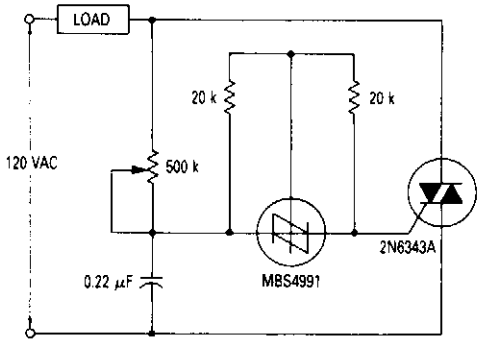
MOTOROLA

Fig. 49-5

Circuit Notes

This wide-range light dimmer circuit uses a unijunction transistor and a pulse transformer to provide phase control for the TRIAC. The circuit operates from a 115 volt, 60 Hz source and can control up to 800 watts of power to incandescent lights. The power to the lights is controlled by varying the conduction angle of the TRIAC from 0° to about 170°. The power available at 170° conduction is better than 97% of that at the full 180°.

LAMP DIMMER

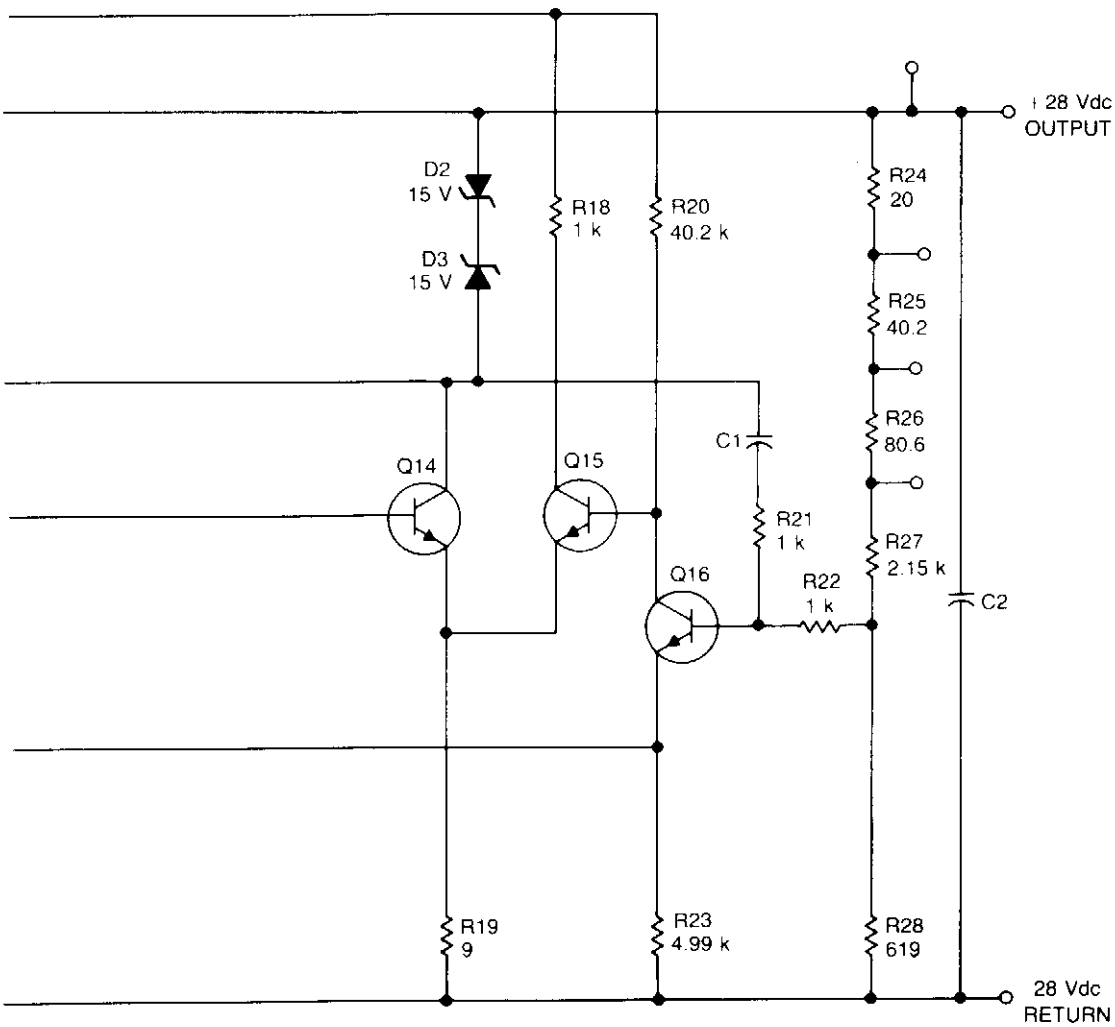


MOTOROLA

Fig. 49-6

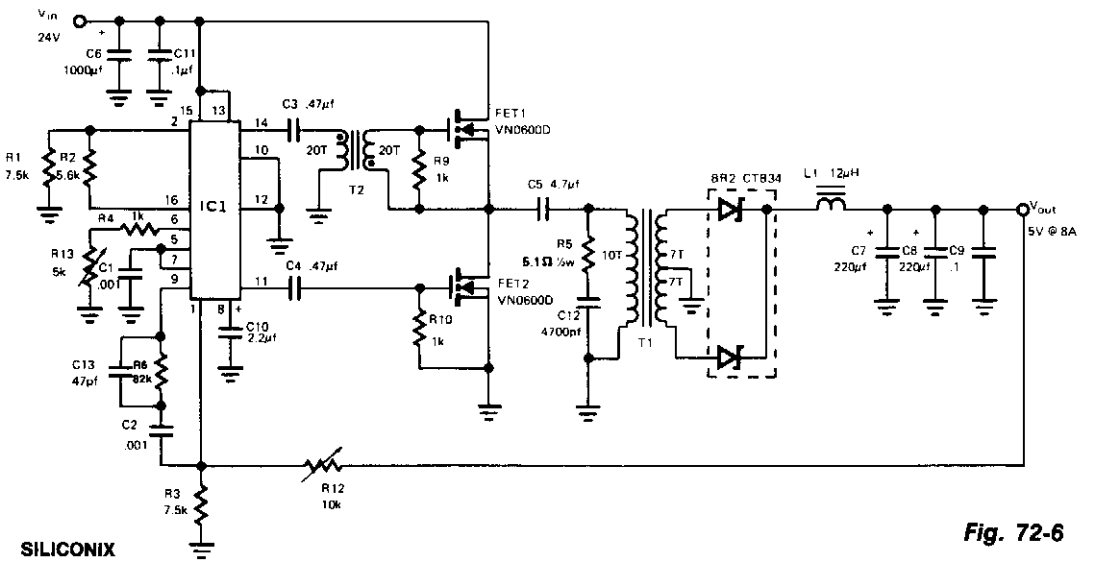
Circuit Notes

A full range power controller suitable for lamp dimming and similar applications operate from a 120 volt, 60 Hz ac source, and can control up to 1000 watts of power to incandescent bulbs. The power to the bulbs is varied by controlling the conduction angle of TRIAC Q1. At the end of each positive half-cycle when the applied voltage drops below that of the capacitor, gate current flows out of the SBS and it switches on, discharging the capacitor to near zero volts. The RC network shown across the TRIAC represents a typical snubber circuit that is normally adequate to prevent line transients from accidentally firing the TRIAC.



through Q16. The other input of the amplifier is connected to a radiation-hardened zener diode, D1. Local feedback using R21 and C1 produces gain to phase shift that are independent of individual component parameters, which provides stable operation into the required loads.

SWITCH MODE POWER SUPPLY



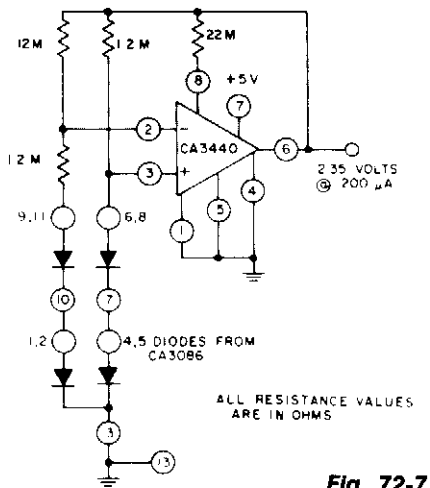
SILICONIX

Fig. 72-6

Circuit Notes

This buck-derived circuit provides up to 8 A at 5 Vdc operating off 24 to 32 Vdc. The two power MOSFETs in the circuit conduct alternately for equal periods. Switching frequency is 150 kHz, set by the PWM125 controller. Output of the two MOSFETs is transformed to a low-voltage level, then rectified. Efficiency of the circuit is 75% when operated in a 22 to 32 V range. Efficiency approaches 90% with higher voltage inputs.

MICROPOWER BANDGAP REFERENCE SUPPLY



GENERAL ELECTRIC/RCA

Fig. 72-7

Circuit Notes

The circuit uses a CA3440 BiMOS op amp and CA3086 transistor array. The no-load current from 5-volt supply is 1.5 µA. Load current can go as high as 200 µA and still maintain output voltage regulation within 0.05%.

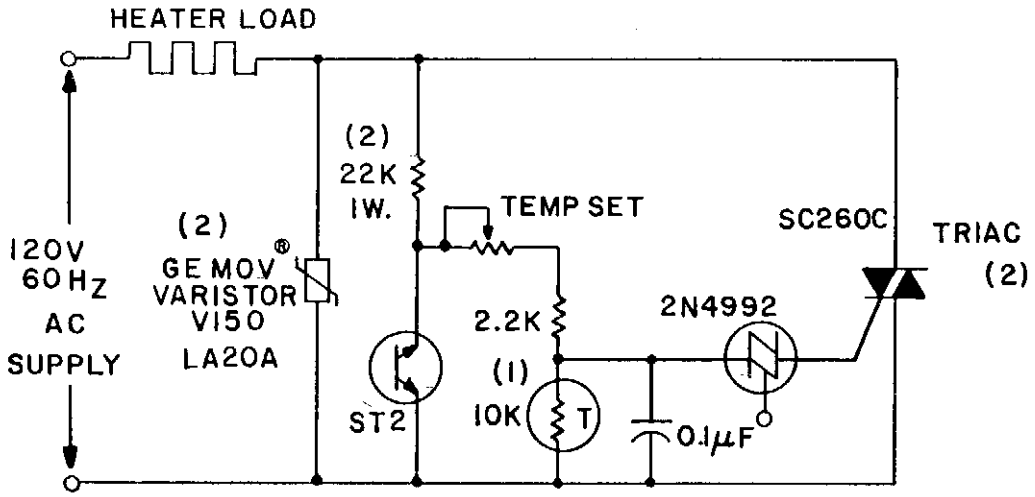
Circuit Notes

The "zero voltage switching" technique is widely used to modulate heating and similar types of ac loads where the time constant associated with the load (tens of seconds to minutes) is sufficiently long to allow smooth proportional modulation by time ratio control, using one complete cycle of the ac input voltage as the minimum switching movement. Despite its attractions, the traditional triac-based ZVS is virtually unusable for the control of very low power loads, especially from 220 volt ac inputs due to the triac's reluctance to latch-on into the near-zero instantaneous currents that flow through it and the load near the ac voltage zero crossover points. The circuit side-steps the latching problem by employing a pair of very sensitive low current reverse blocking thyristors (C106) connected in antiparallel; these are triggered by a simple thermistor modulated differential amplifier (Q1, Q2), with zero voltage logic furnished by an H11AA1 ac input optocoupler. With the NTC thermistor TH calling for heat, transistor Q1 is cut off and Q2 is on, which would normally provide continuous base drive to Q3, with consequent triggering of either SCR, or of SCR 2 via SCR1, depending on phasing of the ac input.

Note that when the ac input voltage is positive with respect to SCR 2, SCR 1 is reverse biased and, in the presence of "gate" current from Q3, behaves as a remote base transistor, whose output provides via blocking diode CR1, positive gate trigger current for SCR 2. When the ac input polarity is reversed (SCR 1's anode positive), SCR 1 behaves as a direct fired conventional thyristor. "Trigger" current to SCR 1, however, is not continuous, even when TH is calling for heat and Q2 is delivering base current to Q3. In this situation, Q3 is inhibited from conduction by the clamping action of PC1, an H11AA photocoupler, except during those brief instants when the ac input voltage is near zero and the coupler input diodes are deprived of current.

Triggering of either SCR can occur only at ac voltage crossing points, and RFI-less operation results. The proportional control feature is injected via the positive feedback action of capacitor CM, which converts the differential amplifier Q1, Q2 into a simple multivibrator, whose duty cycle varies from one to 99 percent according to the resistance of TH. Zener diode Z1 is operational, being preferred when maximum immunity from ac voltage induced temperature drift is desired.

HEATER ELEMENT TEMPERATURE CONTROLLER



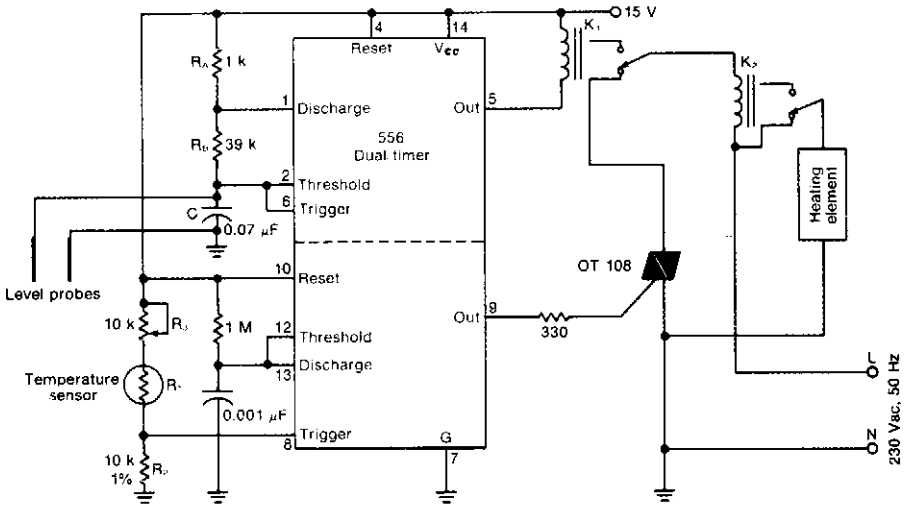
NOTES: 1. ThermistorNational Lead type 1D101, or equivalent.
2. Component values for 220V operation:
Resistor – 47K, 2W
GE-MOV® Varistor – V275LA20A
Triac – SC260E

GENERAL ELECTRIC *Fig. 99-6*

Circuit Notes

The circuit can control up to 6 kW of heating, with moderate gain, using a 25-amp triac (SC260D). Feedback is provided by the negative temperature co-efficient (NTC) thermistor, which is mounted adjacent to the environment being temperature controlled. The temperature set potentiometer is initially adjusted to the desired heating level. As the thermistor becomes heated by the load, its resistance drops, phasing back the conduction angle of the triac, so the load voltage is reduced. The ST2 diac is used as a back-to-back zener diode. Its negative resistance region in its E-I characteristic provides a degree of line voltage stabilization. As the input line voltage increases, the diac triggers earlier in the cycle and, hence, the average charging voltage to the 0.1 μF capacitor, decreases.

DUAL-TIMER CHIP CONTROLS TEMPERATURE WHILE MONITORING LIQUID LEVEL



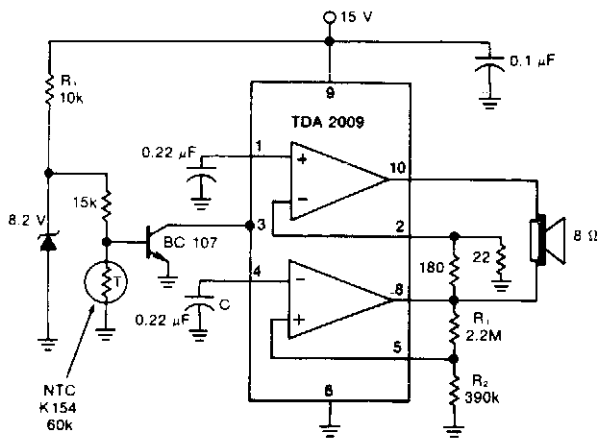
ELECTRONIC DESIGN

Fig. 99-7

Circuit Notes

One-half of a 556 dual timer monitors the temperature of a liquid bath, controlling a heating element that maintains temperature within ±2°C over a 32° - 200°C range. The other half monitors the liquid level, disconnecting the heater when the level drops below a preset point.

TEMPERATURE ALARM



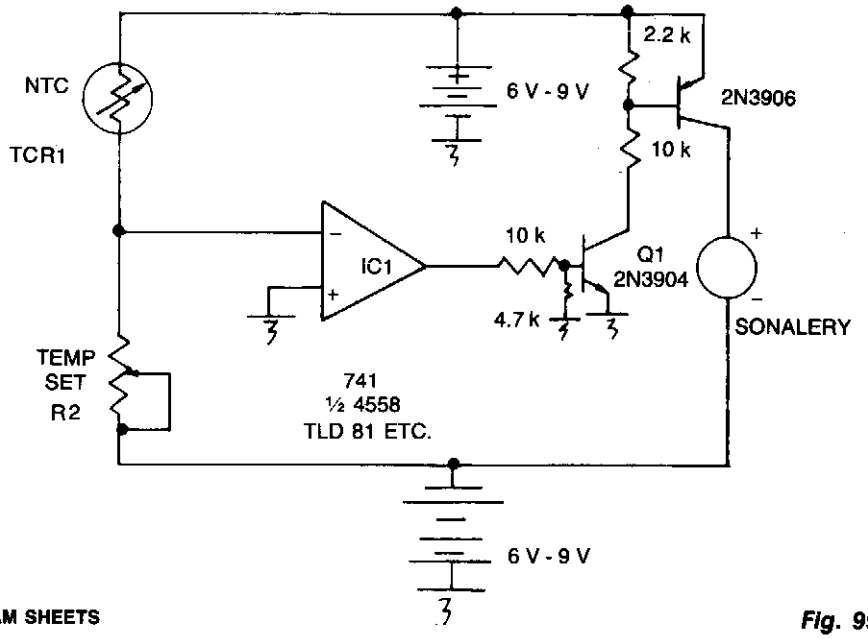
ELECTRONIC DESIGN

Fig. 99-8

Circuit Notes

The mute pin of this dual audio amplifier is used as the trigger for a one chip high-temperature alarm. One-half of the IC is connected as an oscillator and the other boosts the audio alarm outputs to 10W.

ADJUSTABLE THRESHOLD TEMPERATURE ALARM



WILLIAM SHEETS

Fig. 99-9

Circuit Notes

When R1 increases as temp decreases, the output of IC1 goes positive, turning on Q1. Q1 conducts and causes Q2 to conduct, turning on the audible alarm. The threshold is set with potentiometer R2.

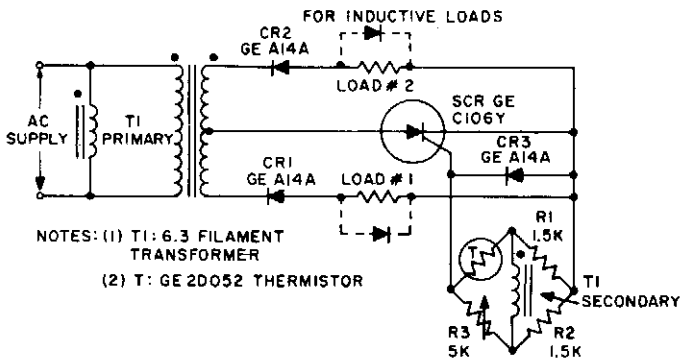
100

Temperature Sensors

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Dual Output, Over-Under Temperature Monitor
Temperature Sensor and DVM Interface
Curvature Corrected Platinum RD Thermometer
Thermocouple Amplifier with Cold Junction Compensation
5-V Powered, Linearized Platinum RTD Signal Conditioner
HI LO Temperature Sensor

DUAL OUTPUT, OVER-UNDER TEMPERATURE MONITOR



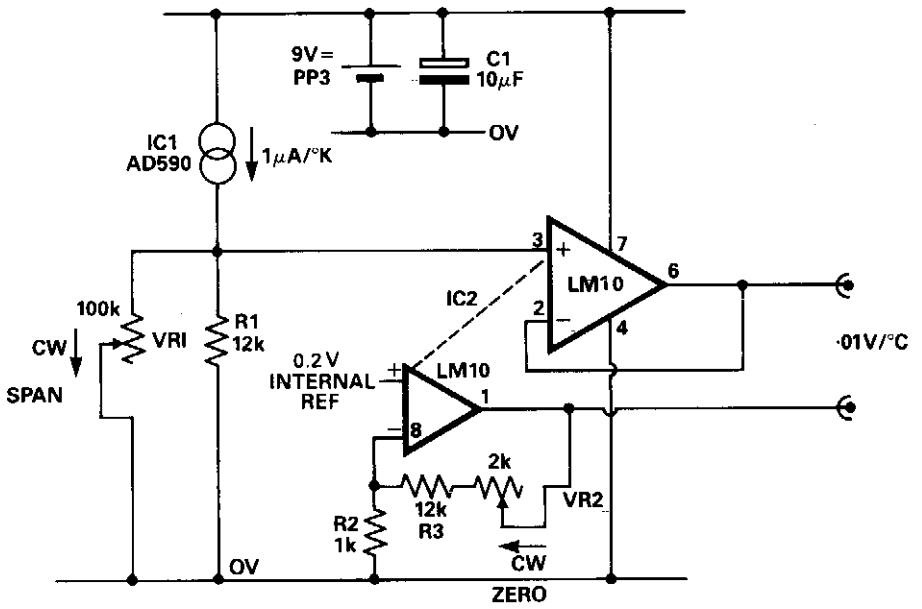
GENERAL ELECTRIC

Fig. 100-1

Circuit Notes

This circuit is ideal for use as an over-under temperature monitor, where its dual output feature can be used to drive HIGH and LOW temperature indicator lamps, relays, etc. T1 is a 6.3 volt filament transformer whose secondary winding is connected inside a four arm bridge. When the bridge is balanced, ac output is zero, and C5 (or C7) receives no gate signal. If the bridge is unbalanced by raising or lowering the thermistor's ambient temperature, and ac voltage will appear across the SCR's gate cathode terminals. Depending in which sense the bridge is unbalanced, the positive gate voltage will be in phase with, or 180° out of phase with the ac supply. If the positive gate voltage is in phase, the SCR will deliver load current through diode CR1 to load (1), diode CR2 blocking current to load (2). Conversely, if positive gate voltage is 180° out of phase, diode CR2 will conduct and deliver power to load (2), CR1 being reverse biased under these conditions. With the component values shown, the circuit will respond to changes in temperature of approximately 1-2°C. Substitution of other variable-resistance sensors, such as cadmium sulfide light dependent resistors (LDR) or strain gauge elements, for the thermistor shown is permissible.

TEMPERATURE SENSOR AND DVM INTERFACE



ELECTRONIC ENGINEERING

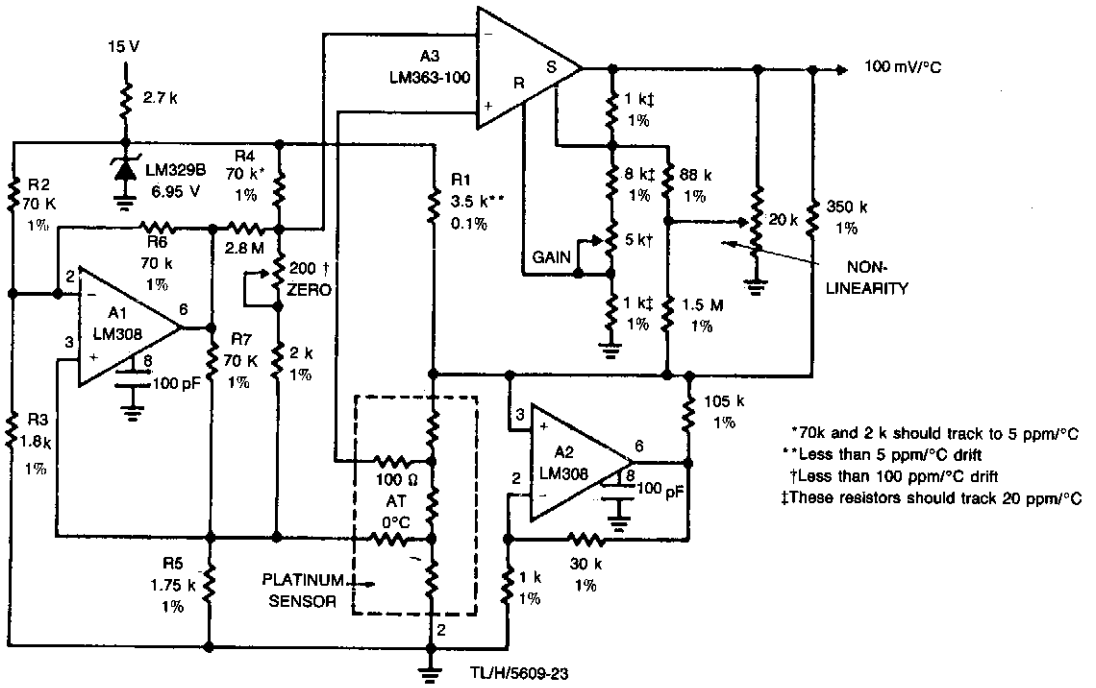
Fig. 100-2

Circuit Notes

The DVM gives a direct indication of the temperature of the sensor in degrees Centigrade. The temperature sensor IC1 gives a nominal 1 μA per degree Kelvin which is converted to 10 mV per degree Kelvin by R1 and VR1. IC2 is a micropower, low input drift op amp with internal voltage reference and amplifier. The main op amp in IC1 is connected as a voltage follower to buffer the sensor voltage at R1.

The second amplifier in IC1 is used to amplify the .2 V internal reference up to 2.73 V in order to offset the 273 degrees below 0°C. The output voltage of the unit is the differential output of the two op amps and is thus equal to 0.01 V per °C.

CURVATURE CORRECTED PLATINUM RTD THERMOMETER



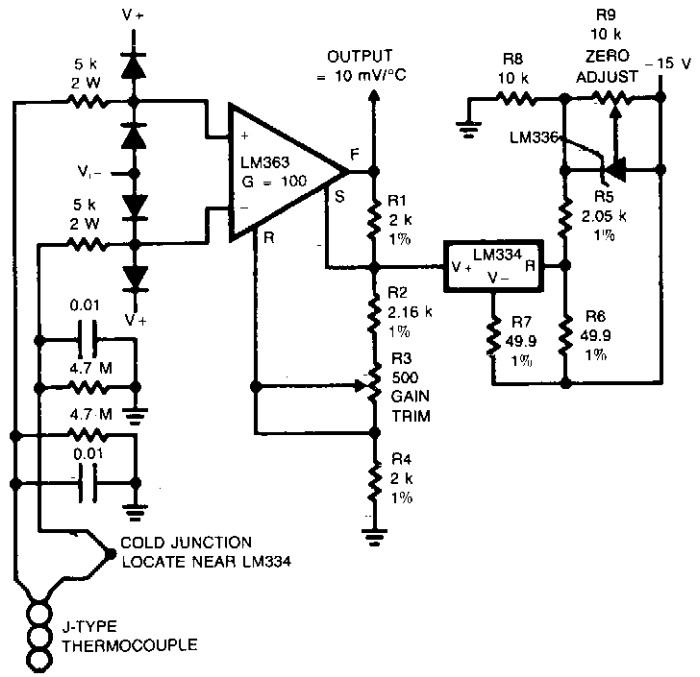
NATIONAL SEMICONDUCTOR CORP.

Fig. 100-3

Circuit Notes

This thermometer is capable of 0.01°C accuracy over −50°C to +150°C. A unique trim arrangement eliminates cumbersome trim interactions so that zero gain, and nonlinearity correction can be trimmed in one even trip. Extra op amps provide full Kelvin sensing on the sensor without adding drift and offset terms found in other designs. A1 is configured as a Howland current pump, biasing the sensor with a fixed current. Resistors R2, R3, R4 and R5 form a bridge driven into balance by A1. In balance, both inputs of A2 are at the same voltage. Since R6 = R7, A1 draws equal currents from both legs of the bridge. Any loading of the R4/R5 leg by the sensor would unbalance the bridge; therefore, both bridge taps are given to the sensor open circuit voltage and no current is drawn.

THERMOCOUPLE AMPLIFIER WITH COLD JUNCTION COMPENSATION



NATIONAL SEMICONDUCTOR CORP.

Fig. 100-4

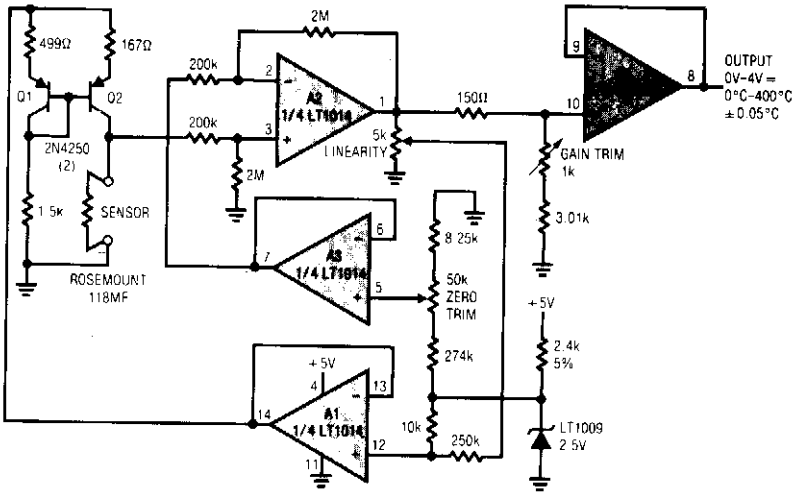
Circuit Notes

Input protection circuitry allows thermocouple to short to 120 Vac without damaging the amplifier.

Calibration:

1. Apply a 50 mV signal in place of the thermocouple. Trim R3 for V_{OUT} = 12.25 V.
2. Reconnect the thermocouple. Trim R9 for correct output.

5-V POWERED, LINEARIZED PLATINUM RTD SIGNAL CONDITIONER

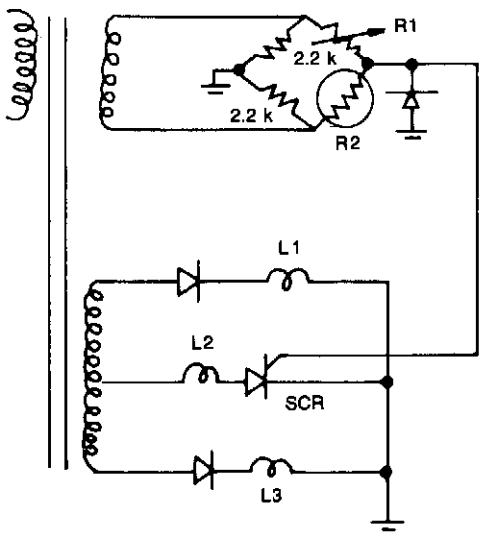


ALL RESISTORS ARE TRW-MAR-6 METAL FILM.
RATIO MATCH 2M 200K ±0.01%.
TRIM SEQUENCE
SET SENSOR TO 0 VALUE
ADJUST ZERO FOR 0V OUT

LINEAR TECHNOLOGY CORP.

Fig. 100-5

HI LO TEMPERATURE SENSOR



WILLIAM SHEETS

Fig. 100-6

Circuit Notes

Resistors R1, R2, and the two 2.2 k resistors form a bridge circuit. R2 is a thermistor, and R1 sets the temperature at which L2 lights. Lower or higher temperatures light L1 or L3 to indicate an over- or under-temperature condition.

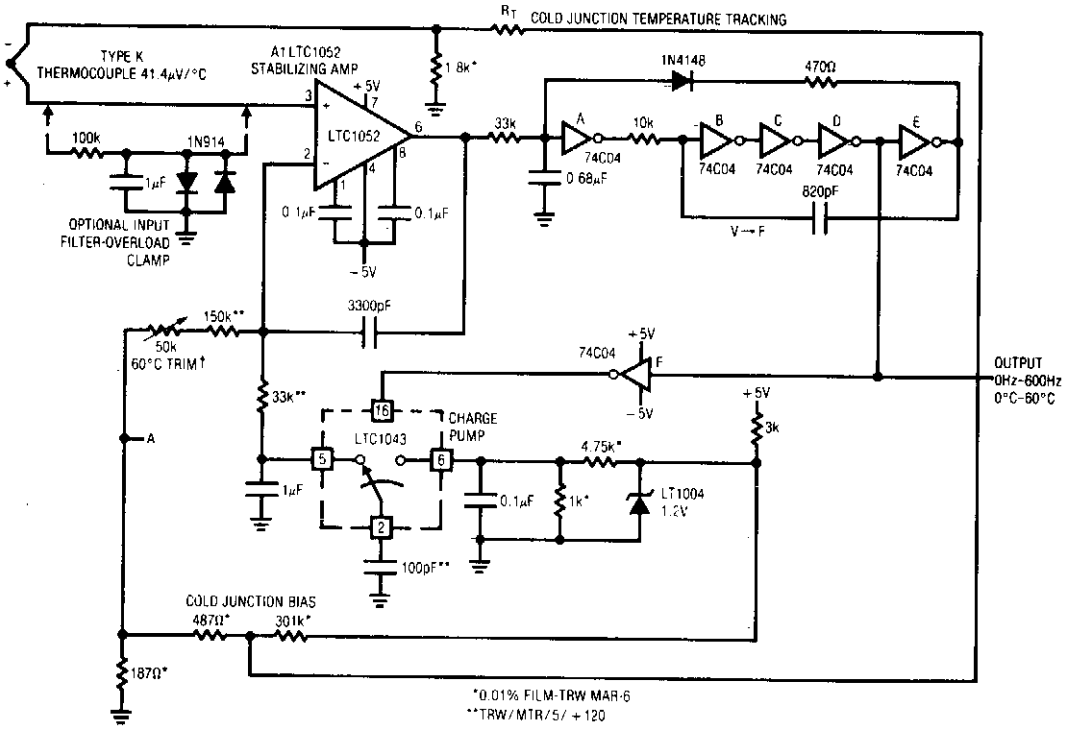
101

Temperature-to-Frequency Converters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Temperature-to-Frequency Converter
Digital Temperature Measuring Circuit

TEMPERATURE-TO-FREQUENCY CONVERTER



*0.01% FILM-TRW MAR-6
**TRW/MTR/5/ +120
R_T = YELLOW SPRINGS INST. #44007
100pF = POLYSTYRENE
†FOR GENERAL PURPOSE (1mV FULL-SCALE) 10-BIT A TO D. REMOVE THERMOCOUPLE–COLD JUNCTION NETWORK, GROUND POINT A, AND DRIVE LTC1052 POSITIVE INPUT.

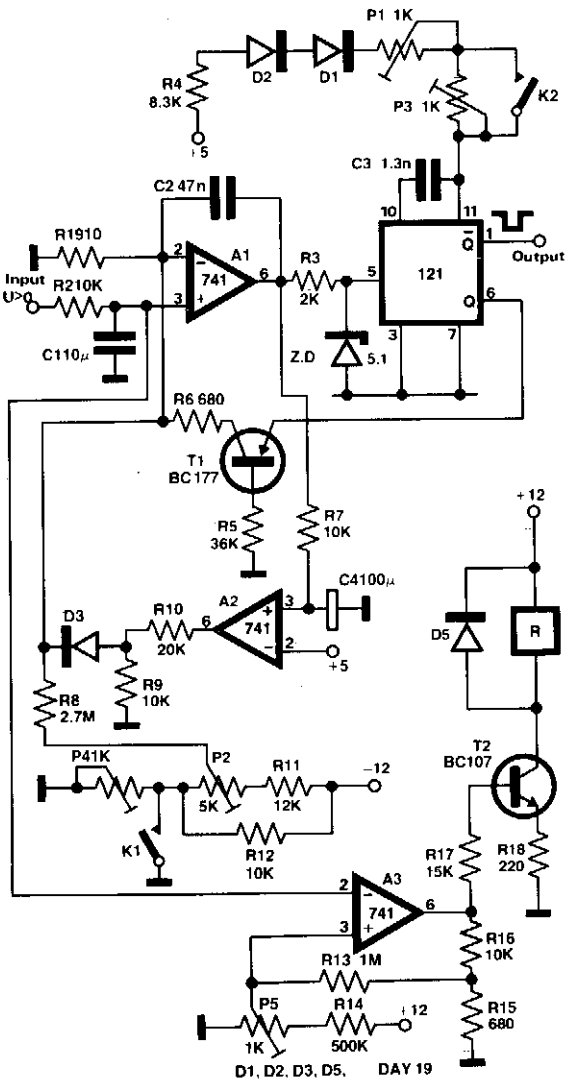
LINEAR TECHNOLOGY CORPORATION

Fig. 101-1

Circuit Notes

A1's positive input is biased by the thermocouple. A1's output drives a crude V → F converter, comprised of the 74C04 inverters and associated components. Each V → F output pulse causes a fixed quantity of charge to be dispensed into the 1 μF capacitor from the 100 pF capacitor via the LT1043 switch. The larger capacitor integrates the packets of charge, producing a dc voltage at A1's negative input. A1's output forces the V– F converter to run at whatever frequency is required to balance the amplifier's inputs. This feedback action eliminates drift and nonlinearities in the V → F converter as an error item and the output frequency is solely a function of the dc conditions at A1's inputs. The 3300 pF capacitor forms a dominant response pole at A1, stabilizing the loop.

DIGITAL TEMPERATURE MEASURING CIRCUIT



Circuit Notes

The output voltage of a thermocouple is converted into frequency measured by a digital frequency meter. The measuring set connected with Ni-NiCr thermocouple permits you to measure the temperatures within the range of 5°C - 800°C with ±1°C error. The output thermocouple signal is proportional to the temperature difference between the hot junction and the thermostat kept at 0°C, it drives the voltage-to-frequency converter changing the analogue input signal into the output frequency with the conversion ratio adjusted in such a way, that the frequency is equal to the measured temperature in Celsius degrees, e.g., for 350°C the frequency value is 350 Hz.

ELECTRONIC ENGINEERING

Fig. 101-2

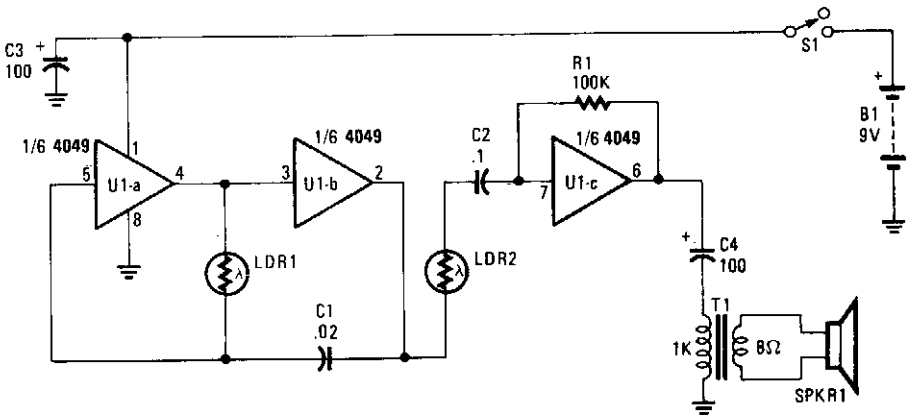
102

Theremins

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Electronic Theremin
Digital Theremin

ELECTRONIC THEREMIN



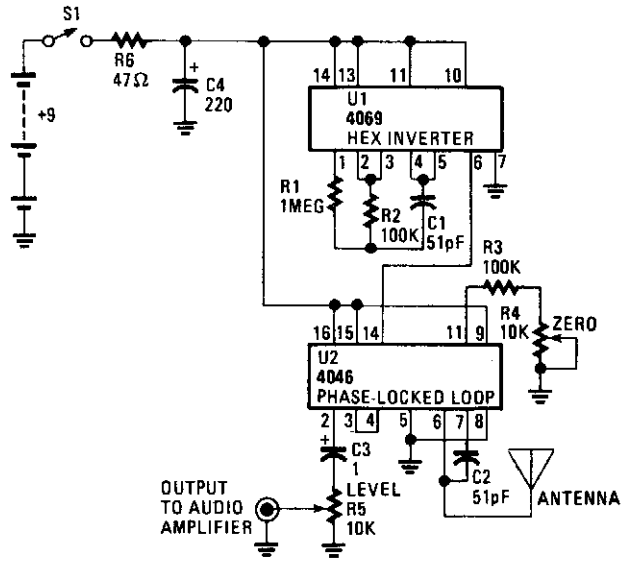
HANDS-ON ELECTRONICS

Fig. 102-1

Circuit Notes

This circuit has the CMOS IC doing double-duty performance. The first two inverters operate as a digital audio oscillator; the third operates as a low-gain linear audio amplifier. As the intensity of the light falling on photoresistor LDR1 increases the oscillator's frequency increases; similarly, the illumination falling on photoresistor LDR2 determines the volume level from the loudspeaker: The more illumination the more volume. If you flop and wave your hands between the two photocells and a light source, a special kind of electronic music will be produced.

DIGITAL THEREMIN



HANDS-ON ELECTRONICS

Fig. 102-2

Circuit Notes

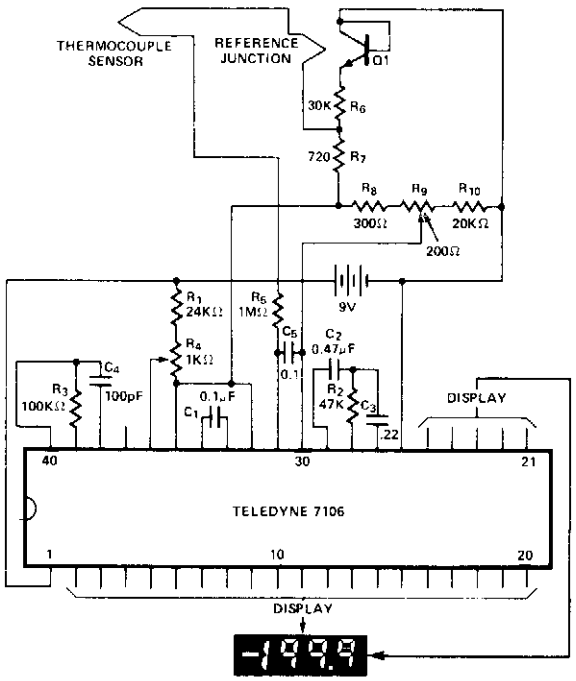
The CD4069 or 74C04 hex inverter—is used as a fixed-frequency oscillator centered around 100 kHz. U2 contains the variable frequency oscillator and balanced modulator. The CD4046 is a phase-locked loop and R3, R4, and C2 determine the center frequency of the on-chip oscillator. The antenna forms a parallel capacitance with C2, which allows the frequency to be shifted several kilohertz by bringing a hand near the antenna. R4, the ZERO control, allows the variable oscillator to be set to the same frequency as the fixed oscillator. When the difference frequency is below 15 Hz, it is below the lower frequency limit of the ear. By setting both oscillators to the same frequency, the Theremin remains silent until the performer brings his or her hand near the antenna. The oscillators are mixed by an exclusive OR gate inside the 4046. That gate acts as a digital balanced modulator, which produces the sum and difference frequencies. The output of the gate is then ac coupled by C3 to LEVEL control R5 and an output jack for connection to an audio amplifier or stereo receiver.

103
Thermometer Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Digital Thermocouple Thermometer
Remote Thermometer
Electronic Thermometer
Differential Thermometer
Basic Digital Thermometer, Kelvin Scale with Zero Adjust
Centigrade Thermometer (0°C-100°C)

DIGITAL THERMOCOUPLE THERMOMETER



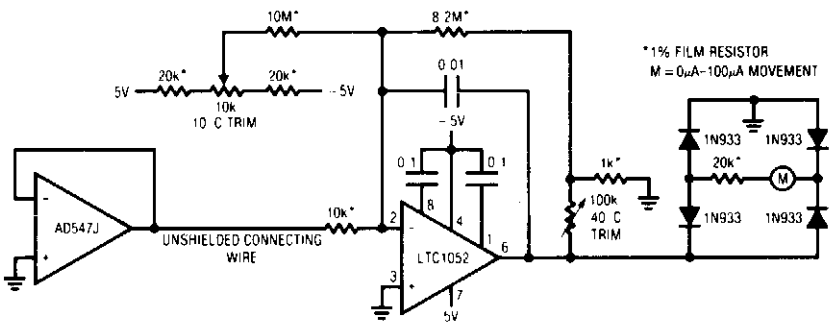
TELEDYNE SEMICONDUCTOR

Fig. 103-1

Circuit Notes

This digital thermocouple thermometer uses one active component and 15 passive components. With this circuit, both type J and type K thermocouples may be used. The type J will measure over the temperature range of 10 to 530°C with a conformity of ±2°C. The type K will measure over a temperature range of 0°C to 1000°C with a conformity of ±3°C.

REMOTE THERMOMETER



LINEAR TECHNOLOGY

Fig. 103-2

Circuit Notes

The low output impedance of a closed loop op amp gives ideal line-noise immunity, while the op amp's offset voltage drift provides a temperature sensor. Using the op amp in this way requires no external components and has the additional advantages of a hermetic package and unit-to-unit mechanical uniformity if replacement is ever required. The op amp's offset drift is amplified to drive the meter by the LTC1052. The diode bridge connection allows either positive or negative op amp temperature sensor offsets to interface directly with the circuit. In this case, the circuit is arranged for a +10°C to +40°C output, although other ranges are easily accommodated. To calibrate this circuit, subject the op amp sensor to a +10°C environment and adjust the 10°C trim for an appropriate meter indication. Next, place the op amp sensor in a +40°C environment and trim the 40°C adjustment for the proper reading. Repeat this procedure until both points are fixed. Once calibrated, this circuit will typically provide accuracy within ±2°C, even in high noise environments.

ELECTRONIC THERMOMETER

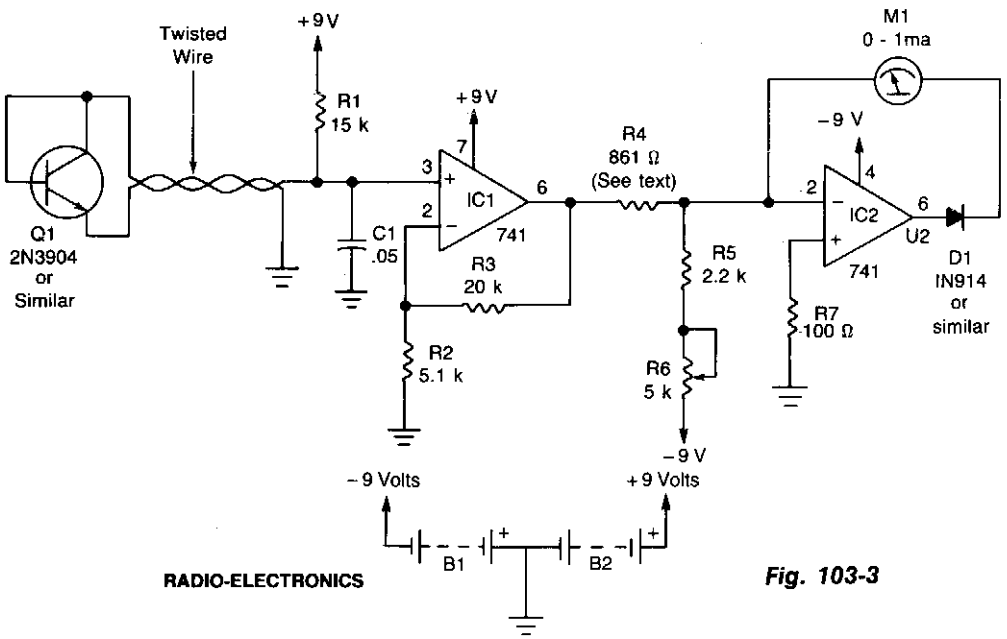


Fig. 103-3

Circuit Notes

An inexpensive electronic thermometer is capable of measuring temperatures over a range of from −30°F to +120°F. A diode-connected 2N3904 transistor used as the temperature sensor forms a voltage divider with R1. As temperature increases, the voltage drop across the transistor changes by approximately −1.166 millivolts-per°F. As a result, the current at pin 3 of IC1, a 741 op amp with a gain of 5, decreases as the temperature measured by the sensor increases.

A second 741 op amp, IC2 is configured as an inverting amplifier. Resistors R5 and R6 calibrate the circuit. Calibration is also straightforward. When properly done, a temperature of −30°F will result in a meter reading of 0 milliamps, while a temperature of 120°F will result in a meter reading of 1 milliamp. Divide the scale between those points into equal segments and mark the divisions with the appropriate corresponding temperatures. The calibration is completed by placing the sensor in an environment with a known temperature, such as an ice-point bath. Place the sensor in the bath and adjust R6 until you get the correct meter reading.

DIFFERENTIAL THERMOMETER

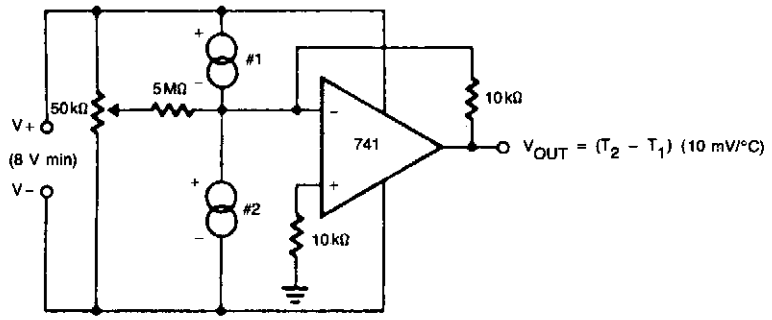


Fig. 103-4

Circuit Notes

The 50 k ohm pot trims offsets in the devices whether internal or external, so it can be used to set the size of the difference interval. This also makes it useful for liquid-level detection (where there will be a measurable temperature difference).

BASIC DIGITAL THERMOMETER, KELVIN SCALE WITH ZERO ADJUST

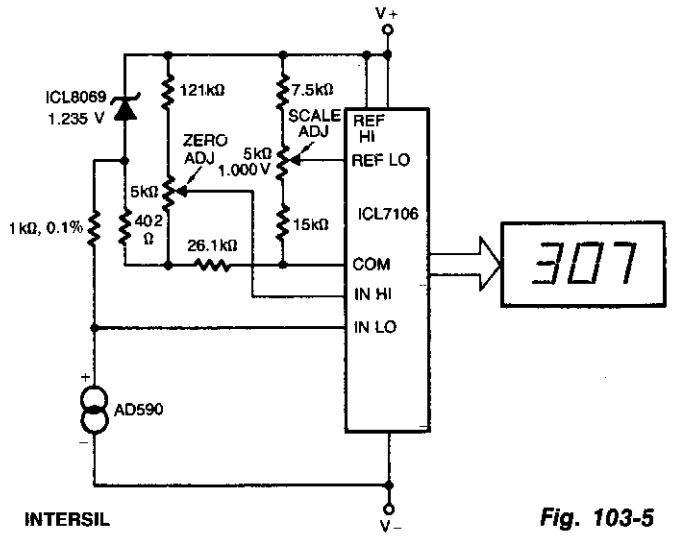
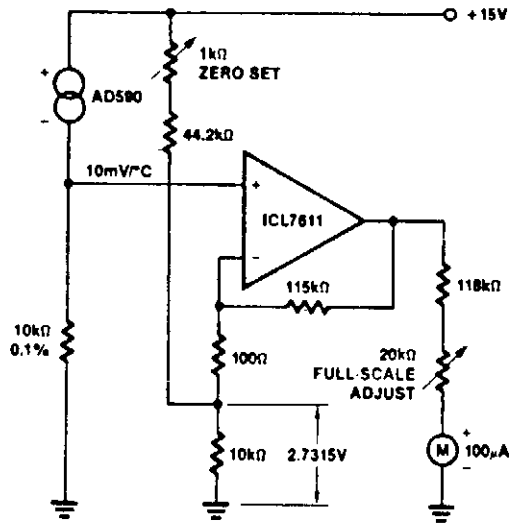


Fig. 103-5

Circuit Notes

This circuit allows "zero adjustment" as well as slope adjustment. The ICL8069 brings the input within the common-mode range, while the 5 k ohm pots trim any offset at 218° K (−55°C), and set the scale factor.

CENTIGRADE THERMOMETER (O°C-100°C)



INTERSIL

Fig. 103-6

Circuit Notes

The ultra-low bias current of the ICL7611 allows the use of large-value gain-resistors, keeping meter-current error under ½%, and therefore saving the expense of an extra meter-driving amplifier.

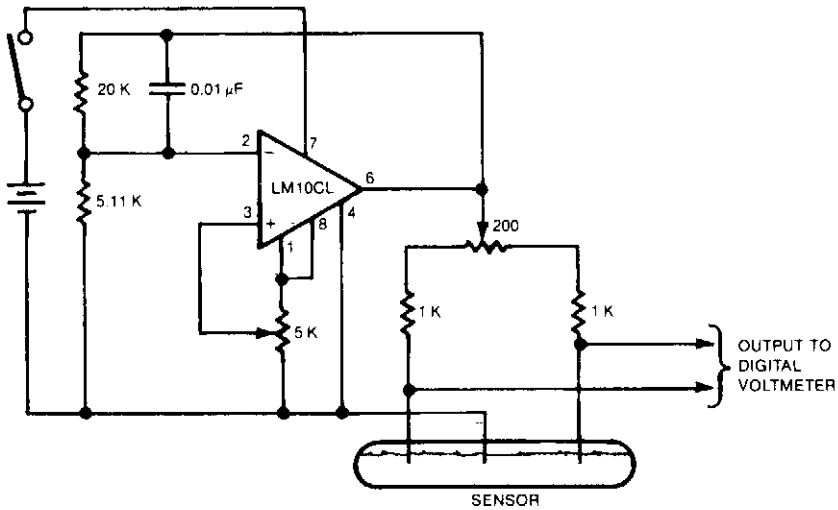
104

Tilt Meters

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Tiltmeter Indicates Sense of Slope
Differential Capacitance Measurement Circuit
Ultra-Simple Level

TILTMETER INDICATES SENSE OF SLOPE



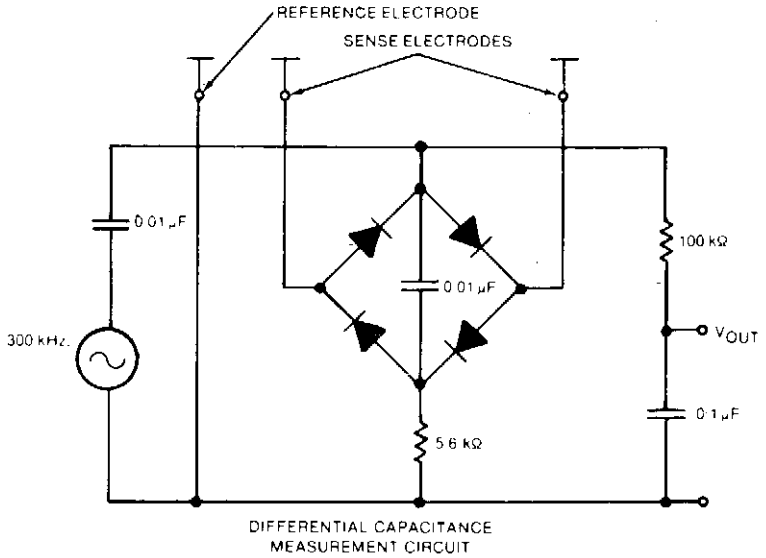
NASA

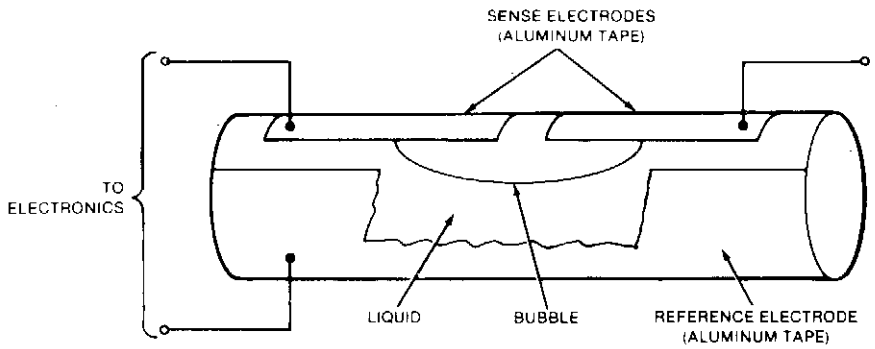
Fig. 104-1

Circuit Notes

Electrodes are immersed in an electrolyte that remains level while the sensor follows the tilt of the body on which it is placed, more of one outer electrode and less of the other are immersed and their resistances fall or rise, respectively. The resistance change causes a change in the output voltage of the bridge circuit. The sensor forms the two lower legs of the bridge, and two 1000 ohm metal film resistors and a 200 ohm ceremet balance potentiometer form the two upper legs. In preparation for use, the bridge is balanced by adjusting the balance potentiometer so that the bridge output voltage is zero when the sensor is level. The bridge input voltage (dc excitation) is adjusted to provide about 10 millivolts output per degree of slope, the polarity indicating the sense of the slope. This scaling factor allows the multimeter to read directly in degrees if the user makes a mental shift of the meter decimal point. The scaling-factor calibration is done at several angles to determine the curve of output voltage versus angle.

DIFFERENTIAL CAPACITANCE MEASUREMENT CIRCUIT





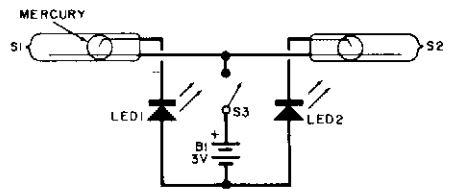
NASA

Fig. 104-2

Circuit Notes

A bubble vial with external aluminum-foil electrodes is the sensing element for a simple indicating tiltmeter. To measure bubble displacement, a bridge circuit detects the difference in capacitance between the two sensing electrodes and the reference electrode. Using this circuit, a tiltmeter level vial with 2 mm deflection for 5 arc-seconds of tilt easily resolves 0.05 arc-second. The four diodes are CA3039, or equivalent.

ULTRA-SIMPLE LEVEL



POPULAR ELECTRONICS *Fig. 104-3*

Circuit Notes

This electronic level uses two LED indicators instead of an air bubble. If the surface is tilted to the right, one LED lights; if it's tilted to the left, the other LED lights. When the surface is level, both LEDs light. It uses two unidirectional mercury switches, S1 and S2. The unidirectional mercury switch has one long electrode and one short, angled electrode. The pool of mercury "rides" on the long electrode and makes contact between the two electrodes if the unit is held in a horizontal position.

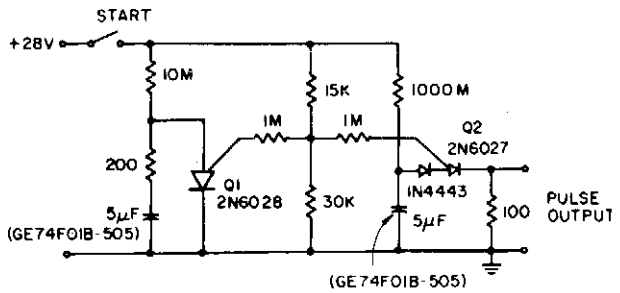
105

Time-Delay Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Hour Time-Delay Sampling Circuit
Time Delay With Constant Current Charging
Low-Cost Integrator Multiplies 555 Timer's Delay
Simple Time-Delay Circuit Using Two SCRs

HOUR TIME-DELAY SAMPLING CIRCUIT



GENERAL ELECTRIC

Fig. 105-1

Circuit Notes

The circuit lowers the effective peak current of the output PUT, Q2. By allowing the capacitor to charge with high gate voltage and periodically lowering gate voltage, when Q1 fires, the timing resistor can be a value which supplies a much lower current than I_P. The triggering requirement here is that minimum charge to trigger flow through the timing resistor during the period of the Q1 oscillator. This is not capacitor size dependent, only capacitor leakage and stability dependent.

TIME DELAY WITH CONSTANT CURRENT CHARGING

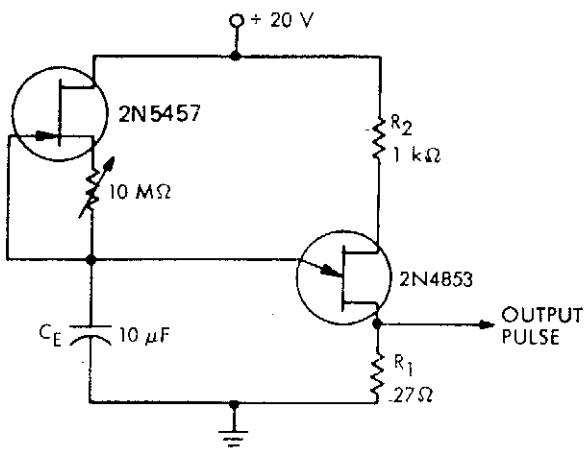
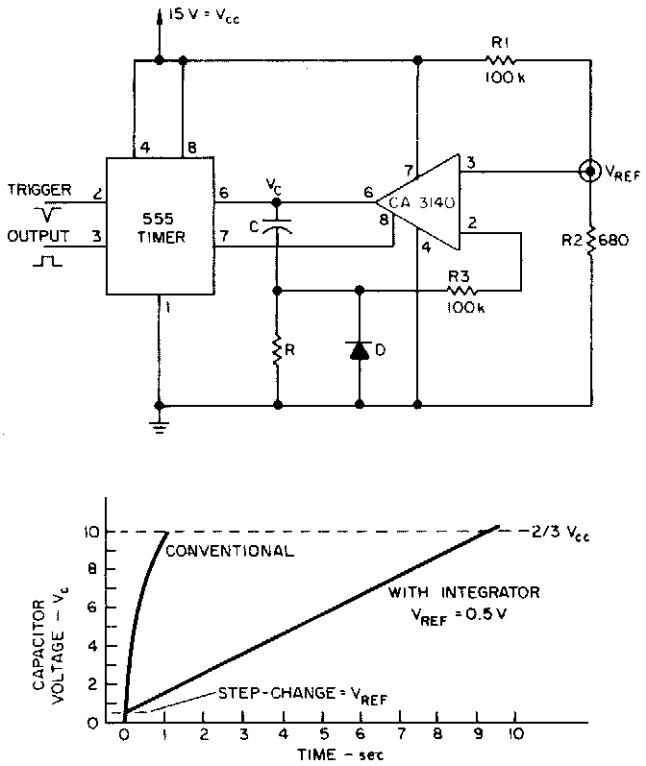


Fig. 105-2

MOTOROLA INC.

LOW-COST INTEGRATOR MULTIPLIES 555 TIMER'S DELAY



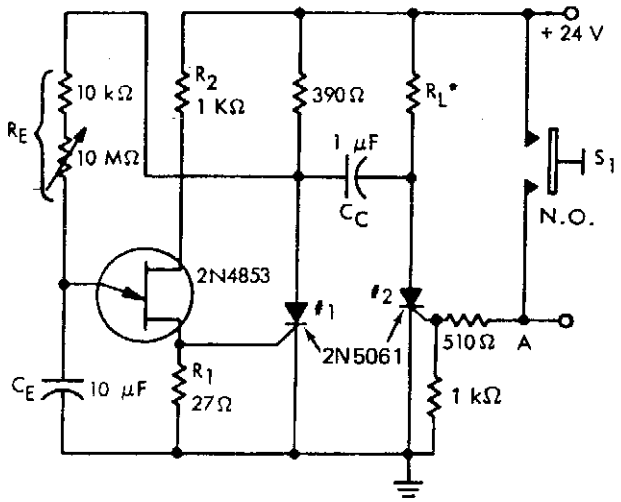
ELECTRONIC DESIGN

Fig. 105-3

Circuit Notes

Long delay times can be derived from a 555 timer with reasonably sized capacitors if an integrator circuit is used. The capacitor's charging time with an integrator circuit can be much longer than with a conventional 555-timer configuration.

SIMPLE TIME-DELAY CIRCUIT USING TWO SCRs



*Value of R_L must be low enough to allow hold current to flow in the SCR.

MOTOROLA INC.

Fig. 105-4

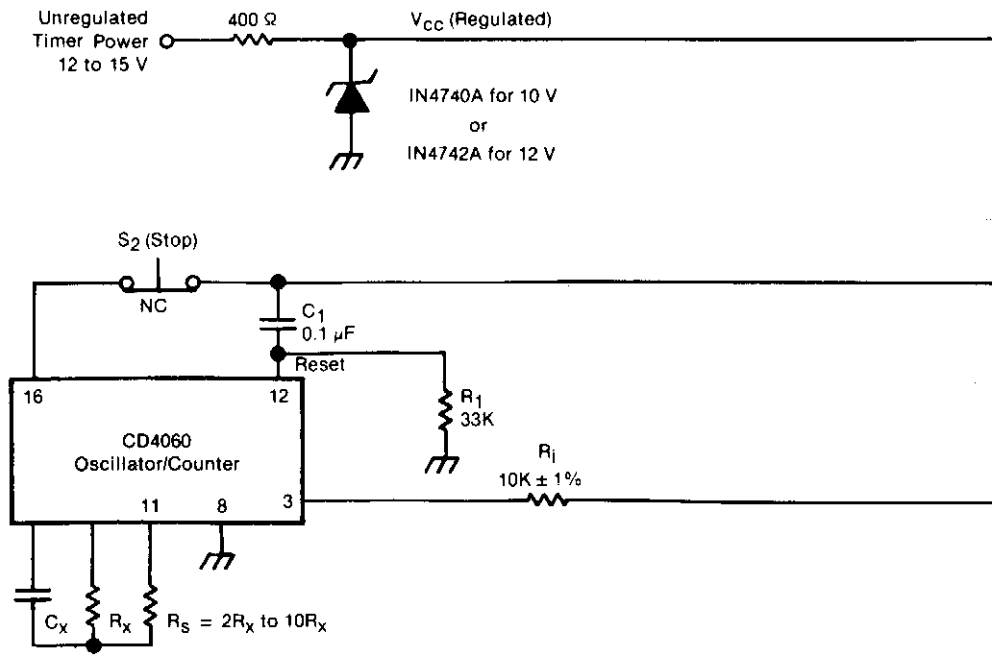
106

Timers

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Long-Term Electronic Timer
Timer with Alarm
Timer Circuit
PUT Long Duration Timer
Programmable Voltage Controlled Timer
Low Power Microprocessor-Programmable Interval Timer
Precision Elapsed Time/Countdown Timer
Adjustable AC Timer .2 to 10 sec.

LONG-TERM ELECTRONIC TIMER

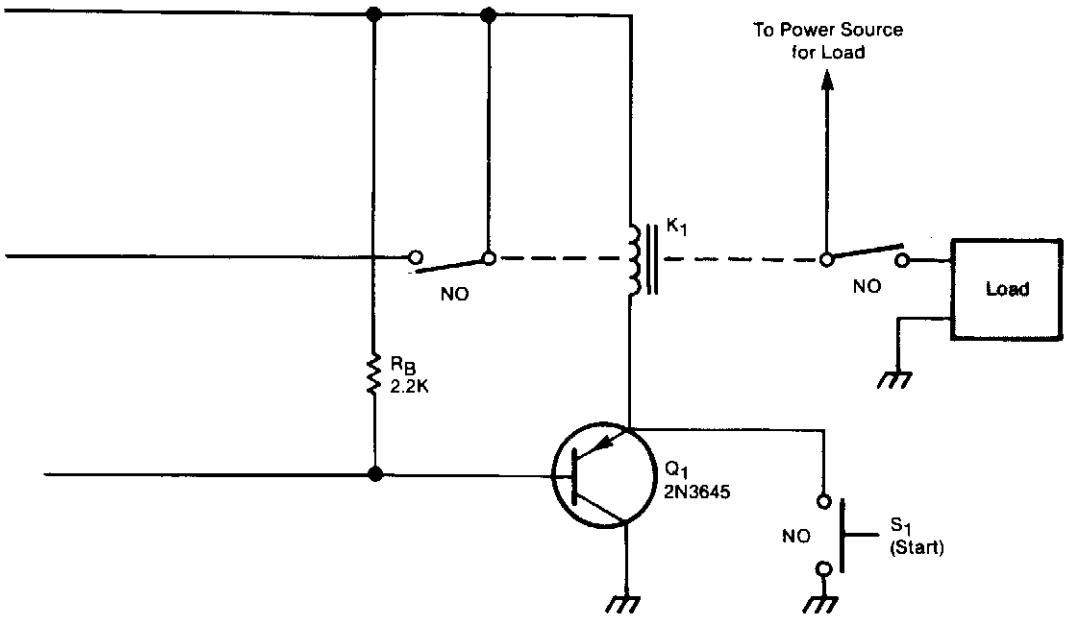


NASA

Fig. 106-1

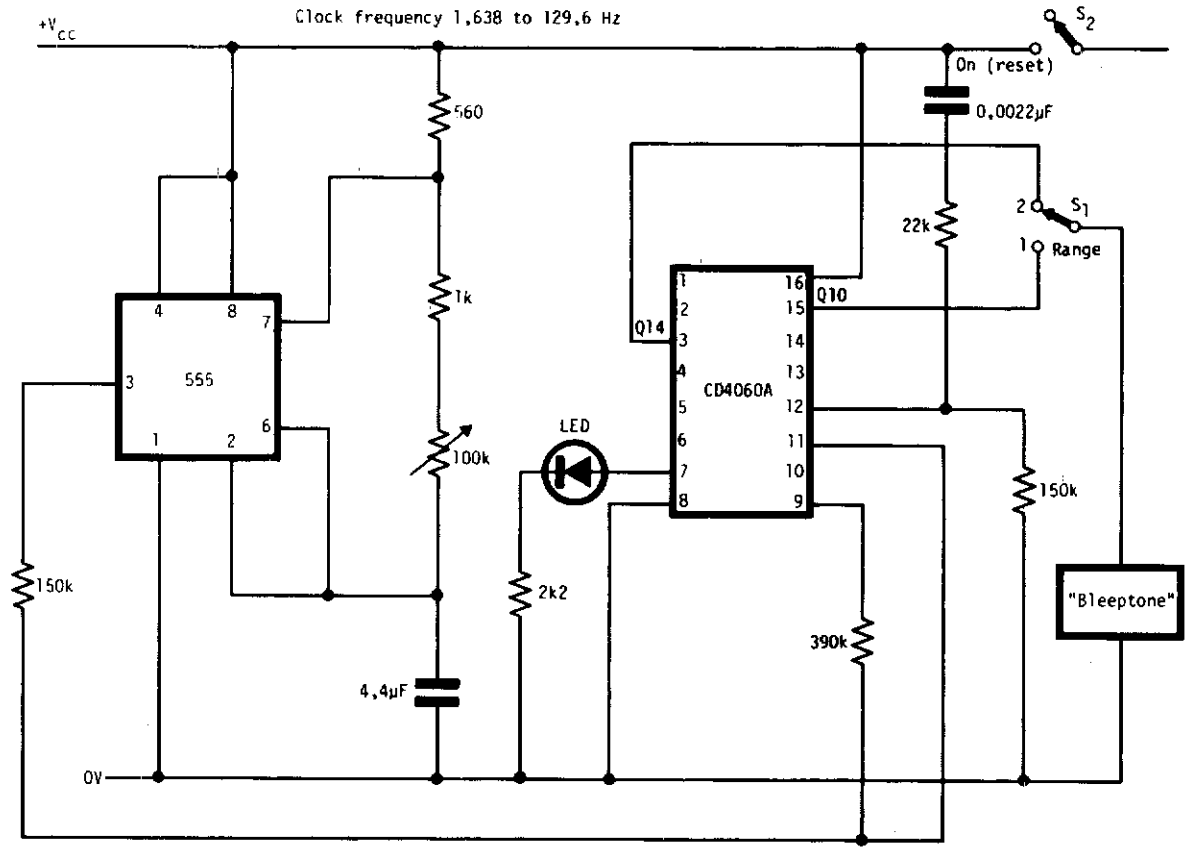
Circuit Notes

The timer includes an oscillator and a counter in an integrated circuit. The timing interval equals the oscillator period multiplied by the number of cycles to be counted. The oscillator frequency depends upon resistor RS and capacitor CX. The number of oscillator cycles to be counted before the counter output changes state is determined by the selection of the counter output terminal, shown here as pin 3. The interval can be set anywhere in the range from fractions of a second to months; it is given by $T = 0.55\ R_S C_X 2^n$, where n is an integer determined by the counter-output selection. Operation is initiated by the closure of momentary switch S1 (or by a command signal having a similar effect). This grounds one side of relay K1, thereby activating the relay



and causing the closure of the switches that supply power to the timer and to the load. The turn-on of V_{CC} at the timer is coupled through C1 to the counter-reset terminal, thus resetting the counter. The initial reset voltage transient is then drained away through R1 to permit normal operation. During the first half cycle of the counter operation, the counter output voltage (at pin 3 in this case) is low. This turns on transistor Q1 so that relay K1 latches on, enabling the timer to continue running even though switch S1 has opened. The oscillator runs while the relay is on. When the number of oscillator cycles reaches the limit, the counter output voltage at pin 3 goes high. This turns off Q1, thereby turning off the relay and returning the system to the original "power-off" state to await the next starting command. The timing cycle can also be interrupted and the system turned off by opening normally-closed switch S2.

TIMER WITH ALARM



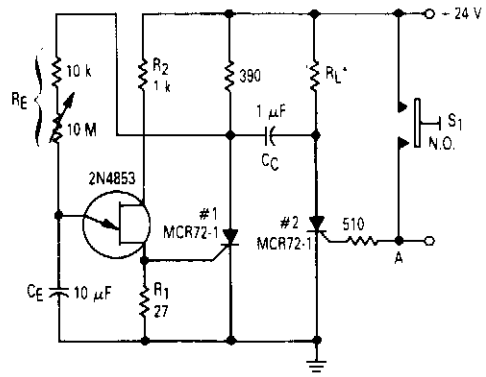
ELECTRONIC ENGINEERING

Fig. 106-2

Circuit Notes

The circuit has two ranges: 10 secs to 5 mins and 1 min to 80 mins. It can be powered by a 9-V battery. With the LED connected as shown a reasonable frequency of flashing occurs throughout the range of operation. This circuit is reset when S2 is closed.

TIMER CIRCUIT



*VALUE OF R_L MUST BE LOW ENOUGH TO ALLOW HOLD CURRENT TO FLOW IN THE SCR.

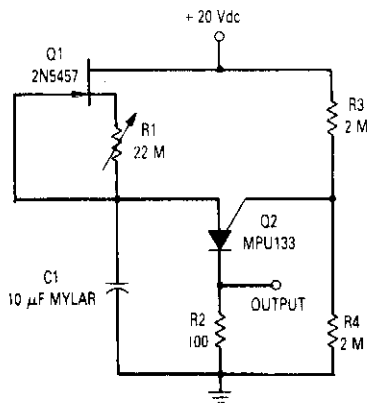
MOTOROLA

Fig. 106-3

Circuit Notes

After one cycle of operation, SCR 1 will be on, and a low value of voltage is applied to the UJT emitter circuit, interrupting the timing function. When pushbutton S1 is pushed, or a positive going pulse is applied at point A, SCR 2 will turn on, and SCR 1 will be turned off by commutating capacitor CC. With SCR 1 off, the supply voltage will be applied to RE and the circuit will begin timing again. After a period of time determined by the setting of RE, the UJT will fire and turn SCR 1 on and commutate SCR 2 off. The time delay is determined by the charge time of the capacitor.

PUT LONG DURATION TIMER



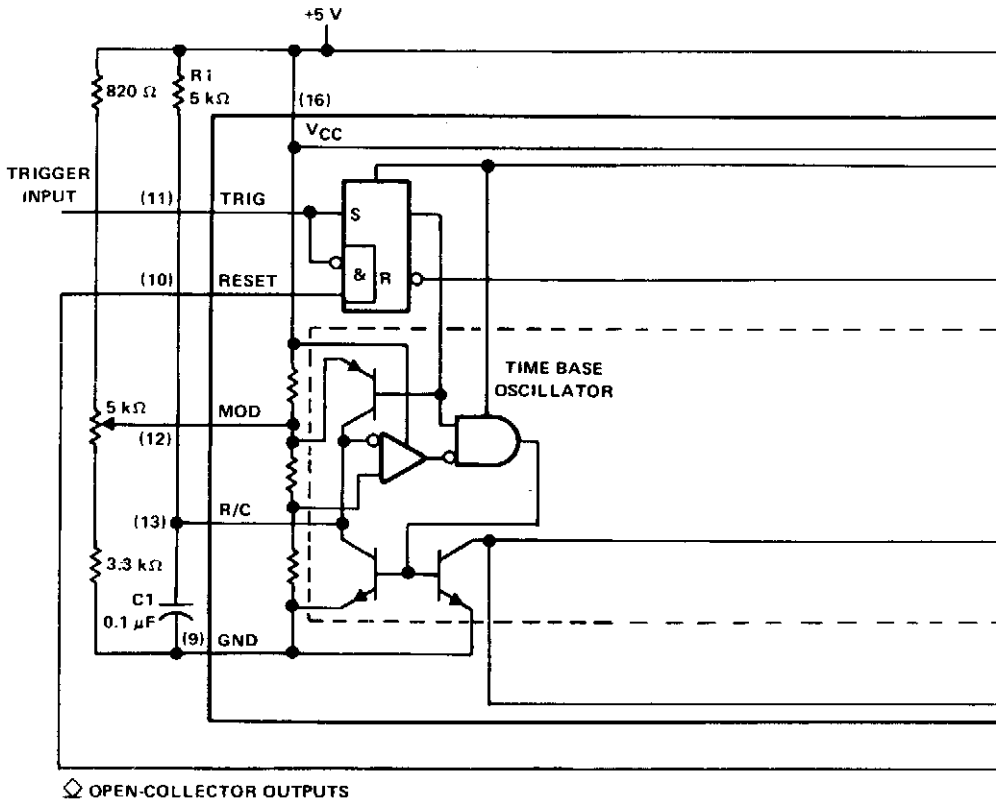
MOTOROLA

Fig. 106-4

Circuit Notes

The time circuit can provide a time delay of up to 20 minutes. The circuit is a standard relaxation oscillator with a FET current source in which resistor R1 is used to provide reverse bias on the gate-to-source of the JFET. This turns the JFET off and increases the charging time of C1. C1 should be a low leakage capacitor such as a mylar type.

PROGRAMMABLE VOLTAGE CONTROLLED TIMER

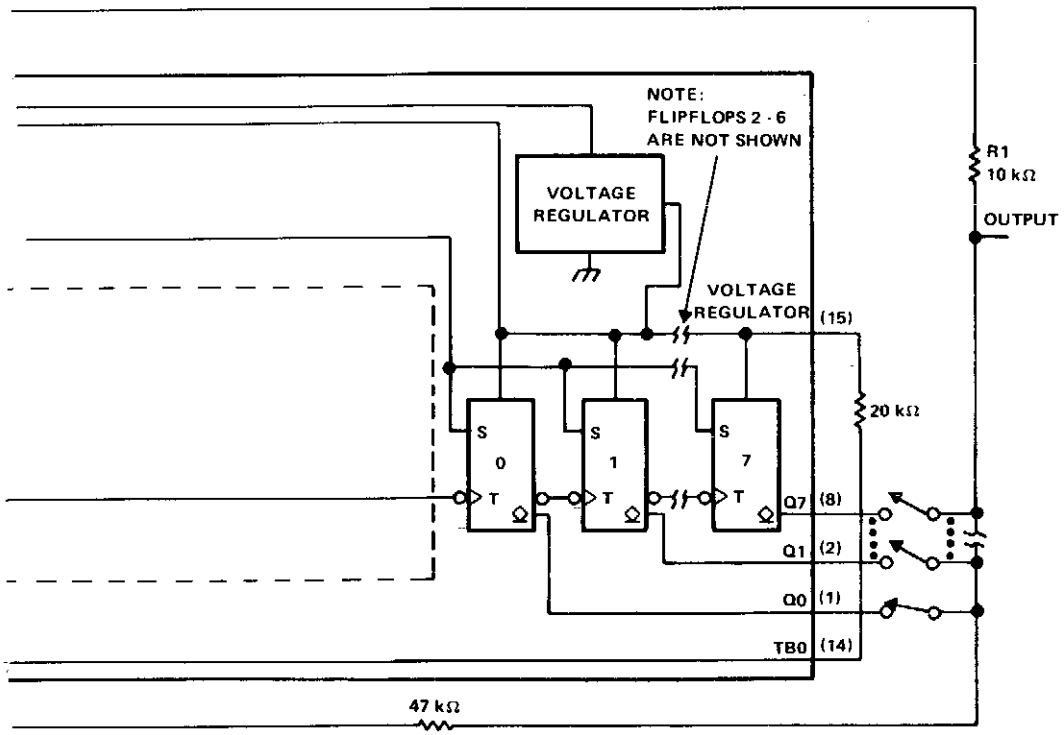


TEXAS INSTRUMENTS

Fig. 106-5

Circuit Notes

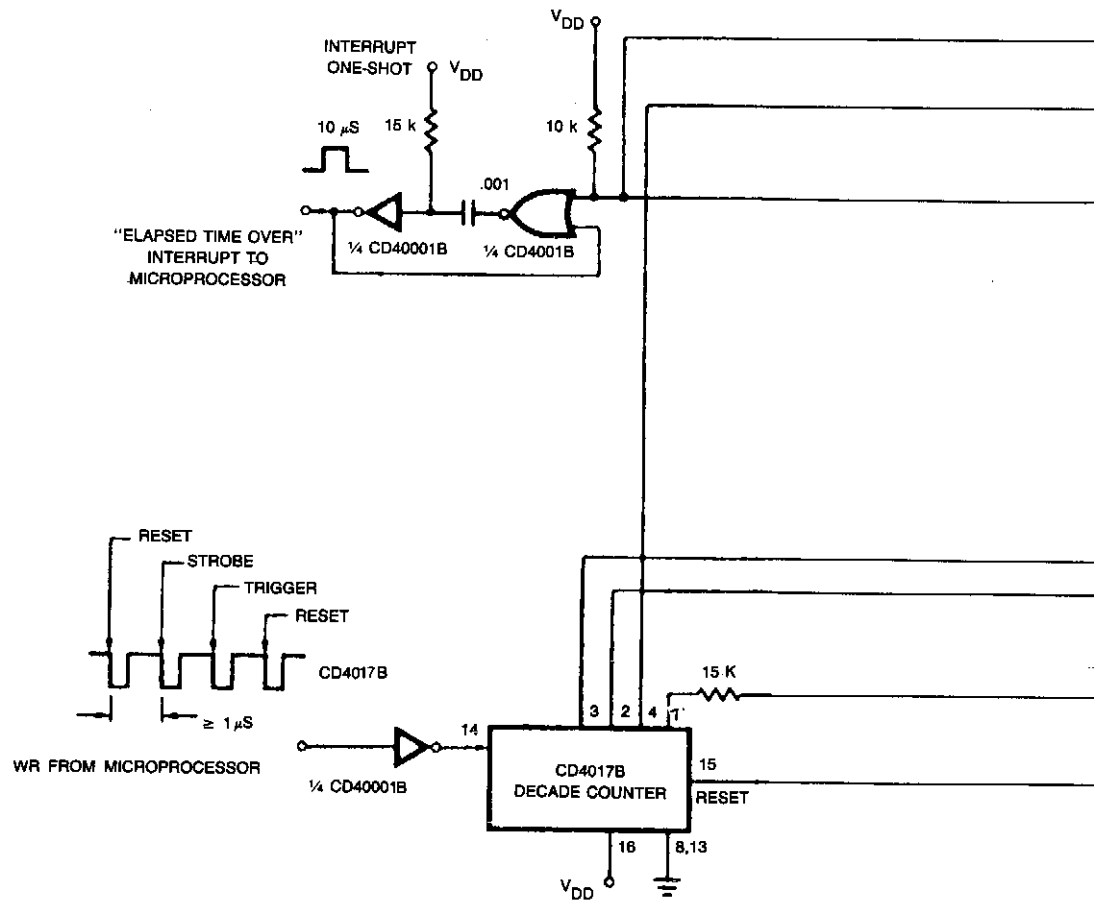
The μA2240 may easily be configured as a programmable voltage controlled timer with a minimum number of external components. The modulation input (pin 12), which allows external adjustment of the input threshold level. A variable voltage is applied from the arm of a 10 k ohm potentiometer connected from V_{CC} to ground. A change in the modulation input voltage will result in a change in the time base oscillator frequency and the period of the time base output (TBO). The TBO has an open-collector output that



is connected to the regulator output via a 10 k ohm pull-up resistor. The output of the TBO drives the input to the 8-stage counter section.

At start-up, a positive trigger pulse starts the TBO and sets all counter outputs to a low state. The binary outputs are open-collector stages that may be connected together to the 10 k ohm pull-up resistor to provide a "wired-OR" output function. This circuit may be used to generate 255 discrete time delays that are integer multiples of the time-base period. The total delay is the sum of the number of time-base periods, which is the binary sum of the Q outputs connected. Delays from 200 μs to 0.223 s are possible with this configuration.

LOW POWER MICROPROCESSOR PROGRAMMABLE INTERVAL TIMER

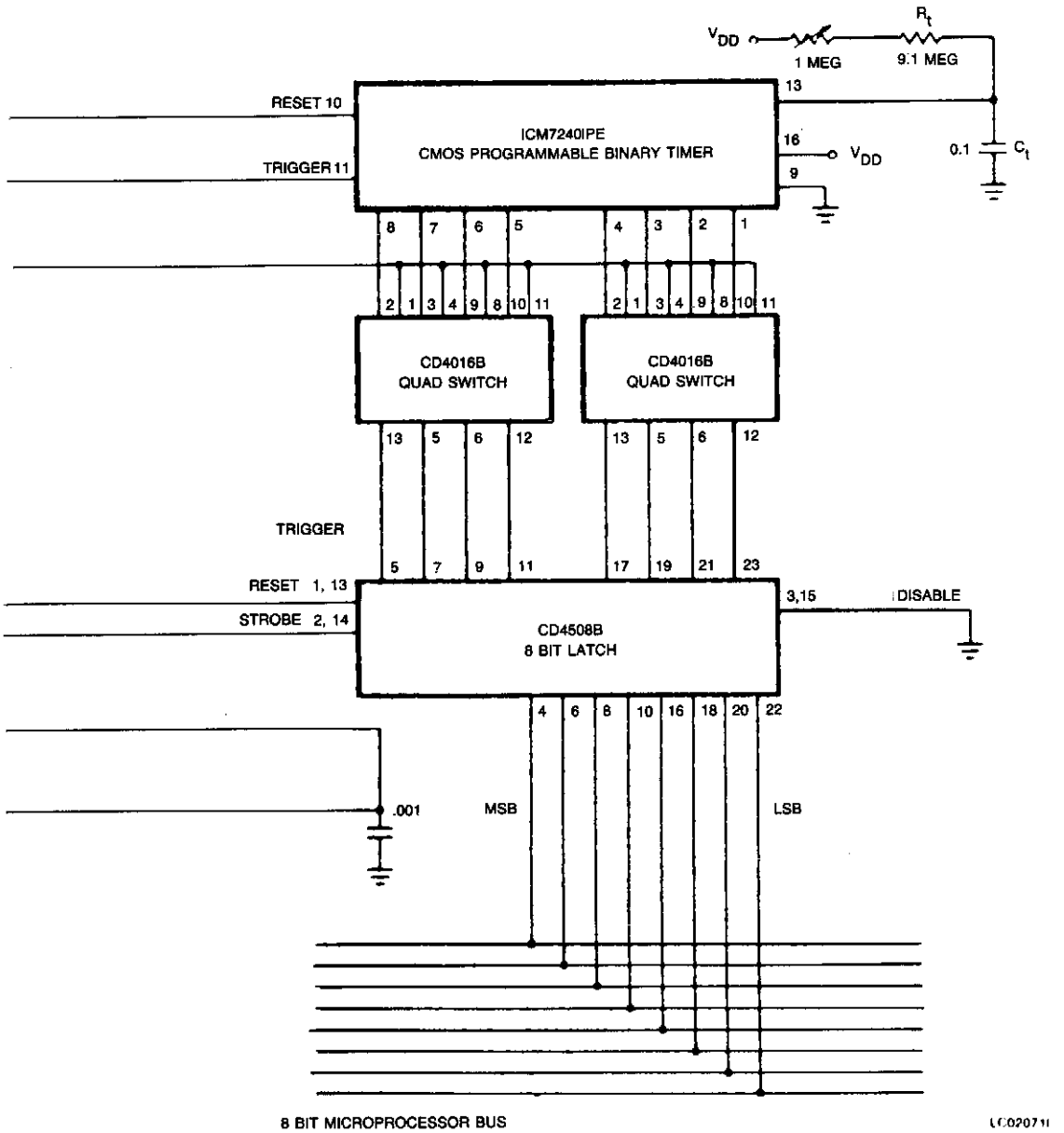


INTERSIL

Fig. 106-6

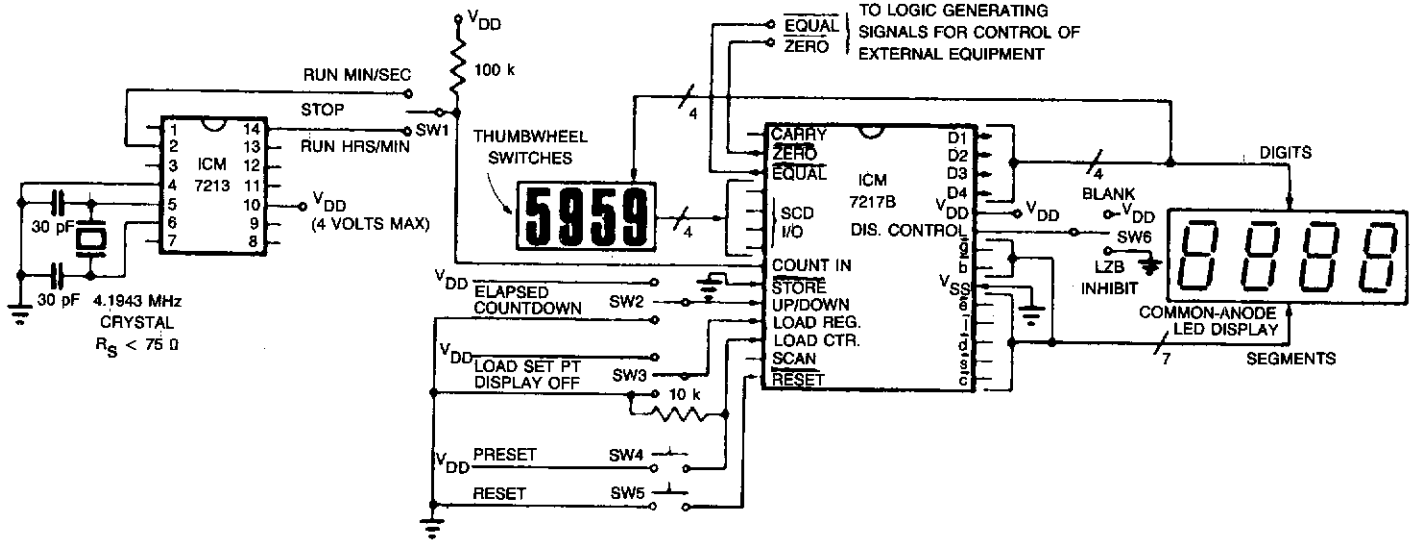
Circuit Notes

The microprocessor sends out an 8-bit binary code on its 8-bit I/O bus (the binary value needed to program the ICM7240), followed by four WRITE pulses into the CD4017B decade counter. The first pulse resets the 8-bit latch, the second strobes the binary value into the 8-bit latch, the third triggers the ICM7240 to begin its timing cycle and the fourth resets the decade counter. The ICM7240 then counts the interval of time



determined by the R-C value on pin 13, and the programmed binary count on pins 1 through 8. At the end of the programmed time interval, the interrupt one-shot is triggered, informing the microprocessor that the programmed time interval is over. With a resistor of approximately 10 M ohm and a can capacitor of 0.1 μF, the time base of the ICM7240 is one second. Thus, a time of 1-255 seconds can be programmed by the microprocessor, and by varying R or C, longer or shorter time bases can be selected.

PRECISION ELAPSED TIME/COUNTDOWN TIMER



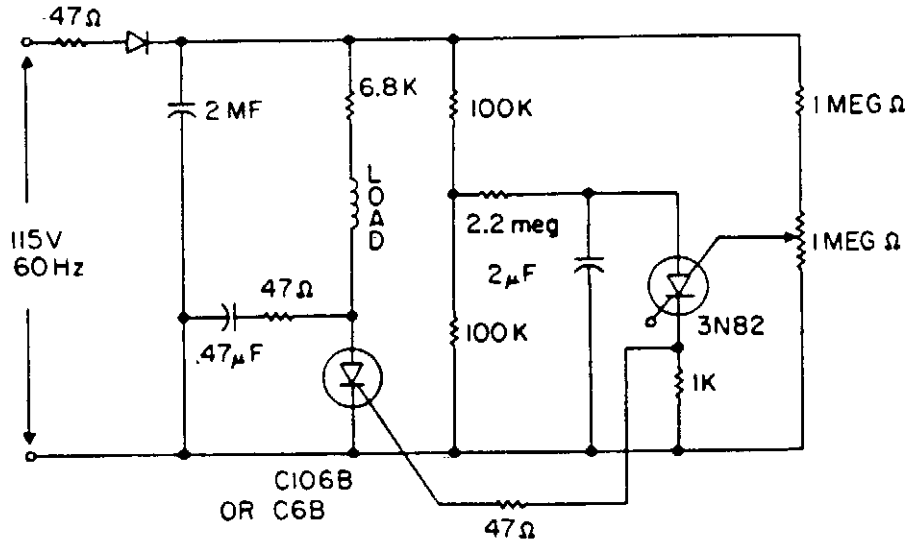
INTERSIL

Fig. 106-7

Circuit Notes

The circuit uses an ICM7213 precision one minute/one second timebase generator using a 4.1943 MHz crystal for generating pulses counted by an ICM7217B. The thumbwheel switches allow a starting time to be entered into the counter for a preset-countdown type timer, and allow the register to be set for compare functions. For instance, to make a 24-hour clock with BCD output the register can be preset with 2400 and the EQUAL output used to reset the counter. Note the 10 k resistor connected between the LOAD COUNTER terminal and ground. This resistor pulls the LOAD COUNTER input low when not loading, thereby inhibiting the BCD output drivers. This resistor should be eliminated and SW4 replaced with an SPDT center-off switch if the BCD outputs are to be used.

ADJUSTABLE AC TIMER .2 TO 10 SEC.



GENERAL ELECTRIC

Fig. 106-8

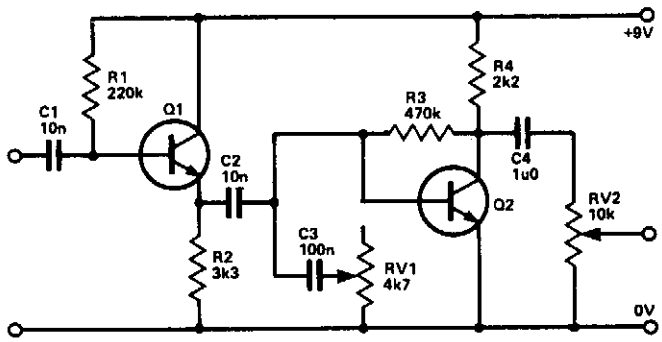
107

Tone Control Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Guitar Treble Boost
Tone Control
Ten Band Graphic Equalizer, Using Active Filters
Tone-Control Audio Amplifier
Mike Preamp with Tone Control
Low Cost High-Level Preamp and Tone Control Circuit
Passive Tone-Control Circuit

GUITAR TREBLE BOOST



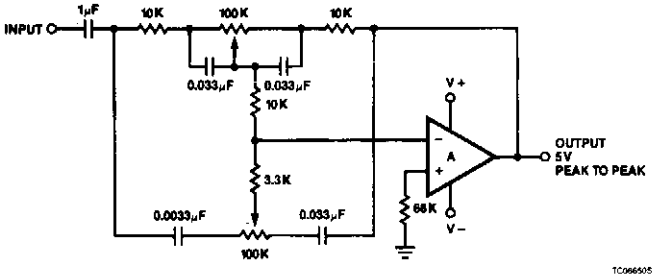
ELECTRONICS TODAY INTERNATIONAL

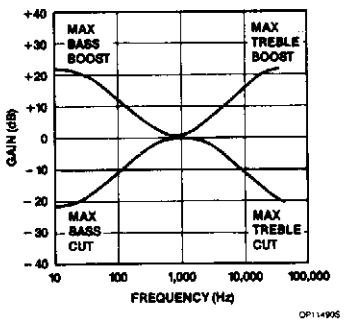
Fig. 107-1

Circuit Notes

Q1 is connected as an emitter follower in order to present a high input impedance to the guitar. C2, being a relatively low capacitance, cuts out most of the bass, and C3 with RV1 acts as a simple tone control to cut the treble, and hence the amount of treble boost can be altered. Q2 is a simple preamp to recover signal losses in C2, C3, and RV1.

TONE CONTROL





NOTES:
1. Amplifier A may be a NE531 or 301. Frequency compensation, as for unity gain non-inverting amplifiers, must be used.
2. Turn-over frequency — 1kHz.
3. Bass boost + 20dB at 20Hz, bass cut - 20dB at 20Hz, treble boost + 19dB at 20Hz, treble cut - 19dB at 20Hz.

All resistor values are in ohms.

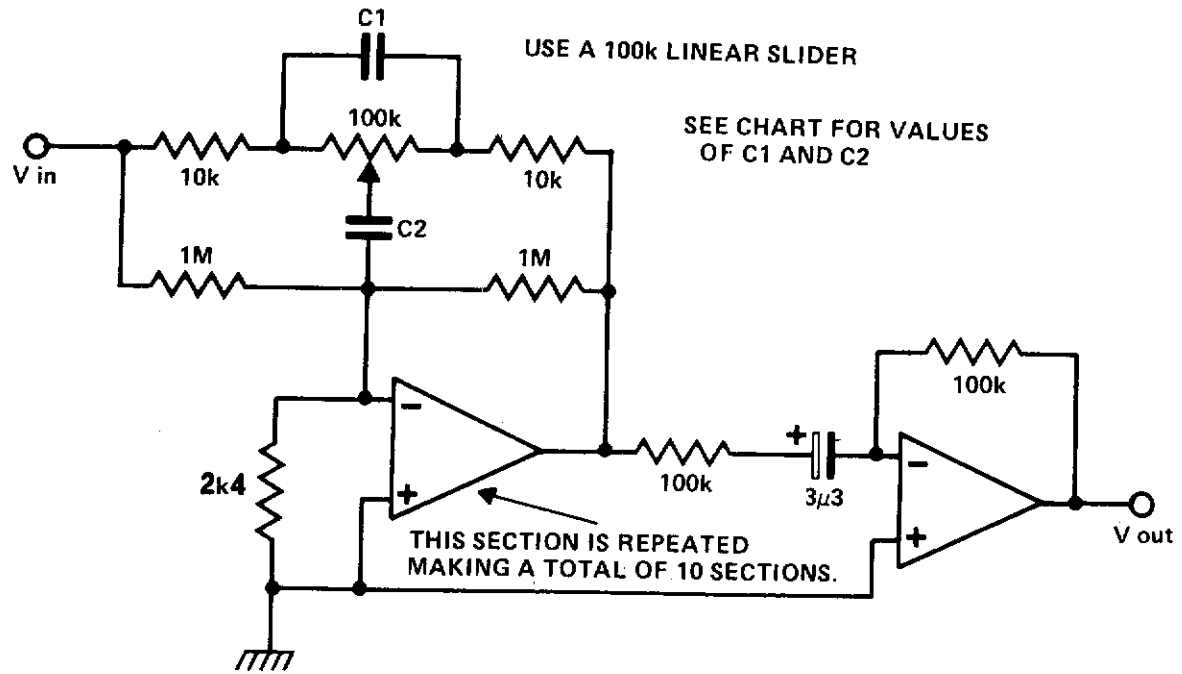
SIGNETICS

Fig. 107-2

Circuit Notes

Tone control of audio systems involves altering the flat response in order to attain more low frequencies or more high ones, dependent upon listener preference. The circuit provides 20 dB of bass or treble boost or cut as set by the variable resistance. The actual response of the circuit is shown also.

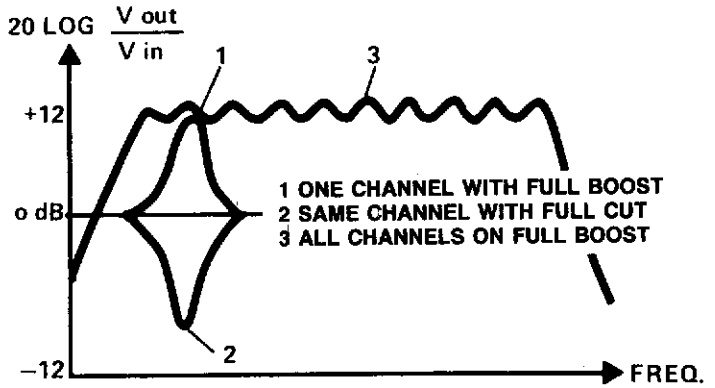
TEN BAND GRAPHIC EQUALIZER, USING ACTIVE FILTERS



ELECTRONICS TODAY INTERNATIONAL

Fig. 107-3

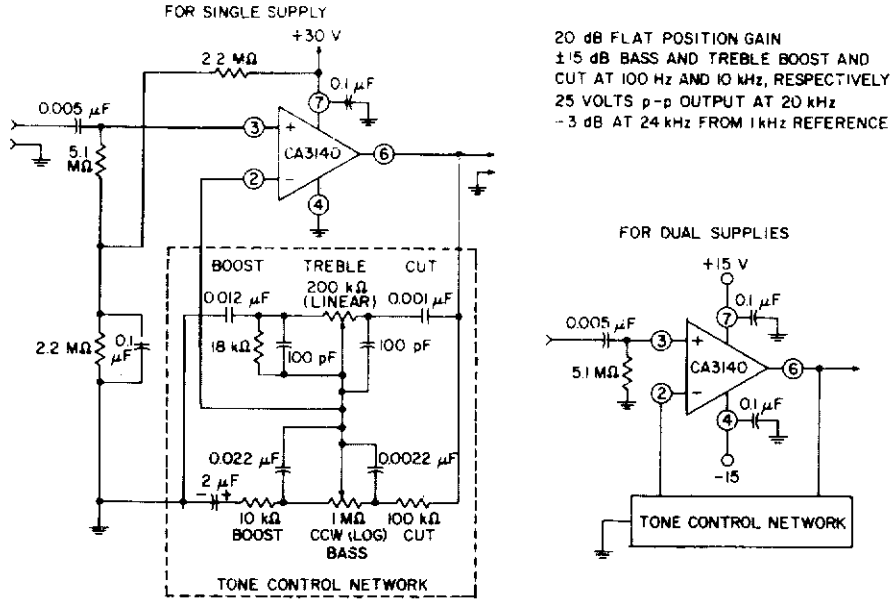
CHANNEL CENTRE FREQ. IN Hz	C1	C2
32	180n	18n
64	100n	10n
125	47n	4n7
250	22n	2n2
500	12n	1n2
1000	5n6	560p
2000	2n7	270p
4000	1n5	150p
8000	680p	68p
16000	360p	36p



Circuit Notes

The above circuit is repeated ten times. Use the table to calculate values.

TONE-CONTROL AUDIO AMPLIFIER



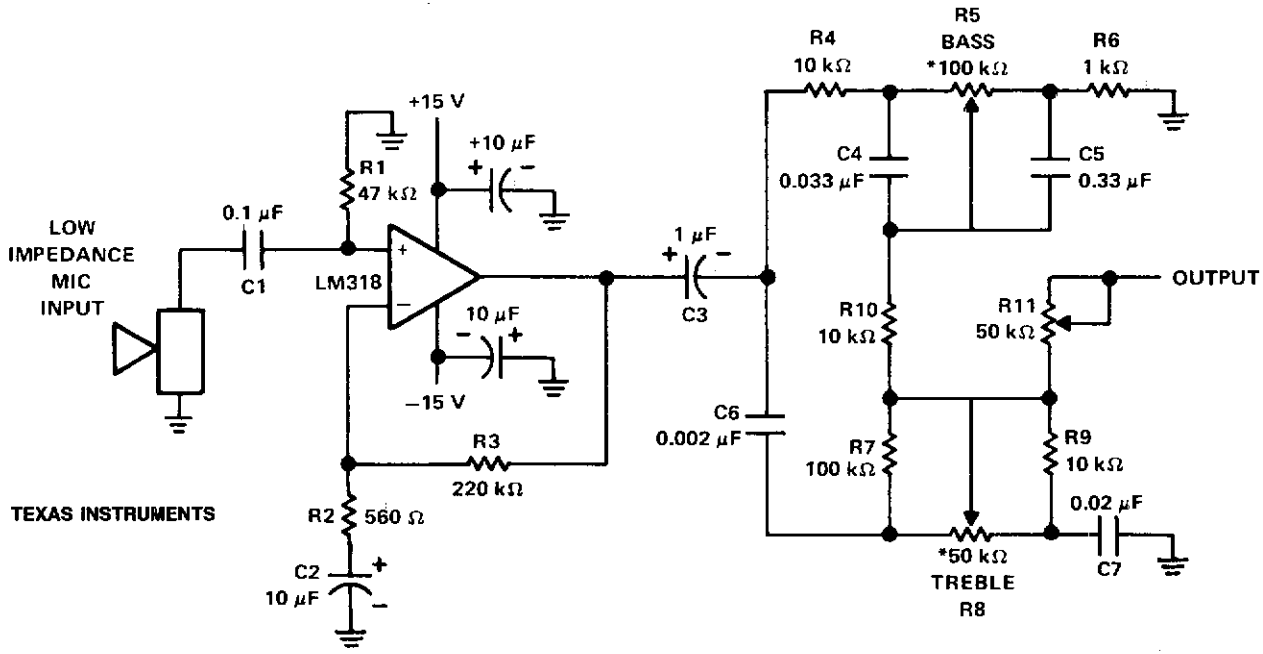
GENERAL ELECTRIC/RCA

Fig. 107-4

Circuit Notes

The circuit makes excellent use of the high slew rate, wide bandwidth, high input impedance, and high-output voltage capability of the CA3140 BiMOS op amp. The wideband gain of this circuit is equal to the ultimate boost or cut plus one, in this case a gain of eleven. For 20-dB boost or cut, input loading is essentially equal to the resistance from terminal 3 to ground.

MIKE PREAMP WITH TONE CONTROL



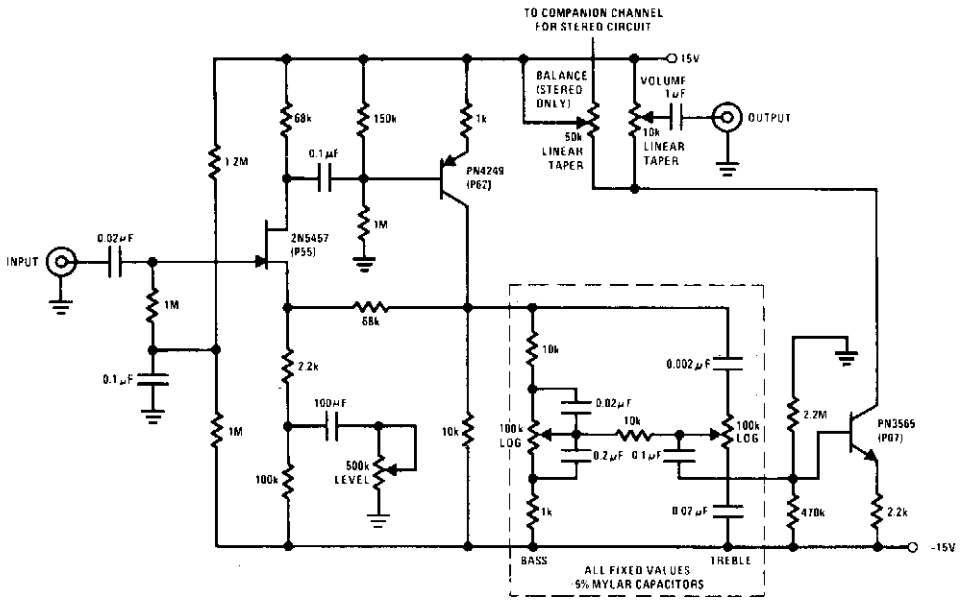
*THE TONE CONTROLS ARE AUDIO TAPER (LOG) POTENTIOMETERS.

Fig. 107-5

Circuit Notes

The LM318 op amp is operated as a standard non-inverting amplifier. Resistor R1 (47 k ohm) provides an input path to ground for the bias current of the non-inverting input. The combination of R2 (560 ohm) and C2 (10 μF) provides a frequency roll-off below 30 Hz. At 30 Hz and above the gain is relatively flat at about 50 dB, set by the ratio R3/R2. R3 (200 k ohm) furnishes negative feedback from the output to the inverting input of the op amp. C3 (1.0 μF electrolytic) ac couples the preamp to the tone control section. The top half of the tone control section is the bass control. The bottom half controls the treble frequency response. These tone controls (R5 and R8) require audio taper (logarithmic) potentiometers. The 50 k ohm potentiometer on the output can be used to set the output or gain of the preamp.

LOW COST HIGH-LEVEL PREAMP AND TONE CONTROL CIRCUIT



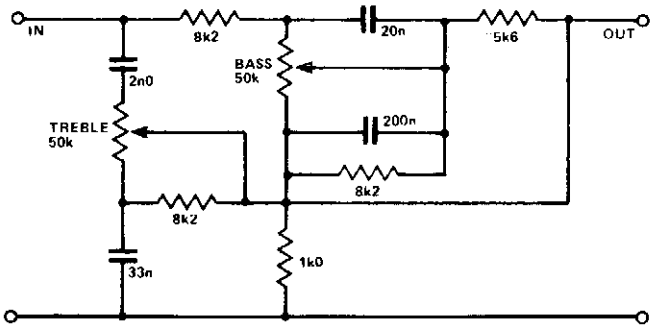
NATIONAL SEMICONDUCTOR CORP.

Fig. 107-6

Circuit Notes

This preamp and tone control uses the JFET to its best advantage; as a low noise high input impedance device. All device parameters are noncritical, yet the circuit achieves harmonic distortion levels of less than 0.05% with a S/N ratio of over 85 dB. The tone controls allow 18 dB of cut and boost; the amplifier has a 1-V output for 100-mV input at maximum level.

PASSIVE TONE-CONTROL CIRCUIT



ELECTRONICS TODAY INTERNATIONAL *Fig. 107-7*

Circuit Notes

A simple circuit using two potentiometers and easily available standard value components provides tone control. The impedance level is suitable for low-level transistor or op amp circuitry.

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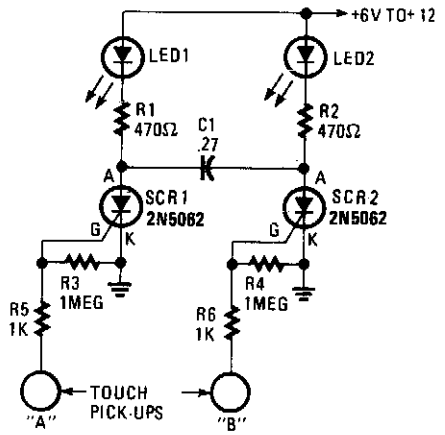
Touch-Switch Circuits

The sources of the following circuits are contained in the Sources section beginning on page 694. The figure number contained in the box of each circuit correlates to the source entry in the Sources section.

Touch On/Off Switch
Touch Switch
Touchomatic

TOUCH ON/OFF SWITCH

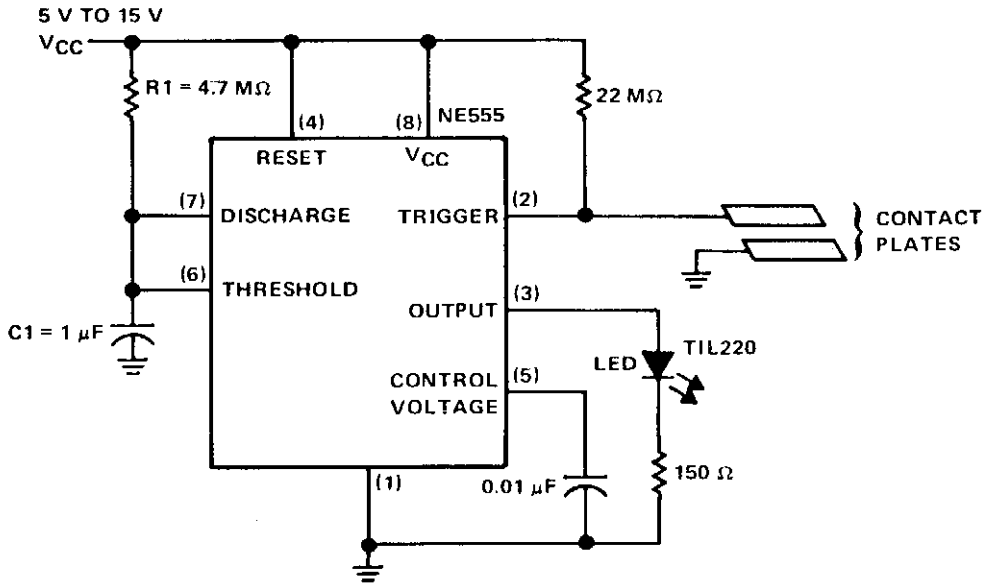


HANDS-ON ELECTRONICS *Fig. 108-1*

Circuit Notes

If a Touch On/Off Switch is desired, this circuit fills the bill. Two sensitive gate SCRs are interconnected, so that when one of the devices is turned on, the other (if on) is forced off. That toggling effect gives an on/off circuit condition for each of the LEDs in the SCR-anode circuits. To turn LED1 on and LED2 off, simply touch the "A" terminal, and to turn LED1 off and LED2 on, the "B" pick-up must be touched. It is possible to simultaneously touch both terminals, causing both SCRs to turn on together. To reset the circuit to the normal one-on/one-off condition, momentarily interrupt the circuit's dc power source. Additional circuitry can be connected to the anode circuit of either or both SCRs to be controlled by the on/off function of the touch switch.

TOUCH SWITCH



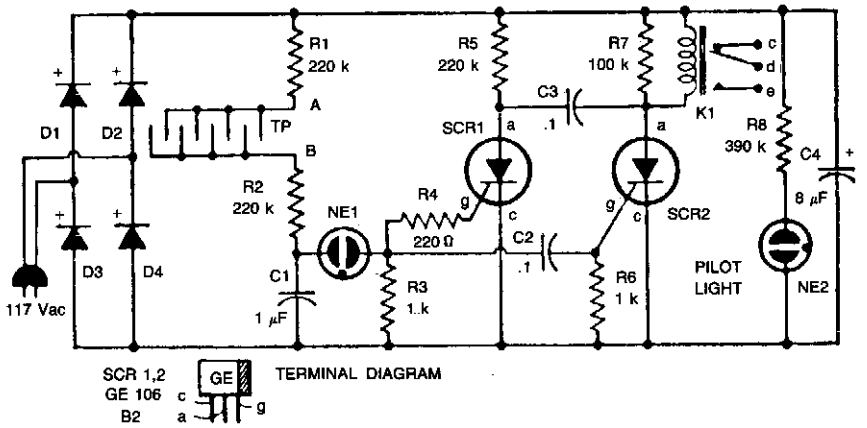
TEXAS INSTRUMENTS

Fig. 108-2

Circuit Notes

The circuit is basically a NE555 monostable, the only major difference being its method of triggering. The trigger input is biased to a high value by the 22 M ohm resistor. When the contact plates are touched, the skin resistance of the operator will lower the overall impedance from pin 2 to ground. This action will reduce the voltage at the trigger input to below the ⅓ V_{CC} trigger threshold and the timer will start. The output pulse width will be T = 1.1 R1C1, in this circuit about 5 seconds. A relay connected from pin 3 to ground instead of the LED and resistor could be used to perform a switching function.

TOUCHOMATIC



TAB BOOKS, INC.

Fig. 108-3

Circuit Notes

When someone touches the touchplate (TP), the resistance of his finger across points A and B is added in series to the combination of R1 and R2, the capacitor C2 begins to charge. When the voltage across C1 is finally sufficient to fire NE1, C1 will begin to discharge. When NE1 fires, it produces a short between its terminals. Since R3 is connected across C1, they are effectively in series after NE1 fires. A voltage spike will then be passed by C2 and this will act as a positive triggering pulse. The pulse is fed to both SCR gates: SCR2 conducts, thereby closing relay K1. With a finger no longer on the touchplate, no more pulses are forthcoming because the C1 charge path is open. The next contact with the touchplate will produce a pulse which triggers SCR1. SCR2 is now off by capacitor C3 which was charged by current passing through R6 and SCR2. The firing of SCR1 in this way places a negative voltage across SCR2 which momentarily drops the relay current to a point below the holding current value of SCR2. (Holding current is the minimum current an SCR requires to remain in a conducting state once its gate voltage is removed.) With SCR2 turned off, the relay will open and SCR1 will turn off due to the large resistance in series with its anode. Starved in this way SCR1 turns off because of a forced lack of holding current.

Sources Index

Chapter 1

Fig. 1-1. Signetics 1987 Linear Data Manual Vol. 2, 2/87, p.7-65.

Fig. 1-2. General Electric Application Note 90.16, p. 25.

Fig. 1-3. Courtesy, William Sheets.

Fig. 1-4. General Electric Application Note 90.16, p. 25.

Fig. 1-5. General Electric/RCA, BiMOS Operational Amplifiers Circuit Ideas, 1987, p. 27.

Fig. 1-6. R-E Experimenters Handbook, p. 157.

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Fig. 2-1. Linear Technology Corporation, Linear Applications Handbook, 1987, p. AN21-2

Fig. 2-2. Linear Technology Corporation, Linear Databook Supplement, 1988, p. S2-34.

Fig. 2-3. Siliconix, Integrated Circuits Data Book, 3/85, p. 10-154.

Fig. 2-4. Linear Technology Corp., Linear Databook, 1986, p. 2-83.

Fig. 2-5. National Semiconductor Corp., Transistor Databook, 1982, p. 11-23.

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Fig. 2-7. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 10/86, p. 4-260.

Fig. 2-8. Electronic Engineering, 11/86, p. 40.

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Fig. 2-12. Electronics Engineering, 9/84, p. 33.

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Fig. 2-25. NASA Tech Briefs, Sept/Oct 1986, p. 43.

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Fig. 3-6. Signetics, 1987 Linear Data Manual, Vol. 2: Industrial, 11/86, p. 5-215.
Fig. 3-7. National Semiconductor Corp., 1984 Linear Supplement Databook, p. S5-126.
Fig. 3-8. Signetics, 1987 Linear Data Manual Vol. 2: Industrial, 12/86, p. 4-67.

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Fig. 4-1. Electronic Design, 5/82, p. 214.
Fig. 4-2. R-E Experimenters Handbook, p. 160.
Fig. 4-3. Popular Electronics, 11/73, p. 50.

Chapter 5

Fig. 5-1. Hands-On Electronics, Summer 1984, p. 77.
Fig. 5-2. Courtesy, William Sheets.
Fig. 5-3. Courtesy, William Sheets.
Fig. 5-4. Electronic Engineering, 5/84, p. 44
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Fig. 5-6. Electronics Today International, 4/85, p. 82.

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Fig. 7-1. Motorola, TMOS Power FET Design Ideas, 1985.
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Fig. 7-6. Radio-Electronics, 6/85, p. 60.
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Fig. 7-10. Texas Instruments, Linear and Interface Circuits Applications, 1985, vol. 1, p. 3-5.
Fig. 7-11. Radio-Electronics, 1979.
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Fig. 7-15. Radio-Electronics, 5/87, p. 10.
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Fig. 7-17. Hands-On Electronics, 4/87, p. 92.
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Chapter 8

Fig. 8-1. CQ, 7/82, p. 18.
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Fig. 8-6. Courtesy, William Sheets.
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